8

Protecting Diodes, Transistors, and Thyristors

All power switching devices attain better switching performance if some form of switching aid circuit, called snubber, is employed. Snubber activation may be either passive or active which involves extra power switches. Only passive snubbers, which are based on passive electrical components, are considered in this chapter, while active snubbers are considered in Chapter 9. Fundamentally, the MOSFET and IGBT do not require switching aid circuits, but circuit imperfections, such as stray inductance and diode recovery, can necessitate the need for some form of switch snubber protection.
Protection in the form of switching aid circuits performs two main functions:

- Divert switching losses from the switch thereby allowing a lower operating temperature, or higher electrical operating conditions for a given junction temperature.
- Prevent transient electrical stressing that may exceed *I-V* ratings thereby causing device failure.

Every switching device can benefit from switching protection circuits, but extra circuit component costs and physical constraints may dictate otherwise.
The diode suffers from reverse recovery current and voltage snap which induces high but short duration circuit voltages. These voltage transients may cause interference to the associated circuit and to nearby equipment. A simple series *R-C* circuit connected in parallel to the stressed device is often used to help suppress the voltage oscillation at diode turn-off. Such a suppression circuit can be used on simple mains rectifying circuits when rectification causes conducted and radiated interference.
Although the MOSFET and IGBT can usually be reliably and safely operated without external protection circuitry, stringent application emission restrictions may dictate the use of snubbers. In specific applications, the IGBT is extensively current derated as its operating frequency increases. In order to attain better device current utilization, at higher frequencies, various forms of switching aid circuits can be used to divert switching losses from the stressed switch.
Generally, all thyristor devices benefit from a turn-on switching aid circuit, which is based on a series connected inductor that is active at thyristor turn-on. Such an

inductive turn-on snubber is obligatory for the high-power GTO thyristor and GCT. In order to fully utilise the GTO thyristor, it is usually used in conjunction with a parallel-connected capacitive turn-off snubber, which decreases device stressing during the turn-off transient. Triacs and rectifying grade SCRs and diodes tend to use a simple *R-C* snubber connected in parallel to the switch to reduce interference. The design procedure of the *R-C* snubber for a diode is different to that for the *R-C* snubber design for a thyristor device, because the protection objectives and initial conditions are different. In the case of a thyristor or rectifying diode, the objective is to control both the voltage rise at turn-off and recovery overshoot effects. For the fast recovery diode or any high-speed switch, the principal objectives are to control the voltage overshoot magnitude at diode snap recovery or at turn-off respectively, which are both exacerbated because of stray circuit inductance carrying current.

8.1 The *R-C* snubber

The series *R-C* snubber is the simplest switching aid circuit and is connected in parallel to the device being aided. It is characterized by having low series inductance and a high transient current rating. These requirements necessitate carbon type resistors for low inductance, below a few watts, and metal film resistors at higher powers. The high current and low inductance requirements are also provided by using metallised, polypropylene capacitors with high *dv/dt* ratings of typically hundreds of V/μs.
Theoretically a purely capacitive snubber would achieve the required protection objectives, but series resistance is added to decrease the current magnitude when the capacitor is discharging and to damp any oscillation by dissipating the oscillatory energy generated at turn-off when an over-voltage tends to occur.

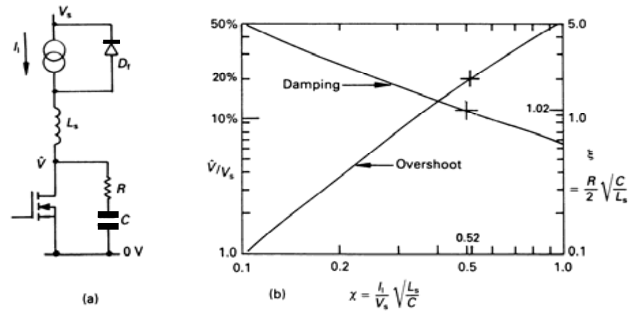

Figure 8.1. *MOSFET drain to source R-C snubber protection:*
(a) MOSFET circuit showing stray inductance, L_s, and R-C protection circuit and
(b) R-C snubber optimal design curves.

8.1.1 R-C switching aid circuit for the MOSFET and the diode

In figure 8.1a, at switch turn-off stray inductance L_s unclamped by the load freewheel diode, D_f, produces an over voltage $\hat{V}$ on the MOSFET or IGBT. The energy associated with the inductor can be absorbed in the shown drain to source *R-C* circuit, thereby containing the voltage overshoot to a safe level. Such an *R-C* snubber circuit is used extensively in thyristor circuits, 8.1.2, for *dv/dt* protection, but in such cases the initial current in the stray inductance is assumed zero. Here the initial inductor current is equal to the maximum load current magnitude, I_ℓ. The design curves in figure 8.1b allow selection of *R* and *C* values based on the maximum voltage overshoot $\hat{V}$ and an initial current factor χ, defined in figure 8.1b. The *C* and *R* values are given by

$$C = L_s\,(I_\ell / \chi V_s)^2 \qquad \text{(F)} \qquad (8.1)$$

$$R = 2\xi\, V_s \chi / I_\ell \qquad (\Omega) \qquad (8.2)$$

If the *R-C* circuit time constant, $\tau = RC$, is significantly less than the MOSFET voltage rise and fall times, t_{rv} and t_{fv}, at reset, a portion of the capacitor energy $\tfrac{1}{2}CV_s^2$, is dissipated in the switch, as well as in *R*. The switch appears as a variable resistor in series with the *R-C* snubber. Under these conditions (t_{fv} and $t_{rv} > RC$) the resistor power loss is approximately by

$$P_R = P_{Ron} + P_{Roff}$$

$$= \frac{\tau}{\tau + t_{fv}} P_{C0} + \frac{\tau}{\tau + t_{rv}} (P_{C0} + P_{L0}) \qquad \text{(W)} \qquad (8.3)$$

where $P_{C0} = \tfrac{1}{2}CV_s^2 f_s$ and $P_{L0} = \tfrac{1}{2}L_s I_\ell^2 f_s$

otherwise (t_{fv} and $t_{rv} < RC$) the resistor losses are the energy to charge and discharge the snubber capacitor, plus the energy stored in the stray inductance, that is $2P_{C0}+P_{L0}$.
Note the total losses are independent of snubber resistance. The snubber resistor determines the time over which the energy is dissipated.
When the *R-C* snubber is employed across a fast recovery diode, the peak reverse recovery current is used for I_ℓ in the design procedure.

Example 8.1: ***R-C snubber design for MOSFETs***

A MOSFET switches a 40 A inductive load on a 200 V dc rail, at 10 kHz. The unclamped drain circuit inductance is 20 nH and the MOSFET voltage rise and fall times are both 100 ns. Design a suitable *R-C* snubber if the MOSFET voltage overshoot is to be restricted to 240 V (that is, 40V overshoot, viz. 20%).

Solution
From figure 8.1b, for 20 per cent voltage overshoot

$$\xi = 1.02, \quad \chi = 0.52$$

Using equations (8.1) and (8.2) for evaluating *C* and *R* respectively,

$$C = L_s\,(I_\ell / \chi V_s)^2 = 20\text{nH}\left(\frac{40\text{A}}{0.52\times 200\text{V}}\right)^2 = 3\text{nF}$$

$$R = 2\xi V_s \chi / I_\ell = 2\times 1.02\times \frac{0.52\times 200\text{V}}{40\text{A}} = 5.3\Omega$$

Use *C*=3.3 nF, 450V dc, metallised polypropylene capacitor and *R*=5.6 Ω.
Since the *RC* time constant, 18.5ns, is short compared with the MOSFET voltage transient times, 100ns, the resistor power rating is given by equation (8.3).

$$P_{C0} = \tfrac{1}{2}CV_s^2 f_s = \tfrac{1}{2}\times 3.3\text{nF}\times 200^2\times 10\text{kHz} = 2.64\text{W}$$

$$P_{L0} = \tfrac{1}{2}L_s I_\ell^2 f_s = \tfrac{1}{2}\times 20\text{nH}\times 40^2\times 10\text{kHz} = 0.16\text{W}$$

$$P_R = \frac{18.5}{100+18.5}\times 2.64\text{W} + \frac{18.5}{100+18.5}\times(2.64\text{W} + 0.16\text{W}) = 0.85\text{W}$$

Use a 5.6 Ω, 1 W carbon composition resistor for low self inductance, with a working voltage of at least 250V dc. Parallel connection of two ½W, carbon composition resistors may be necessary since resistance values below 10Ω are uncommon.
The MOSFET switching losses are increased by $2W_{C0} + P_{L0} - 0.85\text{W} = 4.95\text{W}$.

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8.1.2 R-C snubber circuit for a converter grade thyristor and a triac

The snubber circuit for a low switching frequency thyristor is an anode-to-cathode parallel connected *R-C* series circuit for off-state voltage transient suppression. Series inductance may be necessary (forming a turn-on snubber) to control anode *di/dt* at turn-on. This inductive snubber is essential for the GTO thyristor and the GCT, and will be considered in section 8.3.3.

Off-state dv/dt suppression snubber

Thyristors, other than the GTO thyristor, normally employ a simple *R-C* snubber circuit as shown in figure 8.2. The purpose of the *R-C* snubber circuit is not primarily to reduce turn-off switching loss but rather to prevent false triggering from applied or reapplied anode *dv/dt*, when the switch is in the off-state.
Any thyristor rate of rise of forward-voltage anode *dv/dt* produces a central junction charging current which may cause the thyristor to inadvertently turn on. The critical *dv/dt* is defined as the minimum value of *dv/dt* which will cause switching from the off-state to the on-state. In applications as shown in figure 8.2, an occasional false turn-on is generally not harmful to the triac or the load, since the device and the load only have to survive the surge associated with a half-a-cycle of the ac mains supply.
In other applications, such as reversible converters, a false *dv/dt* turn-on may prove catastrophic. A correctly designed snubber circuit is therefore essential to control the rate of rise of anode voltage.

The action of this *R-C* snubber circuit relies on the presence of inductance in the main current path. Zero inductor current is the initial condition, since the device is in the off-state when experiencing the anode positive *dv/dt*. The inductance may be stray, from transformer leakage or a supply, or deliberately introduced. Analysis is based on the response of the *R-C* portion of an *L-C-R* circuit with a step input voltage and zero initial inductor current. Figure 8.3 shows an *L-C-R* circuit with a step input voltage and the typical resultant voltage across the SCR or *R-C* components. The circuit resistor *R* damps (by dissipating power) any oscillation and limits the capacitor discharge current through the SCR at subsequent device turn-on from the gate.

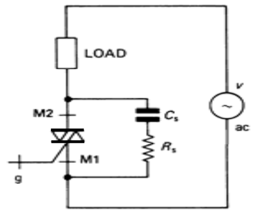

Figure 8.2. *Thyristor (triac) ac circuit with an R-C snubber circuit.*

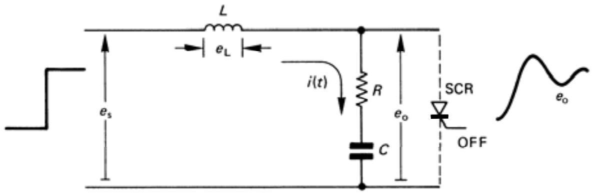

Figure 8.3. *R-C snubber equivalent circuit showing the second-order output response e_o to a step input voltage e_s.*

Based on the snubber circuit analysis presented in the appendix in section 8.5 at the end of this chapter, the maximum *dv/dt*, $\hat{S}$, which is usually specified for a given device, seen by the SCR for a step input of magnitude e_s, is given by

$$\hat{S} = e_s R / L \qquad \text{(V/s)} \qquad (8.4)$$

for a damping factor of $\xi > ½$. That is, after rearranging, the snubber resistance is given by

$$R = L\hat{S}/e_s \qquad \text{(ohms)} \qquad (8.5)$$

while the snubber capacitance is given by

$$C = \frac{4\xi^2 e_s}{R\hat{S}} \qquad \text{(F)} \qquad (8.6)$$

and the peak snubber current is approximated by

$$\hat{I} = \frac{e_s}{R}\frac{2\xi}{\sqrt{1-\xi^2}} \qquad \text{(A)} \qquad \text{for } \xi < 1. \qquad (8.7)$$

Figure 8.4 shows the variation of the various normalised design factors, with damping factor ξ.

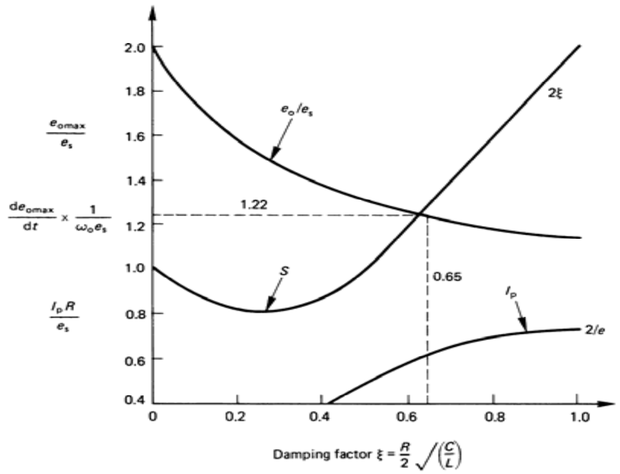

Figure 8.4. *Variation of snubber peak voltage, e_o, maximum de_o/dt, $\hat{S}$; and peak current, I_p; with L-C-R damping factor ξ.*

Example 8.2: ***R-C snubber design for a converter grade thyristor***

Design an *R-C* snubber for the SCRs in a cycloconverter circuit where the SCRs experience an induced *dv/dt* due to a complementary SCR turning on, given

- peak switching voltage, $e_s = 200$ V
- operating frequency, $f_s = 1$ kHz
- *dv/dt* limit, $S = 200$ V/μs.

Assume

- stray circuit $L = 10$ μH
- 22 per cent voltage overshoot across the SCR
- an *L-C-R* snubber is appropriate.

Solution

From equation (8.5) the snubber resistance is given by

$$R = L\hat{S}/e_s$$

$$= \frac{10\mu\text{H}\times 200\text{V}/\mu\text{s}}{200\text{V}} = 10\Omega$$

At turn-on the additional anode current from the snubber capacitor will be 200V/10Ω = 20A, which decays exponentially to zero, with a 1.8μs (10Ω×180nF) *RC* time constant. Figure 8.4 shows the *R-C* snubber circuit overshoot magnitude, $\hat{e}_0/e_s$ for a range of damping factors ξ. The normal range of damping factors is between 0.5 and 1. Thus from figure 8.4, allowing 22 per cent overshoot, implies $\xi = 0.65$. From equation (8.6)

$$C = \frac{4\xi^2 e_s}{R\hat{S}} = \frac{4\times(0.65)^2\times 200\text{V}}{10\Omega\times 200\times 10^6}$$

$$= 180 \text{ nF (preferred value) rated at 244 V peak.}$$

From equation (8.7) the peak snubber current during the applied *dv/dt* is

$$\hat{I} = \frac{e_s}{R}\frac{2\xi}{\sqrt{1-\xi^2}}$$

$$= \frac{200\text{V}}{10\Omega}\frac{2\times 0.65}{\sqrt{1\text{-}0.65^2}} = 34 \text{ A}$$

The 10 ohm snubber resistor losses are given by

$$P_{10\Omega} = Ce_0^2 f_s$$

$$= 180\times10^{-9}\times244^2\times1\times10^3 = 11\text{W}$$

The resistor carries current to both charge (maximum 34A) and discharge (initially 20A) the capacitor. The necessary 10Ω, 11W resistor must have lower inductance, hence two 22 Ω, 7W, 500V dc working voltage, metal oxide film resistors can be parallel connected to achieve the necessary ratings.

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Variations of the basic snubber circuit are shown in figure 8.5. These circuits use extra components in an attempt to control SCR initial *di/dt* arising from snubber discharge through R_L at turn-on.

An *R-C* snubber can be used across a diode in order to control voltage overshoot at diode snap-off during reverse recovery, as a result of stray circuit inductance, as considered in 8.1.1.

The *R-C* snubber can provide decoupling and transient overvoltage protection on both ac and dc supply rails, although other forms of *R-C* snubber circuit may be more applicable, specifically the soft voltage clamp.

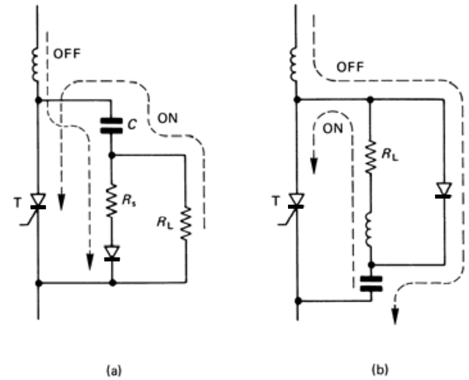

Figure 8.5. *Variations of the basic thyristor R-C snubber:*
(a) R_s<<R_L and (b) transistor-type R-C-D snubber, R_s = 0.

8.2 The soft voltage clamp

A primary function of the basic *R-C* snubber is to suppress voltage overshoot levels. The *R-C* snubber commences its clamping action from zero volts even though the objective is to clamp the switch voltage to the supply voltage level, V_s. Any clamping action below V_s involves the unnecessary transfer of energy. The soft voltage clamp reduces energy involvement since it commences clamping action once the switch voltage has reached the supply voltage V_s, and the voltage overshoot commences.

The basic *R-C-D* soft voltage clamp is shown in figure 8.6a, with resistor *R* parasitic inductance, L_R, and stray or deliberately introduced unclamped inductance *L*, shown.
The voltage clamp functions at switch turn-off once the switch voltage exceeds V_s. The capacitor voltage does not fall below the supply rail voltage V_s. Due to the stored energy in *L*, the capacitor *C* charges above the rail voltage and *R* limits current magnitudes as the excess capacitor charge discharges through *R* in to V_s. All the energy stored in *L*, $\frac{1}{2}LI_m^2$, is dissipated in *R*. The inductor current i_L and capacitor voltage V_c waveforms are shown in figure 8.6b.
At switch turn-on, the diode D blocks, preventing discharge of C which remains charged to V_s.

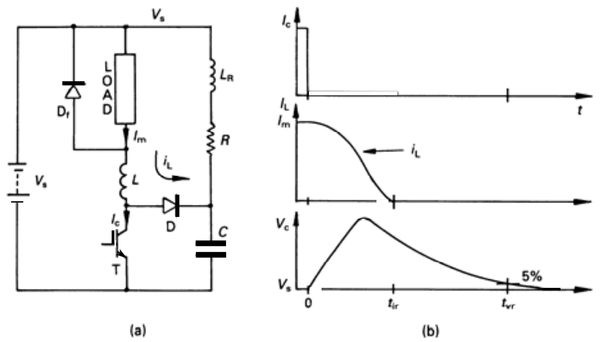

Figure 8.6. *Soft voltage clamp:*
(a) circuit diagram and (b) turn-on inductor current, I_L, and capacitor voltage, V_c, at switch turn-off.

The energy drawn from the supply V_s as the capacitor overcharges, is returned to the supply as the capacitor discharges through *R* into the supply. The net effect is that only the energy in *L*, $\frac{1}{2}LI_m^2$, is dissipated in *R*.
Analysis is simplified if the resistor inductance L_R is assumed zero. The inductor current decreases from I_m to 0 according to

$$i_L(\omega t) = I_m \frac{\omega_0}{\omega} e^{-\alpha t}\cos(\omega t - \phi) \qquad \text{(A)} \qquad (8.8)$$

where

$$\alpha = \tfrac{1}{2}RC \quad \text{(s)} \qquad \omega_0 = 1/\sqrt{LC} \quad \text{(rad/s)}$$

$$\omega = \sqrt{\omega_0^2 - \alpha^2} \quad \text{(rad/s)} \qquad \phi = \tan^{-1}\frac{\alpha}{\omega} \quad \text{(rad)}$$

The inductor current reaches zero, termed the current reset time, t_{ir}, in time

$$t_{ir} = (\tfrac{1}{2}\pi + \phi)/\omega \qquad \text{(s)} \qquad (8.9)$$

which must be shorter than the switch minimum off-time, $\check{t}_{off}$. The capacitor charges from V_s according to

$$V_C(\omega t) = V_s + \frac{I_m}{\omega C} e^{-\alpha t}\sin\omega t \qquad \text{(V)} \qquad (8.10)$$

The maximum capacitor voltage, hence maximum switch voltage, occurs for large *R*

$$\hat{V}_C = V_s + I_m\sqrt{\frac{L}{C}} \qquad \text{(V)} \qquad (8.11)$$

Once the current in *L* has reduced to zero the capacitor discharges to V_s exponentially, with a time constant *RC*.
The practical *R-C* circuit, which includes the stray inductance L_R, must be over-damped, that is

$$R > 2\sqrt{\frac{L_R}{C}} \qquad (\Omega) \qquad (8.12)$$

The capacitor voltage reset time t_{vr} is the time for the capacitor to discharge to within 5 per cent of V_s, as shown in figure 8.6b.
The stray inductance L_R increases the peak capacitor voltage and increases the voltage reset time. Design of the voltage clamp, including the effects of L_R, is possible with the aid of figure 8.7. Design is based on specifying the maximum voltage overshoot, V_{cp} and minimizing the voltage reset time, t_{vr}, which limits the upper switching frequency, f_s, where $f_s \le 1/t_{vr}$ such that $\check{t}_{off} \ge t_{ir}$.

Example 8.3: ***Soft voltage clamp design***
A 5 μH inductor turn-on snubber is used to control diode reverse recovery current and switch turn-on loss, as shown in figure 8.6a. The maximum collector current is 25 A, while the minimum off-time is 5 μs and the maximum operating frequency is 50 kHz.

i. Assuming an independent *L-C* resonant transfer from *L* to *C* and a subsequent *R-C* discharge cycle, calculate soft voltage clamp *R* and *C* requirements.
ii. Use figure 8.7 to determine the voltage clamp requirements if the discharge (reset) resistor inductance L_R is
   (a) 0
   (b) 1.0μH.

In each case, the maximum switch overshoot is to be restricted to 50 V.

Solution
i. Assuming all the inductor energy is transferred to the clamp capacitor, before any discharge through R occurs, then from equation (8.11), for a 50 V capacitor voltage rise

$$50 = I_m\sqrt{L/C}$$

that is, C = 5 μH/(50V/25A)^2 = 1.25 μF (use 1.2 μF).

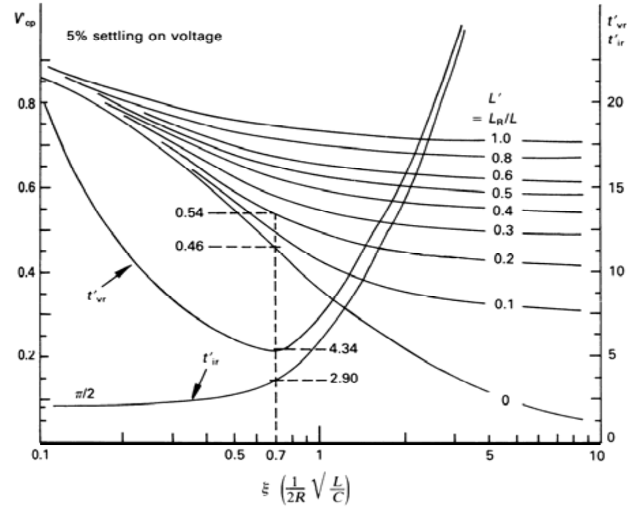

Figure 8.7. *Voltage clamp capacitor normalised peak over-voltage, V'_{Cp}, versus damping factor, ξ, for different resistor normalised inductances, L′, and voltage and current normalised settling times,* t'_{vr}, $t'_{ir} = V_{cp}/\{I_m\sqrt{L/C}\}$, $t'_{vr} = t_{rv}/\omega_0$, $t'_{ir} = t_{ir}/\omega_0$.

From equation (8.9), for $R = 0$, the energy transfer time (from L to C) is

$$t_{ir} = \tfrac{1}{2}\pi\sqrt{LC} = \tfrac{1}{2}\pi\sqrt{5\mu\text{H}\times 1.25\mu\text{F}} = 4\mu\text{s}$$

which is less than the switch minimum off-time of 5 µs.
If the maximum operating frequency is 50 kHz, the capacitor must discharge in 20 - 4 = 16 µs. Assuming five RC time constants for capacitor discharge

$$5\times RC = 16\mu\text{s}$$

$$R = 16\mu\text{s}/(5\times 1.2\mu\text{F}) = 2\tfrac{2}{3}\Omega \quad (\text{use } 2.4\Omega)$$

The resistor power rating is

$$P_R = \tfrac{1}{2}LI_m^2 f_s = \tfrac{1}{2}\times 5\mu\text{H}\times 25^2\times 50\text{kHz} = 78\text{W}$$

Obviously with a 2.4 Ω discharge resistor and 50V overshoot, discharge current would flow as the capacitor charges above the voltage rail. A smaller value of C could be used. A more accurate estimate of C and R values is possible, as follows.

ii. (a) $L_R = 0$, $L' = 0$
From figure 8.7, for the minimum voltage reset time

$$V'_{cp} = 0.46, \quad t'_{ir} = 2.90, \quad t'_{rv}\,4.34, \quad \text{and} \quad \xi = 0.70$$

From $V'_{cp} = V_{cp}/I_m\sqrt{L/C}$

$$0.46 = 50\text{V}/25\text{A}\sqrt{5\mu\text{H}/C} \quad \text{gives } C = 0.27\mu\text{F}$$

From $\xi = \tfrac{1}{2R}\sqrt{L/C}$, $R = \tfrac{1}{2\xi}\sqrt{L/C} = \dfrac{1}{2\times 0.7}\sqrt{5\mu\text{H}/0.27\mu\text{F}} = 3.2\Omega$

(Use 3.3 Ω, 78 W)
The reset times are given by

$$t_{vr} = t'_{vr}\sqrt{LC} = 4.34\times 1.16 = 5\mu\text{s} \quad (<20\mu\text{s})$$

$$t_{ir} = t'_{ir}\sqrt{LC} = 2.9\times 1.16 = 3.4\mu\text{s} \quad (<5\mu\text{s})$$

It is seen that smaller capacitance (1.2 µF *vs* 0.27 µF) can be employed if simultaneous *L-C* transfer and *R-C* discharge are accounted for. The stray inductance of the resistor discharge path has been neglected. Any inductance decreases the effectiveness of the *R-C* discharge. Larger C than 0.27 µF and R < 3.3Ω are needed, as is now shown.
ii. (b) $L_R = 1\mu\text{H}$, $L' = L_R/L = 0.2$
In figure 8.7, for a minimum voltage reset time, ξ = 0.7, V'_{cp} = 0.54 when the L'=0.2 curve is used. The normalised reset times are unchanged, that is t'_{ir} = 2.9 and t'_{vr} = 4.34. Using the same procedure as in part b(i)

$$0.54 = 50\text{V}/25\text{A}\sqrt{5\mu\text{H}/C} \quad \text{gives } C = 0.37\mu\text{F} \quad (\text{use } 0.39\mu\text{F})$$

$$R = \tfrac{1}{2\xi}\sqrt{L/C} = \frac{1}{2\times 0.7}\sqrt{5\mu\text{H}/0.39\mu\text{F}} = 2.6\Omega \quad (\text{use } 2.7\Omega, 78\text{W})$$

Since resistor inductance has been accounted for, parallel connection of four 10Ω, 25W wire-wound aluminium clad resistors can be used.

$$t_{vr} = 4.34\times 1.4 = 6\mu\text{s} \quad (< 20\mu\text{s})$$

$$t_{ir} = 2.90\times 1.4 = 4\mu\text{s} \quad (< 5\mu\text{s})$$

Note that circuit supply voltage V_s is not a necessary design parameter, other than to specify the capacitor absolute dc voltage rating.

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8.3 Switching-aid circuits

Optimal gate drive electrical conditions minimize collector (or drain or anode) switching times, thus minimizing switch electrical stresses and power losses. Proper gate drive techniques greatly enhance the switching robustness and reliability of a power switching device. Switching-aid circuits, commonly called *snubber circuits,* can be employed to further reduce device switching stresses and losses. Optimal gate drive conditions minimise the amount of snubbering needed.

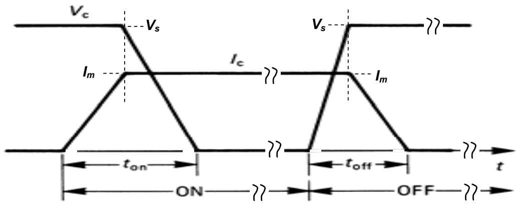


Figure 8.8. *Idealised collector switching waveforms for an inductive load.*

During both the switch-on and the switch-off intervals, for an inductive load as considered in chapter 6.2, an instant exists when the switch simultaneously supports the supply voltage V_s and conducts the full load current I_m, as shown in figure 8.8. The gate drive conditions cannot alter this peak power loss but can vary the duration of the switching periods (t_{on} and t_{off}). From chapter 6, the switching losses, W, dissipated as heat in the switch, are given by

for turn-on $W_{on} = \frac{1}{2}V_s I_m t_{on}$ (J) (8.13)

for turn-off $W_{off} = \frac{1}{2}V_s I_m t_{off}$ (J) (8.14)

In order to reduce switching losses, two snubber circuits can be employed on a power switching device, one operational during switch turn-on, the other effective during turn-off. In the case of the turn-off snubber, energy is diverted from the switch turning off into a parallel capacitor as shown in figure 8.9a. The turn-on snubber utilises an inductor in series with the collector as shown in figure 8.9b in order to control the rate of rise of anode current during the collector voltage fall time. For both snubbers, the *I-V* SOA trajectory is modified to be within that area shown in figure 6.8.

- *An* ***inductive turn-on snubber*** is essential for the GTO thyristor and the GCT in order to control the initial *di/dt* current to safe levels at switch turn-on. In large area thyristor devices, the inductor controlled current increase at turn-on allows sufficient time for the silicon active area to spread uniformly so as to conduct safely the prospective load current. Special thyristor gate structures such as the amplifying gate, as shown in figure 3.23, allow initial anode *di/dt* values of up to 1000 A/us. Use of a turn-on snubber with the MOSFET and the IGBT is limited but may be used because of freewheel diode imposed limitations rather than an intrinsic need by the switch.
- *The* ***capacitive turn-off snubber*** is used extensively on the GTO thyristor. The *R-D-C* circuit is necessary to ensure that GTO turn-off occurs at a low anode-to-cathode voltage, preventing excessive power loss at the central GTO junction during reverse recovery. Larger area GTOs employ 1 to 8 μF in an *R-D-C* turn-off snubber and at high voltages and frequencies the associated losses, $\frac{1}{2}C_s V_s^2 f_s$, tend to be high. To reduce this loss, GTOs with an increased SOA, namely GCTs, for use without a turn-off snubber are available. These devices under utilise their voltage and current density capabilities as compared with when a turn-off snubber is used.

While the switching performance of IGBTs and MOSFETs can be enhanced by using the turn-off snubber, it is not a prerequisite for safe, reliable switch operation.

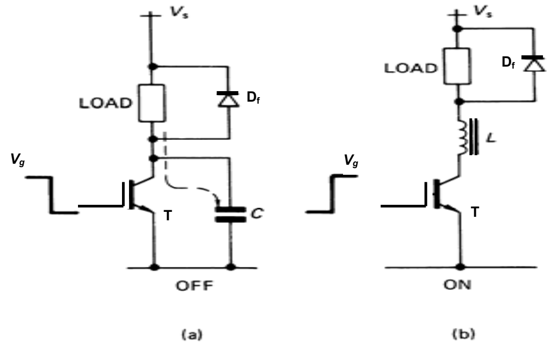


Figure 8.9. *Basic switching-aid circuits comprising:*
(a) a capacitor for current shunting at switch turn-off and
(b) an inductor for limiting the rate of rise of principal current at turn-on.

8.3.1 *The turn-off snubber circuit - assuming a linear current fall*

Figure 8.10 shows a complete turn-off snubber circuit comprising a capacitor-diode plus resistor combination across the anode-to-cathode/collector-to-emitter terminals of the switching device. At switch turn-off, load current is diverted into the snubber capacitor *C* via the diode D, while the switch principal current decreases. The anode/collector voltage is clamped to the capacitor voltage, which is initially zero. The larger the capacitor, the slower the anode/collector voltage rises for a given load current and, most importantly, turn-off occurs without a condition of simultaneous supply voltage and maximum load current (V_s,I_m). Figure 8.11 shows the anode/collector turn-off waveforms for different magnitudes of snubber capacitance. The GTO/IGBT tail current has been neglected, thus the switching device is analysed without any tail current. For clarity the terminology to be henceforth used, refers to an IGBT, viz., collector, emitter, and gate.

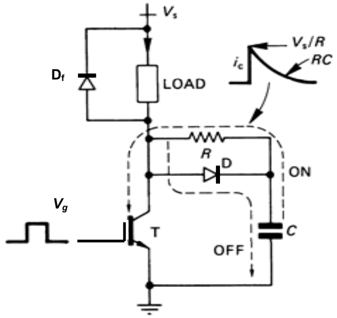


Figure 8.10. *Practical capacitive turn-off snubber showing capacitor charging and discharging paths during device switching.*

Figure 8.11a shows turn-off waveforms for a switch without a snubber, where it has been assumed that the collector voltage rise time is short compared with the collector current fall time, which is given by $i_c(t) = I_m(1 - t/t_{fi})$. For low capacitance values, the snubber capacitor (whence collector) may charge to the rail voltage before the collector current has fallen to zero, as seen in figure 8.11b. For larger capacitance, the collector current reaches zero before the capacitor (whence collector) has charged to the rail voltage level, as shown in figure 8.11c.
For analysis, the collector voltage rise time for an unaided switch is assumed zero. The device switch-off energy losses without a snubber, as shown in figure 8.11a, are given by

$$W = \tfrac{1}{2} V_s I_m t_{fi} \qquad \text{(J)} \qquad (8.15)$$

With a snubber circuit, switch losses are decreased as shown in figure 8.11d, but snubber (resistor) losses are incurred. After turn-off the capacitor is charged to the rail voltage. This stored energy, $\tfrac{1}{2} C_s V_s^2$, is subsequently dissipated as heat in the snubber circuit resistor at subsequent switch turn-on, when an *R-C* discharge occurs. If the snubber *RC* time constant is significantly shorter than the switch voltage fall time at turn-on, the capacitor energy dissipated in the resistor is less than $\tfrac{1}{2} C_s V_s^2$ and switch losses are increased as considered in 8.1.1. A range of capacitance values exists where the total losses - snubber plus switch - are less than those losses incurred if the same device is switched unaided, when losses as given by equation (8.14) result. Two distinct snubber design cases exist, as indicated by figures 8.11b and 8.11c. The two possibilities and the associated circuit voltage and current waveforms in each case are shown in detail in figure 8.12. The waveforms are based on satisfying Kirchhoff's

voltage and current laws for each case.
If the snubber capacitor charges fully before the collector current has reached zero then the switch losses are given by

$$W_t = \tfrac{1}{2} V_s I_m t_{fi} \left(1 - \tfrac{2}{3} k + \tfrac{1}{2} k^2\right) \qquad \text{(J)} \qquad (8.16)$$

for $k \le 1$, where $k = \tau / t_{fi}$, as defined in figures 8.11b and 8.13.
Alternatively, with larger capacitance, if the snubber capacitor does not charge fully to V_s until after the collector current reaches zero, that is $k \ge 1$, then the switch losses are given by

$$W_t = \tfrac{1}{2} V_s I_m t_{fi} \; \frac{1}{6(2k-1)} \qquad \text{(J)} \qquad (8.17)$$

for $k \ge 1$ as defined in figures 8.11c and 8.13. Initially the capacitor voltage increase is quadratic, then when the collector current reaches the load current level, the capacitor voltage increase becomes linear.

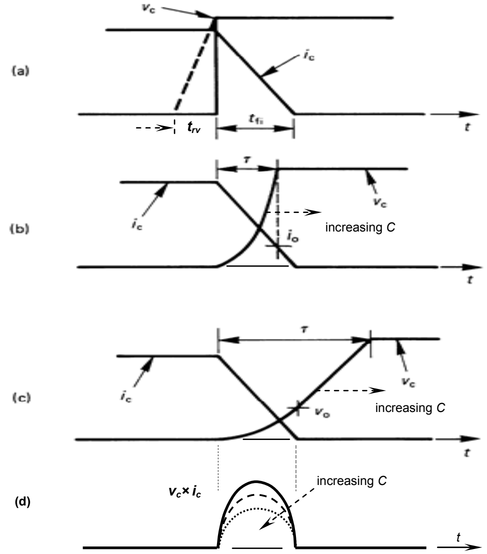


Figure 8.11. *Switch turn-off waveforms: (a) unaided turn-off; (b) turn off with small snubber capacitance; (c) turn-off with large snubber capacitance; (d) and reset power losses.*

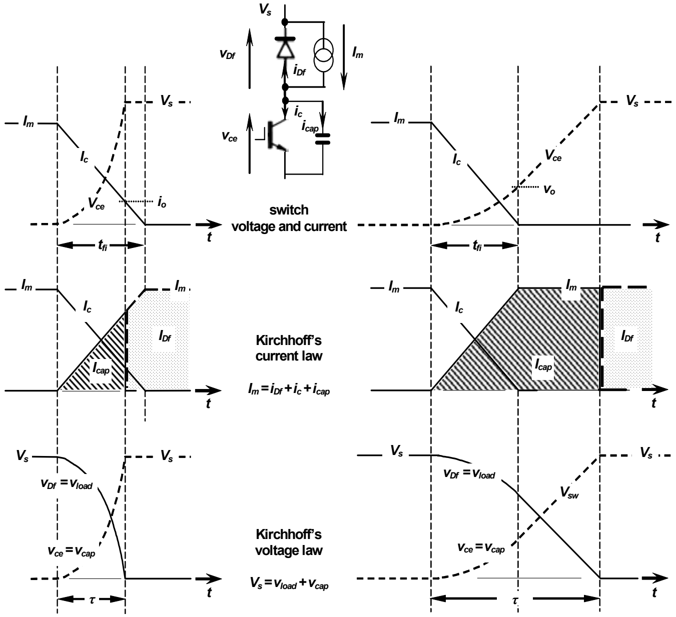


Figure 8.12. *Switch turn-off waveforms satisfying Kirchhoff's laws: (a) turn-off with small snubber capacitance and (b) turn-off with large snubber capacitance.*

These losses, normalised with respect to the unaided switch losses given by equation (8.15), are plotted in figure 8.13. The switch and capacitor components contributing to the total losses are also shown. A number of important points arise concerning turn-off snubbers and snubber losses.

(a) Because of current tailing, voltage overshoot, and the assumption that the voltage rise time t_{rv} is insignificantly short, practical unaided switch losses, equation (8.14), are approximately twice those indicated by equation (8.15).

(b) As the snubber capacitance increases, that is, k increases, the switch loss is progressively reduced but at the expense of increased snubber associated loss.

(c) If $k \leq 1.41$ the total losses are less than those for an unaided switch. In the practical case $k \leq 2.70$ would yield the same condition.

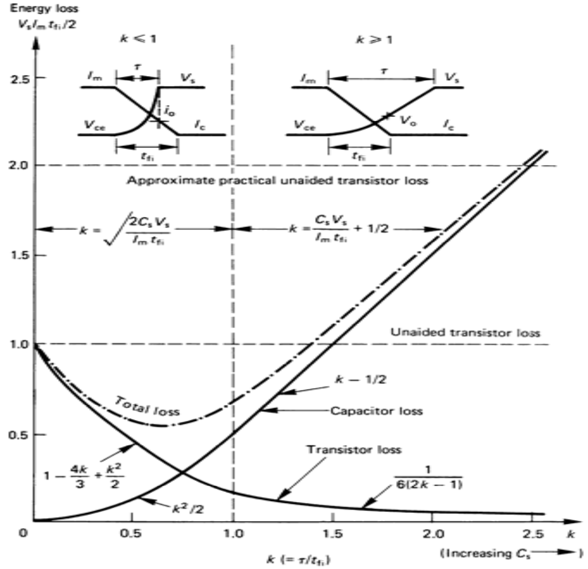


Figure 8.13. *Loss components for a switch at turn-off when employing a capacitance-type snubber and assuming the collector current falls according to* $i_c = I_m(1-t/t_{fi})$.

(d) A minimum total loss (switch plus capacitor) condition exists. When $k = 2/3$ the total losses are only $5/9$ those of an unaided switch. The snubber capacitance for this condition is given by

$$C_s = \frac{2}{9}\frac{I_m t_{fi}}{V_s} \qquad \text{(F)} \qquad (8.18)$$

(e) Losses are usually minimised at the maximum loss condition, that is maximum load current I_m. At lower currents, the capacitor charging time is increased.

(f) Snubbers not only reduce total losses, but because the loss is distributed between the switch and resistor, more effective heat dispersion can be achieved.

(g) High switch current occurs at turn-on and incorporates the load current I_m, the snubber capacitor exponential discharge $\frac{V_s}{R}\left(1-e^{-t/CR}\right)$, and any freewheel diode reverse recovery current.

The capacitor energy $\tfrac{1}{2}C_sV_s^2$ is removed at turn-on and is exponentially dissipated mainly in the snubber circuit resistor. The power rating of this resistor is dependent on the maximum switching frequency and is given by

$$P_{R_s} = \tfrac{1}{2}C_sV_s^2 f_s \qquad \text{(W)} \qquad (8.19)$$

Two factors specify the snubber circuit resistance value.

- The snubber circuit *RC* time constant period must ensure that after turn-on the capacitor discharges before the next turn-off is required. If $\check{t}_{on}$ is the minimum switch on-time, then $\check{t}_{on} = 5R_sC_s$, is sufficient to ensure the correct snubber circuit initial conditions, namely, zero capacitor volts.
- The resistor initial current at capacitor discharge is V_s / R_s. This component is added to the load current at switch turn-on, hence adding to the turn-on stresses. The maximum collector current rating must not be exceeded. In order to reduce the initial discharge current, a low valued inductor can be added in series with the resistor, (or a wire-wound resistor used), thus producing an overdamped *L-C-R* discharge current oscillation at turn-on.

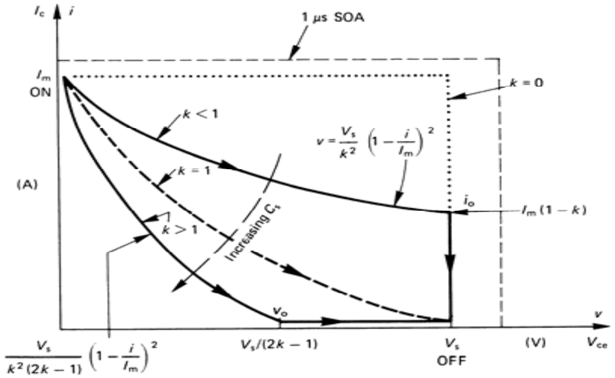


Figure 8.14. *The collector I-V trajectory at turn-off with a switching-aid circuit.*

As a result of utilising a turn-off snubber the collector trajectory across the SOA is modified as shown in figure 8.14. It is seen that the undesired unaided condition of simultaneous supply voltage V_s and load current I_m is avoided. Typical trajectory

conditions for a turn-off snubbered device are shown for three situations, depending on the relative magnitudes of t_{fi} and τ. A brief mathematical derivation describing the turn-off switching-aid circuit action is presented in the appendix in section 8.6 at the end of this chapter.

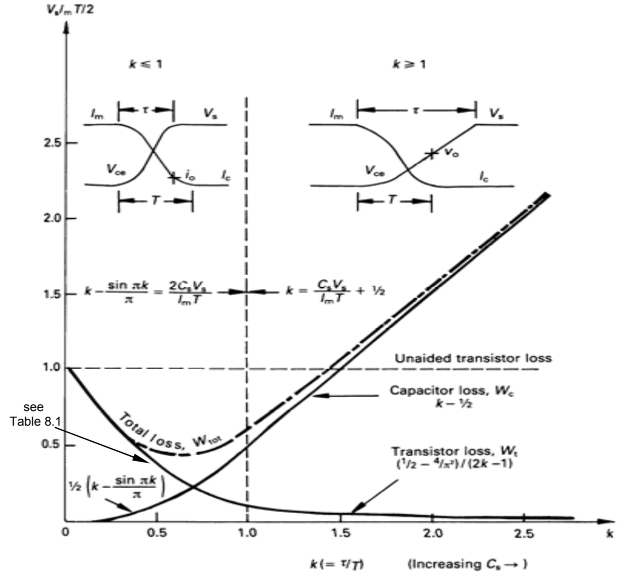


Figure 8.15. *Loss components for a switch at switch-off when employing a capacitance-type snubber and assuming a collector fall current according to* $i_c = \tfrac{1}{2}I_m\{1 + \cos(\pi t/T)\}$.

Table 8.1. Normalised switching loss components at turn-off with a cosinusoidal current fall of half period T

$\times V_s I_m T/2$ loss	$k = \tau/T \leqslant 1$	$k = \tau/T \geqslant 1$
Switch W_t	$\dfrac{\dfrac{k^2}{2} + \dfrac{2}{\pi^2}(\cos \pi k - 1) + \dfrac{k}{\pi}\sin \pi k - \dfrac{\sin^2 \pi k}{2\pi^2}}{\left(k - \dfrac{\sin \pi k}{\pi}\right)}$ $+ 1 - k - \dfrac{\sin \pi k}{\pi}$	$\dfrac{\dfrac{1}{2} - \dfrac{4}{\pi^2}}{2k - 1}$
Resistor W_c	$\left(k - \dfrac{\sin \pi k}{\pi}\right) \Big/ 2$	$k - \dfrac{1}{2}$
W_{tot}	$W_t + W_c$	$\left(k - \dfrac{1}{2}\right) + \left(\dfrac{1}{4} - \dfrac{2}{\pi^2}\right) \Big/ \left(k - \dfrac{1}{2}\right)$

8.3.2 *The turn-off snubber circuit - assuming a cosinusoidal current fall*

As an alternative to a linear current fall at turn-off, it may be more realistic to assume that the current falls cosinusoidally according to

$$i_c(t) = \tfrac{1}{2} I_m (1 + \cos \pi t / T) \qquad \text{(A)} \qquad (8.20)$$

for $0 \le t \le T$, as shown in figure 8.15.

As with a linear current fall, two cases exist.

(i) $\tau \le T$ $(k \le 1)$, that is the snubber capacitor charges to V_s in time τ, before the switch current reaches zero, at time T.

(ii) $\tau \ge T$ $(k \ge 1)$, that is the snubber capacitor charges to the supply V_s after the switch current has fallen to zero.

These two cases are shown in figure 8.15 where k is defined as τ / T. Using a similar analysis as presented in the appendix (section 8.6), expressions can be derived for switch and snubber resistor losses. These and the total losses for each case are summarised in table 8.1.

Figure 8.15 shows that a minimum total loss occurs, namely

$$W_{total} = 0.41 \times \tfrac{1}{2} V_s I_m T \quad \text{at} \quad k = 0.62$$

when

$$C_s = 0.16 \frac{I_m T}{V_s} \qquad \text{(F)} \qquad (8.21)$$

For $t_{fi} < 0.85T$, a cosinusoidal fall current predicts lower total losses than a linear fall current, with losses shown in figure 8.13.

Example 8.4: ***Turn-off snubber design***

A 600V, 100A machine field winding is switched at 10kHz. The switch operates with an on-state duty cycle ranging between 5% and 95% ($5\% \le \delta \le 95\%$) and has a turn-off linear current fall time of 100ns, that is, $i_c(t) = 100 \times (1 - t / 100\text{ns})$.

i. Estimate the turn-off losses in the switch.
ii. Design a capacitive turn-off snubber using the dimensionally correct identity $i = Cdv/dt$. What is the capacitor voltage when the current reaches zero.
iii. Design a capacitive turn-off snubber such that the switch voltage reaches 600V as its conducting current reaches zero.

In each case calculate the percentage decrease in switch turn-off power dissipation.

Solution

i. The switch un-aided turn-off losses are given by equation (8.14). The turn-off time is greater than the current fall time (since the voltage rise time t_{rv} has been neglected), thus the turn-off switching losses will be greater than

$$W_{off} = \tfrac{1}{2} V_s I_m t_{off} = \tfrac{1}{2} \times 600\text{V} \times 100\text{A} \times 100\text{ns} = 3\text{mJ}$$

$$P_{off} = W_{off} \times f_s = 3\text{mJ} \times 10\text{kHz} = 30\text{W}$$

ii. Use of the equation $i=Cdv/dt$ results in a switch voltage that reaches the rail voltage after the collector current has fallen to zero. From $k = \frac{1}{2} + C_s V_s / I_m t_{fi}$ in figure 8.13, k=3/2 satisfies the dimensionally correct capacitor charging equation. Substitution into $i=Cdv/dt$ gives the necessary snubber capacitance

$$100\text{A} = C\,\frac{600\text{V}}{100\text{ns}}$$

$$\text{that is}\quad C = 16\tfrac{2}{3}\,\text{nF}$$

Use an 18nF, 1000V dc, metallised polypropylene, high *dv/dt* capacitor.
The snubber capacitor discharges at switch turn-on, and must discharge during the minimum on-time. That is

$$\check{t}_{on} = 5\,CR$$

$$5\%\text{ of } 1/10\text{kHz} = 5\times R\times 18\text{nF}$$

$$\text{that is}\quad R = 55.5\Omega \quad \text{Use } 56\Omega$$

The discharge resistor power rating is independent of resistance and is given by

$$P_{56\Omega} = \tfrac{1}{2} C V_s^2 f_s$$

$$= \tfrac{1}{2}\times 18\text{nF}\times 600\text{V}^2\times 10\text{kHz} = 32.4\text{W} \quad \text{Use 50W.}$$

The resistor can be wire-wound, the internal inductance of which reduces the initial peak current when the capacitor discharges at switch turn-on. The maximum discharge current in to the switch during reset, which is added to the 100A load current, is

$$I_{56\Omega} = V_s / R = 600\text{V}/56\Omega = 10.7\text{A}$$

which decays exponential to zero in five time constants, 5µs. The peak switch current (neglecting freewheel diode recovery) is 100A+10.7A=110.7A, at turn-on.
At switch turn-off, when the switch current reduces to zero, the snubber capacitor has charged to a voltage less than the 600V rail voltage, specifically

$$v_0 = \frac{1}{C}\int i_{cap}\,dt$$

$$= \frac{1}{18\text{nF}}\int_0^{100ns} 100\text{A}\times\left(\frac{t}{100\text{ns}}\right) dt = 277\text{V} \qquad (300\text{V with } 16\tfrac{2}{3}\,\text{nF})$$

The switch turn-off losses are reduced from 30W to

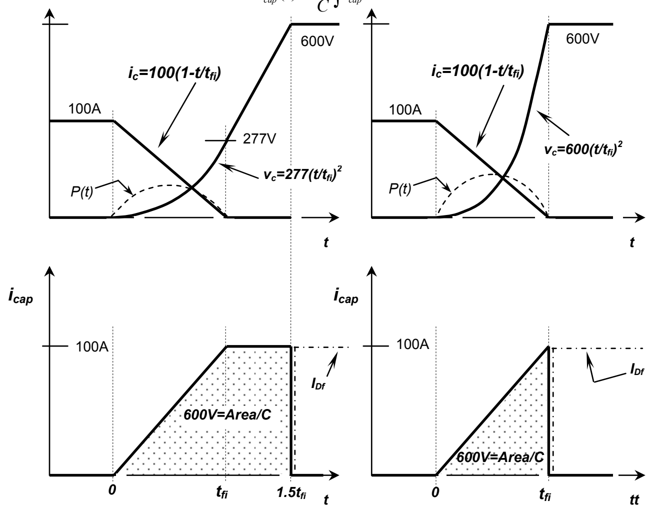
$$P_{off} = f_s \int_0^{100\text{ns}} i_c v_{ce}\,dt = f_s \int_0^{100\text{ns}} I_m\left(1\text{-}\frac{t}{100\,\text{ns}}\right) \times v_0\left(\frac{t}{100\text{ns}}\right)^2 dt$$

$$= f_s \int_0^{100\text{ns}} 100\text{A}\left(1\text{-}\frac{t}{100\,\text{ns}}\right) \times 277\text{V}\left(\frac{t}{100\text{ns}}\right)^2 dt = 2.3\text{W}$$

The total turn-off losses (switch plus snubber resistor) are 2.3W+32.4W=34.7W, which is more than the 30W for the unaided switch. Since the voltage rise time has been neglected in calculating the un-aided losses, 34.7W would be expected to be less than the practical un-aided switch losses. The switch losses have been reduced by 92⅓%.

iii. As the current in the switch falls linearly to zero, the capacitor current increases linearly to 100A (k = 1), such that the load current remains constant, 100A. The capacitor voltage increases in a quadratic function according to

$$v_{cap}(t) = \frac{1}{C}\int i_{cap}\,dt$$



Example 8.4. part (b) **Example 8.4. part (c)**

The capacitor charges quadratically to 600V in 100ns, as its current increases linearly from zero to 100A, that is

$$600\text{V} = \frac{1}{C}\int_0^{100ns} 100\,\frac{t}{100\text{ns}}\,dt$$

$$\text{that is } C = 8\tfrac{1}{3}\,\text{nF}$$

Use a 10nF, 1000V dc, metallised polypropylene, high *dv/dt* capacitor.
The necessary reset resistance to discharge the 10nF capacitor in 5µs is

$$5\mu\text{s} = 5\times R\times 10\text{nF}$$

$$\text{that is } R = 100\Omega$$

The power dissipated in the reset resistor is

$$P_{100\Omega} = \tfrac{1}{2} C V_s^2 f_s$$
$$= \tfrac{1}{2} \times 10\text{nF} \times 600\text{V}^2 \times 10\text{kHz} = 18\text{W}$$

Use a 100Ω, 25W, wire-wound, 600V dc withstand voltage, metal clad resistor.

The resistance determines the current magnitude and the period over which the capacitor energy is dissipated. The resistance does not determine the amount of energy dissipated. The capacitor exponentially discharges with an initial current of 600V/100Ω = 6A, which adds to the 100A load current at switch turn-on. The peak switch current is therefore 100A+6A = 106A, at turn-on.

The energy dissipated in the switch at turn-off is reduced from 30W when un-aided to

$$P_{off} = f_s \int_0^{100\text{ns}} i_c v_c \, dt = f_s \int_0^{100\text{ns}} I_m \left(1 - \frac{t}{100\,\text{ns}}\right) \times V_s \left(\frac{t}{100\text{ns}}\right)^2 dt$$
$$= f_s \int_0^{100\text{ns}} 100\text{A} \left(1 - \frac{t}{100\,\text{ns}}\right) \times 600\text{V} \left(\frac{t}{100\text{ns}}\right)^2 dt = 5\text{W}$$

The total losses (switch plus snubber resistor) with a turn-off snubber are 5W+18W =23W, which is less than the 30W for the unaided switch. The switch losses have been decreased by 83⅓%.

Note that the losses predicted by the equations in figure 8.13 amount to 5W+15W =20W. The discrepancy is due to the fact that the preferred value of 10nF (rather that the calculated 8⅓nF, k =1.2) has been used for the resistor loss calculation.

♣

8.3.3 *The turn-on snubber circuit - with non-saturable (air-core) inductance*

A turn-on snubber comprises an inductor-diode combination in the collector circuit as shown in figure 8.16. At turn-on the inductor controls the rate of rise of collector current and supports a portion of the supply voltage while the collector voltage falls. At switch turn-off the energy stored in the inductor, $\tfrac{1}{2} L_s I_m^2$, is transferred by current through the diode and dissipated in the diode D_s and in the resistance of the inductor.

Figure 8.17 shows collector turn-on waveforms with and without a turn-on snubber circuit. The turn-on losses associated with an unaided switch, figure 8.17a, neglecting the current rise time, are given by

$$W = \tfrac{1}{2} V_s I_m t_{fv} \qquad \text{(J)} \qquad (8.22)$$

where it is assumed that the collector current rise time is zero and that the collector voltage falls linearly, according to $v_c(t) = V_s(1 - t / t_{fv})$.

When an inductive turn-on snubber circuit is employed, collector waveforms as in figure 8.17b or 8.17c result.

The two possibilities and the associated circuit voltage and current waveforms in each case are shown in detail in figure 8.18. The waveforms are based on satisfying Kirchhoff's voltage and current laws for each case.

For low inductance the collector current reaches its maximum value I_m before the collector voltage has reached zero. As shown in figure 8.17b, the collector current increases quadratically $i_c(t)=I_m(t/\tau)^2$ and the total turn-on losses (switch plus snubber resistor) are given by

$$W_t = \tfrac{1}{2} V_s I_m t_{fv} \left(k^2 - \tfrac{4}{3}k + 1\right) \qquad \text{(J)} \qquad (8.23)$$

for $k \le 1$, where $k = \tau/t_{fv}$ as defined in figure 8.17. These losses include both switch losses and stored inductor energy subsequently dissipated. For higher snubber inductance, the collector voltage reaches zero before the collector current reaches the load current level. Initially the inductor current increases quadratically $i_{Ls}(t)=i_0(t/t_{fv})^2$, then when the collector voltage has reached zero, the current increases linearly. The total losses are given by

$$W_t = \tfrac{1}{2} V_s I_m t_{fv} \left(k^2 - k + \tfrac{1}{3}\right) \big/ \left(k - \tfrac{1}{2}\right) \qquad \text{(J)} \qquad (8.24)$$

Note that these equations are similar to those for the turn-off snubber, except that the current fall time is replace by the voltage fall time. The normalised loss components for the capacitive snubber in figure 8.13 are valid for the inductive turn-on snubber.

Minimum total turn-on losses of 5/9 those of the un-aided case, occur at k = ⅔ when

$$L_s = \frac{2}{9} \frac{V_s t_{fv}}{I_m} \qquad \text{(H)} \qquad (8.25)$$

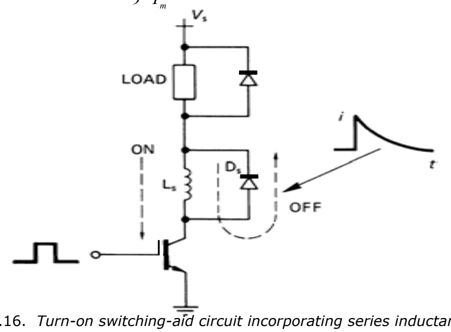


Figure 8.16. *Turn-on switching-aid circuit incorporating series inductance, L_s.*

At switch turn-off, the snubber inductance stored energy is dissipated as heat in the snubber freewheeling diode path. The maximum power loss magnitude is dependent on the operating frequency and is given by

$$P_{L_s} = \tfrac{1}{2} L_s I_m^2 f_s \qquad \text{(W)} \qquad (8.26)$$

This power is dissipated in both the inductor winding resistance and freewheeling diode D_s. The resistance in this loop is usually low and therefore a long L/R dissipating time constant may result. The time constant is designed such that $\check{t}_{off} = 5L_s/R$ where $\check{t}_{off}$ is the minimum device off-time. The time constant can be reduced either by adding series resistance or a Zener diode as shown in figure 8.19.

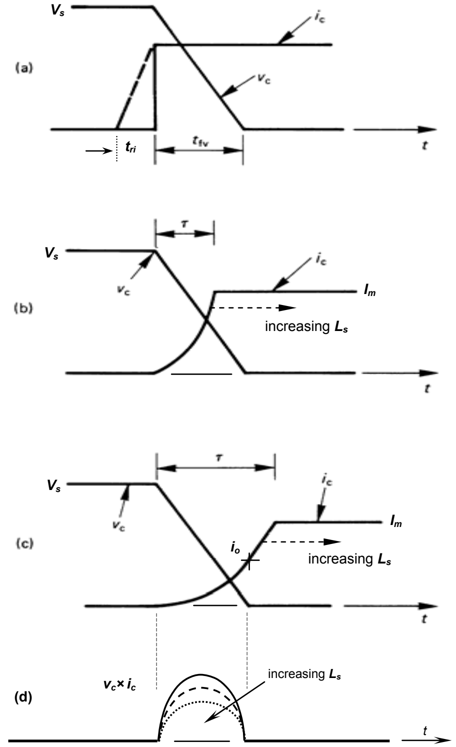


Figure 8.17. *Switch voltage and current collector waveforms at turn-on: (a) without a snubber; (b) and (c) with an inductive snubber; and (d) switch power losses.*

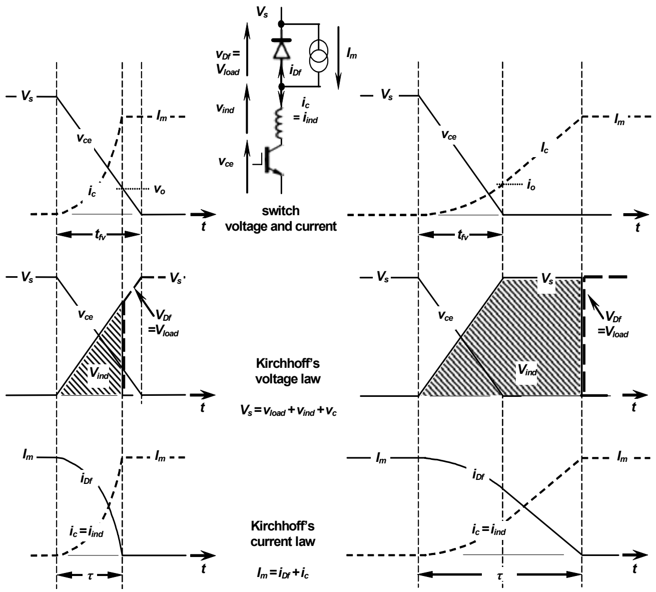


Figure 8.18. *Turn-on snubber waveforms satisfying Kirchhoff's laws: (a) turn-on with small snubber inductance and (b) turn-on with large snubber inductance.*

A disadvantage of adding series resistance R as in figure 8.19a is that the switch collector voltage at turn-off is increased from V_s to $V_s + I_mR$. The resistor must also have low self-inductance in order to allow the collector current to rapidly transfer from the switch to the resistor/diode reset circuit. The advantage of using a Zener diode as in figure 8.19b is that the maximum overvoltage is fixed, independent of the load current magnitude. For a given overvoltage, the Zener diode absorbs the inductor-stored energy quicker than would a resistor (see example 6.3 and problem 8.9). The advantages of using resistive dissipation are lower costs and more robust heat dissipation properties.

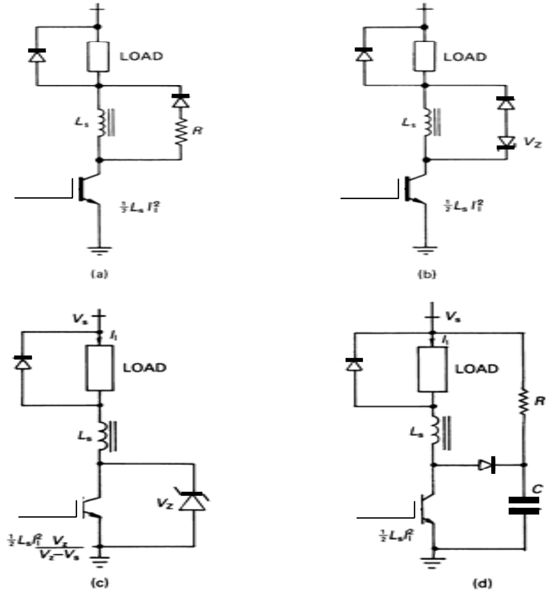


Figure 8.19. *Four turn-on snubber modifications for increasing the rate of release of inductor L_s stored energy: (a) using a power resistor; (b) using a power Zener diode; (c) parallel switch Zener diode, $V_Z > V_s$; and (d) using a soft voltage clamp.*

Alternatively the Zener diode can be placed across the switch as shown in figure 8.19c. The power dissipated is increased because of the energy drawn from the supply, through the inductor, during reset. At higher power, the soft voltage clamp shown in figure 8.19d, and considered in section 8.2, can be used. At switch turn-off, the energy stored in L_s, along with energy from the supply, is transferred and stored in a clamp capacitor. Simultaneously energy is dissipated in R and returned to the supply as the capacitor voltage rises. The advantage of this circuit is that the capacitor affords protection directly across the switch, but with lower loss than a Zener diode as in figure 8.19c. The energy loss equation for each circuit is also shown in figure 8.19. Figure 8.20 shows how a switch turn-on snubber circuit modifies the SOA trajectory during switch-on, avoiding a condition of simultaneous maximum voltage V_s and current I_m.

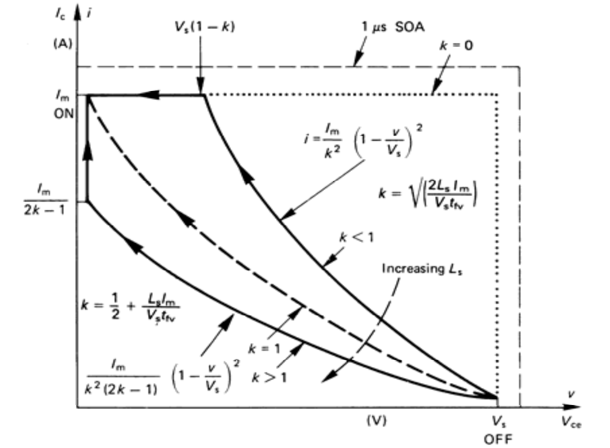


Figure 8.20. *The collector I-V trajectory at turn-on with a switching-aid circuit.*

Example 8.5: ***Turn-on air-core inductor snubber design***

A 600V, 100A machine field winding is switched at 10kHz. The switch operates with an on-state duty cycle between 5% and 95% (5% ≤ δ ≤ 95%) and has a turn-on voltage fall time of 100ns, that is, $v_c(t) = 600\text{V}(1 - t/100\text{ns})$.

i. Estimate the turn-on losses of the switch.
ii. Design an inductive turn-on snubber using the dimensionally correct identity *v=Ldi/dt*. What is the current magnitude in the turn-on inductor when the switch voltage reaches zero.
iii. Design an inductive turn-on snubber such that the switch current reaches 100A as its supported voltage reaches zero.

In each snubber case, using first a resistor and second a Zener diode for inductor reset, calculate the percentage decrease in switch power dissipation at turn-on.

Solution

i. The switch un-aided turn-on losses are given by equation (8.13). The turn-on time is greater than the voltage fall time (since the current rise time t_{ri} has been neglected), thus the turn-on switching losses will be greater than

$$W_{on} = ½ V_s I_m t_{on} = ½ \times 600\text{V} \times 100\text{A} \times 100\text{ns} = 3\text{mJ}$$
$$P_{on} = W_{on} \times f_s = 3\text{mJ} \times 10\text{kHz} = 30\text{W}$$

ii. Use of the equation *v=Ldi/dt* results in a switch current that reaches the load current magnitude after the collector voltage has fallen to zero. From $k = ½ + L_s I_m / V_s t_{fv}$ in figure 8.20, *k*=3/2 satisfies the dimensionally correct inductor equation. Substitution into *v=Ldi/dt* gives the necessary snubber inductance

$$600\text{V} = L\frac{100\text{A}}{100\text{ns}}$$
that is $L = 600$ nH

The snubber inductor releases its stored energy at switch turn-off, and must discharge during the switch minimum off-time, $\check{t}_{off}$. That is

$$\check{t}_{off} = 5L / R$$
$$5\% \text{ of } 1/10\text{kHz} = 5 \times 0.6\mu\text{H} / R$$
that is $R = 0.6\ \Omega$

Use the preferred value 0.68Ω, which reduces the *L/R* time constant.
The discharge resistor power rating is independent of resistance and is given by

$$P_{0.68\Omega} = ½ L I_m^2 f_s$$
$$= ½ \times 600\text{nH} \times 100\text{A}^2 \times 10\text{kHz} = 30\text{W}$$

The resistor in the circuit in figure 8.19a must have low inductance to minimise voltage overshoot at switch turn-off. Parallel connection of metal oxide resistors may be necessary to fulfil both resistance and power rating requirements. The maximum switch over-voltage at turn-off, (assuming zero resistor inductance), at the commencement of core reset, which is added to the supply voltage, 600V, is

$$V_{0.68\Omega} = I_m R = 100\text{A} \times 0.68\Omega = 68\text{V}$$

which decays exponential to zero volts in five time constants, 5μs. The maximum switch voltage is 600V+68V=668V, at turn-off. The reset resistor should be rated at 0.68Ω, 30W, metal film, 750V dc working voltage.
A Zener diode, as in figure 8.19b, of $V_z = L I_m / \check{t}_{off} = 0.6\mu\text{H} \times 100\text{A}/5\mu\text{s} = 12\text{V}$, will reset the inductor in the same time as 5 *L/R* time constants. The switch voltage is clamped to 612V during the 5μs inductor reset time at turn-off.
At turn-on when the switch voltage reduces to zero, the snubber inductor current (hence switch current) is less than the load current, 100A, specifically

$$i_0 = \frac{1}{L}\int v_{ind}\, dt$$
$$= \frac{1}{600\text{nH}} \int_0^{100\text{ns}} 600\text{V} \times \left(\frac{t}{100\text{ns}}\right) dt = 50\text{A}$$

The switch turn-on losses are reduced from 30W to

$$P_{on} = f_s \int_0^{100\text{ns}} i_c v_c\, dt = f_s \int_0^{100\text{ns}} V_s \left(1 - \frac{t}{100\text{ns}}\right) \times i_0 \left(\frac{t}{100\text{ns}}\right)^2 dt$$
$$= f_s \int_0^{100\text{ns}} 600\text{V} \left(1 - \frac{t}{100\text{ns}}\right) \times 50\text{A} \left(\frac{t}{100\text{ns}}\right)^2 dt = 2.5\text{W}$$

The total turn-on losses (switch plus snubber resistor) are 2.5W + 30W = 32.5W, which is more than the 30W for the unaided switch. Since the current rise time t_{ri} has been neglected in calculating the 30W un-aided turn-on losses, it would be expected that 32.5W would be less than the practical un-aided case. The switch losses are decreased by 92⅔%, from 30W down to 2.5W.

iii. As the voltage across the switch falls linearly to zero from 600V, the series inductor voltage increases linearly to 600V (*k*=1), such that the voltage sum of each component amounts to 600V. The inductor current increases in a quadratic function according to

$$i_{ind}(t) = \frac{1}{L}\int v_{ind}\, dt$$

The inductor current increases quadratically to 100A in 100ns, as its voltage increases linearly from zero to 600V, that is

$$100\text{A} = \frac{1}{L}\int_0^{100ns} 600\text{V}\ {}^{t}\!/_{100\text{ns}}\, dt$$
that is L = 300nH

The necessary reset resistance to reduce the 300nH inductor current to zero in 5μs is

$$\check{t}_{off} = 5\mu\text{s} = 5 \times 0.3\mu\text{H} / R$$
that is $R = 0.3\Omega$

Use the preferred value 0.33Ω in order to reduce the time constant.
The power dissipated in the 0.33Ω reset resistor is

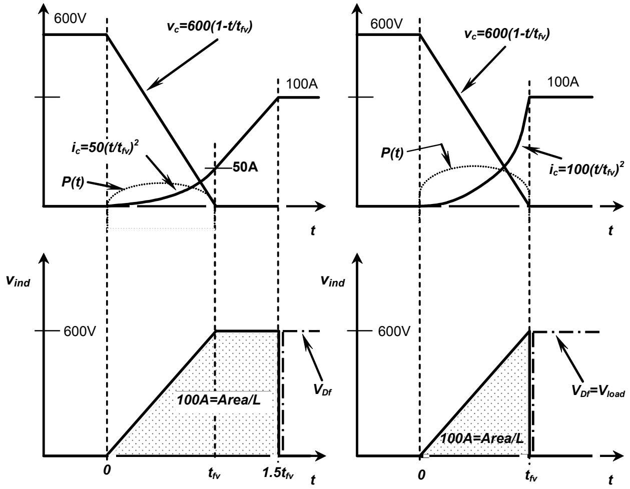
$$P_{0.33\Omega} = ½ L I_m^2 f_s$$
$$= ½ \times 300\text{nH} \times 100\text{A}^2 \times 10\text{kHz} = 15\text{W}$$

The resistance determines the voltage magnitude and the period over which the inductor energy is dissipated, not the amount of inductor energy to be dissipated. The inductor peak reset voltage is 100A×0.33Ω=33V, which is added to the supply voltage of 600V, giving 633V across the switch at turn-off. That is, use an 0.33Ω, 15W metal film, 750V dc working voltage resistor.
A Zener diode, as in figure 8.19b, of $V_z = L I_m / \check{t}_{off} = 0.3\mu\text{H} \times 100\text{A}/5\mu\text{s} = 6\text{V}$ (use 6.8V), will reset the inductor in the same time as 5 *L/R* time constants. The switch voltage is clamped to 606.8V during the $\check{t}_{off} = 5\mu\text{s}$ inductor reset time at turn-off.
The energy dissipated in the switch at turn-on is reduced from 30W to

$$P_{on} = f_s \int_0^{100\text{ns}} i_c v_c dt \;= f_s \int_0^{100\text{ns}} V_s \left(1\text{-}\frac{t}{100\text{ns}}\right) \times I_m \left(\frac{t}{100\,\text{ns}}\right)^2 \; dt$$

$$= f_s \int_0^{100\text{ns}} 600\text{V}\left(1\text{-}\frac{t}{100\text{ns}}\right) \times 100\text{A}\left(\frac{t}{100\,\text{ns}}\right)^2 \; dt = 5\text{W}$$

The total turn-on snubber losses (switch plus snubber resistor) are 5W+15W=20W, which is less than the 30W for the unaided switch. The switch losses, with an inductive turn-on snubber, are decreased by 83⅓%, from 30W to 5W.



Example 8.5. part (b) **Example 8.5. part (c)**

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8.3.4 The turn-on snubber circuit - with saturable ferrite inductance

The purpose of a turn-on snubber circuit is to allow the switch collector voltage to fall to zero while the collector current is low. Device turn-on losses are thus reduced, particularly for inductive loads, where during switching the locus point (V_s, I_m) occurs. This turn-on loss reduction effect can be achieved with a saturable inductor in the

circuit shown in figure 8.21a, rather than using a non-saturable (air core) inductor as previously considered in section 8.3.3. The saturable inductor in the snubber circuit is designed to saturate after the collector voltage has fallen to zero, at point **Y** in figure 8.21. Before saturation the saturable inductor presents high reactance and only a low magnetising current flows. From Faraday's equation, assuming the collector voltage fall to be linear, $V_s\left(1 - t / t_{fv}\right)$, the saturable inductor ℓ_s must satisfy

$$v_\ell = N\frac{d\phi}{dt} = NA\frac{dB}{dt} \qquad (8.27)$$

Rearranging, using an inductor voltage $v_\ell(t) = V_s - v_c(t) = V_s\; t / t_{fv}$, and integrating gives

$$B_s = \frac{1}{NA}\int_0^{t_{fv}} v_\ell(t)\; dt = \frac{1}{NA}\int_0^{t_{fv}} V_s \frac{t}{t_{fv}} dt \qquad (8.28)$$

which yields the following identity

$$V_s = \frac{2NA\;B_s}{t_{fv}} \qquad \text{(V)} \qquad (8.29)$$

where N is the number of turns,
A is the core area, and
B_s is the core ferro-magnetic material saturation flux density.

The inductor magnetising current I_M should be much less than the load current magnitude I_m, $I_M << I_m$, and the magnetising current at saturation is given by

$$I_M = H_s L_{eff} / N \qquad \text{(A)} \qquad (8.30)$$

where L_{eff} is the core effective flux path length and H_s is the magnetic flux intensity at the onset of saturation. Before core saturation the inductance is given by

$$L = N\Phi / I = N^2 / \Re = \mu_0 \mu_r A N^2 / L_{eff} \qquad \text{(H)} \qquad (8.31)$$

When the core saturates the inductance falls to that of an air core inductor (μ_r=1) of the same turns and dimensions, that is, the incremental inductance is

$$L_{sat} = \mu_0 A N^2 / L_{eff} \qquad \text{(H)} \qquad (8.32)$$

The energy stored in the inductor core is related to the *B-H* area shown in figure 8.21c and magnetic volume, and is approximated by

$$W_{\ell_s} = ½ B_s H_s A L_{eff} = ½ L I_M^2 \qquad \text{(J)} \qquad (8.33)$$

The collector turn-on waveforms are shown in figure 8.21b, while the corresponding *B-H* curve and SOA trajectories are illustrated in figure 8.21 parts c and d. It will be seen in figure 8.21b that little device turn-on electrical stressing occurs.

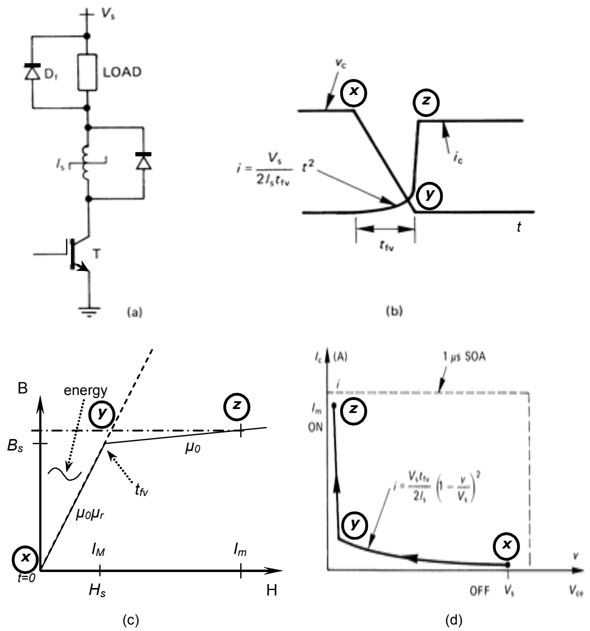


Figure 8.21. *Switch turn-on characteristics when a saturable inductor is used in the turn-on snubber: (a) circuit diagram; (b) collector voltage and current waveforms; (c) magnetic core B-H curve trajectory; and (d) safe operating area I-V turn-on trajectory.*

Example 8.6: ***Turn-on ferrite-core inductor snubber design***

A 600V, 100A machine field winding is switched at 10kHz. The switch operates with an on-state duty cycle between 5% and 95% (5% ≤ δ ≤ 95%) and has a turn-on voltage fall time of t_{fv}=100ns, that is, $v_c(t) = 600\text{V}(1 - t/100\text{ns})$.

i. Design a saturable inductor turn-on snubber that saturates as the collector voltage reaches zero, using a ferrite core with the following parameters.

- A = 0.4 sq cm
- L = 4cm
- B_s = 0.4T
- H_s = 100At/m

ii. Calculate the switch losses at turn-on when using the saturable reactor. What is the percentage reduction in switch turn-on losses?

iii. If an air cored inductor is used to give the same switch turn-on loss, what are the losses at reset?

Solution

From *example 8.5*, the unaided switch turn-on loss is 30W.

i. From equation (8.29) the number of turns is

$$N = ½ V_s t_{fv} / AB_s$$
$$= ½ \times 600\text{V} \times 0.1\mu\text{s} / 0.4 \times 10^{-4} \times 0.4\text{T} \approx 2 \text{ turns}$$

The magnetising current I_M at saturation, that is, when the collector voltages reaches zero, is given by equation (8.30)

$$I_M = H_s L_{eff} / N$$
$$= 100\text{At/m} \times 0.04 / 2 = 2\text{A}$$

Since I_M<<I_m, (2A<<100A), this core with 2 turns produces satisfactory turn-on snubber action, resulting in greatly reduced switch losses at turn-on.

From equation (8.31) the inductance before saturation is

$$L = NAB_s / I_M$$
$$= 2 \times 0.4 \times 10^{-4} \times 0.4\text{T} / 2\text{A} = 16\mu\text{H}$$

The incremental inductance after saturation, from equation (8.32), is given by

$$L_{sat} = N^2 A \mu_0 / L_{eff}$$
$$= 2^2 \times 0.4 \times 10^{-4} \times 4\pi \times 10^{-7} / 0.04 = 50\text{nH}$$

From equation (8.33) the energy stored in the core and released as heat in the reset resistor is

$$W_L = ½ L I_M^2 \quad \left(= ½ B_s H_s L_{eff} A\right)$$
$$= ½ \times 16\mu\text{H} \times 2^2 = 32\mu\text{J}$$
$$P_L = W_L \times f_s = 32\mu\text{J} \times 10\text{kHz} = 0.32\text{W}$$

The time $\check{t}_{off}$ for core reset via the resistor in five *L/R* time constants, is dominated by the 16µH section (the pre-saturation section) of the *B-H* curve, thus

$$\check{t}_{off} = 5\mu\text{s} = 5 \times 16\mu\text{H} / R$$

that is R = 16Ω

Use a 15Ω, 1W, carbon composition resistor, for low inductance.

This resistance results in a switch voltage increase above 600V of 15Ω×100A=1500V at turn-off. This high-voltage may be impractical in terms of the switch and resistor voltage ratings.

Alternatively, the Zener diode clamps shown in figures 8.19 b or c, may be suitable to dissipate the 0.32W of stored magnetic energy. The Zener voltage is determined by assuming that a fixed Zener voltage results in a linear decrease in current from 2A to zero in 5μs. That is

$$W_L = V_Z \int_0^{\check{t}_{off}} i_{ind}\,dt \quad \left(= \tfrac{1}{2} L I_M^2\right)$$

$$32\mu\text{J} = \tfrac{1}{2} \times V_Z \times 2\text{A} \times 5\mu\text{s}$$

$$\text{that is } V_Z = 6.4\text{V}$$

Use a 6.8V, 1W Zener diode.

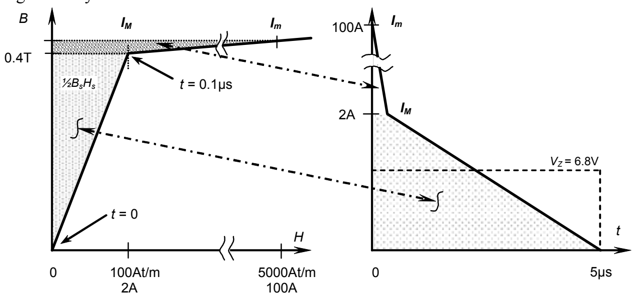
Series connected Zener diodes in parallel with the switch would need to dissipate 30W. The energy associated with saturation is small and is released in an insignificant time compared to the 5μs minimum off-time. The advantage of the Zener diode clamping approach, as opposed to using a resistor, is that the maximum switch voltage is clamped to 606.8V, even during the short, low energy period when the inductor current falls from 100A to 2A.

ii. The switch turn-on losses with the saturable reactor are given by

$$P_{on} = f_s \int_0^{100\text{ns}} i_c v_c\,dt = f_s \int_0^{100\text{ns}} V_s\left(1 - \frac{t}{100\text{ns}}\right) \times I_M \left(\frac{t}{100\text{ns}}\right)^2 dt$$

$$= f_s \int_0^{100\text{ns}} 600\text{V}\left(1 - \frac{t}{100\text{ns}}\right) \times 2\text{A}\left(\frac{t}{100\text{ns}}\right)^2 dt = 0.1\text{W}$$

The switch losses at turn-on have been reduced from 30W to 0.1W, a 99⅔% decrease in losses. The total losses (switch plus Zener diode) are 0.1W+0.32W=0.42W, which is significantly less than the 30W in the un-aided case.



Example 8.6

iii. If an air core inductor of 16μH were to replace the saturable reactor, the stored energy released would give losses

$$W = \tfrac{1}{2} L I_m^2$$

$$= \tfrac{1}{2} \times 16\mu\text{H} \times 100^2 = 80\text{mJ}$$

$$P = W \times f_s = 80\text{mJ} \times 10\text{kHz} = 800\text{W}$$

Clearly the use of an air cored inductor rather than a saturable reactor, to achieve the same switch loss of 0.1W at turn-on, is impractical.

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8.3.5 The unified snubber circuit

Figure 8.21 shows a switching circuit which incorporates both a turn-on and turn-off snubber circuit. Both C_s and L_s are dimensioned by the analysis outlined in sections 8.3.1 and 8.3.3, respectively. The power rating of the dissipating resistor *R* incorporates a contribution from both the turn-on inductor L_s and turn-off capacitor C_s, according to

$$P_{R_s} = \tfrac{1}{2}\left(L_s I_m^2 + C_s V_s^2\right) f_s \quad \text{(W)} \qquad (8.34)$$

Calculated resistance values to satisfy minimum off and on time reset according to $\check{t}_{on} \ge 5R_sC_s$ and $\check{t}_{off} \ge 5L_s/R_s$, may result in irreconcilable resistance requirements. The snubber capacitor discharges at turn-on via an *L-C-R* circuit rather than the usual *R-C* circuit, hence reducing the turn-on current stressing of the switch.

An important by-product from using a turn-on snubber circuit is that the inductor controls the reverse recovery process of the load freewheeling diode at switch turn-on.

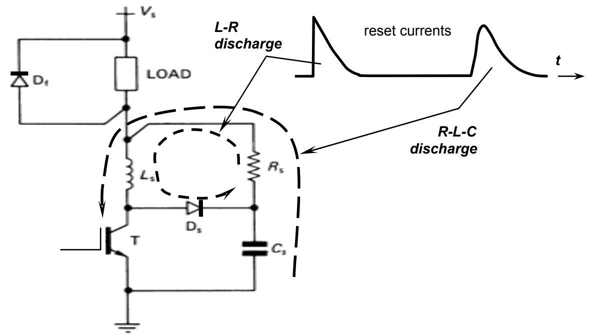

Figure 8.21. *Unified snubber incorporating both a turn-on and a turn-off circuit which share the one dissipation reset resistor.*

8.4 Snubbers for bridge legs

Figure 8.23 shows three typical switch bridge leg configurations used in inverters as shown in figures 14.1 and 14.3. The inductive turn-on snubber L and capacitive turn-off snubber C_s are incorporated into the bridge legs as shown in each circuit in figure 8.23.

The combinational snubber circuit in figure 8.23a can be used to minimise the number of snubber components. The turn-on snubber inductance L_s, reset resistor R, and snubber capacitor C_{sc}, are common to any number of bridge legs. The major disadvantage of this circuit is that turn-off snubber action associated with the lower switch is indirect, relying on low inductance decoupling through C_s and C_{sc}.

With an inductive load, unwanted turn-off snubber action occurs during the switch modulation sequence as shown in figures 8.23b and 8.23c. When the upper switch T_u is turned off as in figure 8.23b the load current I_m is diverted to the freewheel diode D_f. While D_f conducts the capacitor C, discharges to zero through the resistor R, as shown, dissipating energy $½C_sV_s^2$. When the switch T_u is turned on, the load current is provided via the switch T_u and the snubber capacitor C_s is charged through the series turn-on snubber inductance, as shown in figure 8.23c. A lightly damped L-C oscillation occurs and C_s is over charged. Advantageously, the recovery voltage of the freewheel diode D_f is controlled by the capacitor voltage rise.

The unwanted snubber action across the non-power conducting switch can be avoided in some applications by using a series blocking diode as shown in figure 8.23d. The diode D_b prevents C_s from discharging into the load as occurs with the lower switch in figure 8.23b. A blocking diode can be used to effectively disable the internal parasitic diode of the MOSFET. Adversely, the blocking diode increases the on-state losses.

In reactive load applications, bridge legs are operated with one switch on, with only a short underlap when both switches are off. Thus although the snubber capacitor cannot discharge into the load in figure 8.23d, it always discharges through the switch T, regardless of load current flow through the switch.

In IGBT and MOSFET applications, the conventional R-C-D turn-off snubber is not usually required. But because of diode recovery limitations, a turn-on snubber may be necessary. In low frequency applications, a single turn-on snubber inductor can be used in the dc link as shown in figure 8.24a. Snubber circuit design is based on the turn-on snubber presented in 8.3.3. The circuit in figure 8.24b is based on the conventional turn-on snubber being incorporated within the bridge leg. Figures 8.24c and d show turn-on snubbers which use the soft voltage clamp, presented in 8.2, to reset the snubber inductor current to zero at turn-off.

In each circuit at switch turn-off, t_3, the energy $½LI_m^2$ stored in the turn-on snubber inductor is dissipated in the resistor of the discharge circuit. The energy $½LI_{rm}^2$ in L, due to diode recovery, is dissipated in the resistor at time t_1, in circuits (a), (b) and (c). In figure 8.24d the energy in excess of that associated with the load, $½LI_m^2$, due to diode recovery, is dissipated in the switch and its parallel connected diode. At time t_1

$$W = ½LI_{rm}^2 + \frac{V_{ce}}{V_{ce}+V_{Df}} LI_{rm}I_m \tag{8.35}$$

is dissipated in the two semiconductor components. Since the energy is released into a low voltage $v_{ce} + v_{Df}$, the reset time $t_2 - t_1$ is large.

Coupling of the inductors in figures 8.24c and d does not result in any net energy savings.

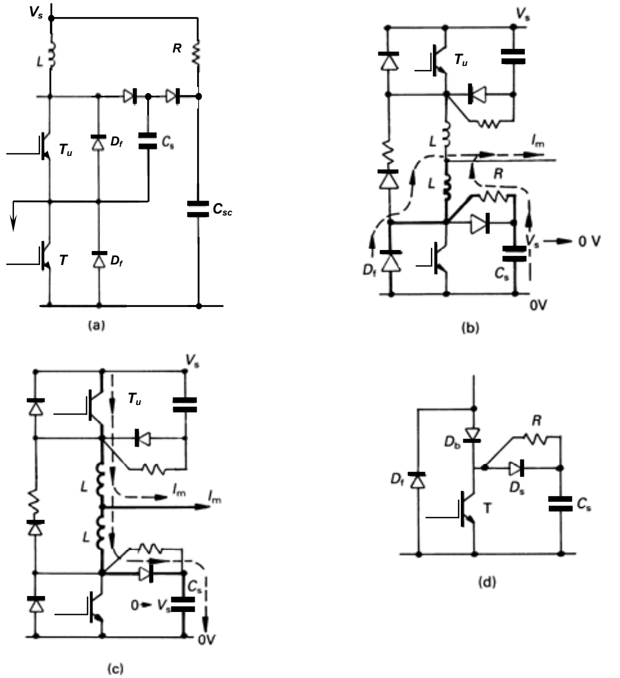

Figure 8.23. *Bridge leg configurations: (a) Undeland leg snubber circuit; (b) leg with turn-on snubber and turn-off snubber C_s discharge path shown; (c) L-C oscillation at switch-on; and (d) blocking circuit to prevent snubber capacitor discharge when D_f conducts.*

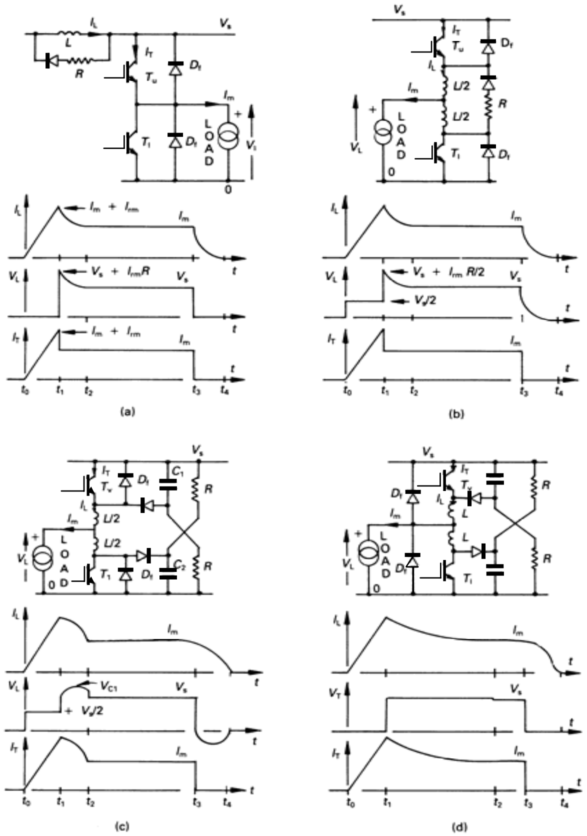

Figure 8.24. *Turn-on snubbers for bridge legs:*
(a) single inductor in dc link; (b) unified L-R-D snubber; (c) soft voltage clamp; and (d) soft voltage clamp with load clamped.

8.5 Appendix: Turn-off *R-C* snubber circuit analysis

When a step input voltage is applied to the *L-C-R* circuit in figure 8.3, a ramped voltage appears across the *R-C* part of the circuit. If this *dv/dt* is too large, a thyristor in the off-state will turn on as a result of the induced central junction displacement current, which causes injection from the outer junctions.
The differential equations describing circuit current operation are

$$(D^2 + 2\xi_0 D + \omega_0^2)\, I = 0 \qquad (8.36)$$

and

$$(\tau D + 1) I = CDe_0 \qquad (8.37)$$

where D = differential operator = d/dt
and ξ = damping ratio = $\tfrac{1}{2}R\sqrt{C/L}$
ω_0 = natural frequency = $1/\sqrt{LC}$
ω = oscillation frequency = $\omega_0\sqrt{1-\xi^2}$

Solution of equations (8.36) and (8.37), for I_o= 0, leads to
(a) The snubber current

$$I(t) = \frac{e_s}{R}\frac{2\xi}{\sqrt{1-\xi^2}}\, e^{-\xi\omega_0 t}\sin\omega t \qquad \text{(A)} \qquad (8.38)$$

(b) The rate of change of snubber current

$$\frac{dI}{dt} = \frac{e_s}{L}\; e^{-\xi\omega_0 t}\left(\cos\omega t - \frac{\xi}{\sqrt{1-\xi^2}}\sin\omega t\right) \qquad \text{(A/s)} \qquad (8.39)$$

(c) Snubber *R-C* voltage

$$e_0 = e_s\left(1 - e^{-\xi\omega_0 t}\cos\omega t - \frac{\xi}{\sqrt{1-\xi^2}}\sin\omega t\right) \qquad \text{(V)} \qquad (8.40)$$

(d) The rate of change of *R-C* voltage

$$\frac{de_0}{dt} = \omega_0 e_s\, e^{-\xi\omega_0 t}\left(2\xi\cos\omega t + \frac{1-2\xi}{1-\xi^2}\sin\omega t\right) \qquad \text{(V/s)} \qquad (8.41)$$

The maximum value expressions for each equation can be found by differentiation
(a) Maximum snubber current

$$I_p = \frac{e_s}{R}\, 2\xi\, e^{\left(-\xi\cos^{-1}\xi/\sqrt{1-\xi^2}\right)} \qquad \text{(A)} \qquad (8.42)$$

when $\cos\omega t = \xi$

(b) The maximum snubber *di/dt* is given by

$$\frac{dI_p}{dt} = \frac{e_s - e_0}{L} \qquad \text{(A/s)} \qquad (8.43)$$

(c) Maximum *R-C* voltage

$$\hat{e}_o = e_s\left(1 + e^{\left\{-\xi\cos^{-1}\xi \big/ \sqrt{1-\xi^2}\right\}}\right) \qquad \text{(V)} \qquad (8.44)$$

when $\cos \omega t = 2\xi^2 - 1$

(d) Maximum slew rate, $\dfrac{de_0}{dt} = \hat{S}$

for $\xi < ½$

$$\hat{S} = e_0\omega_0 e^{\left\{-\xi\cos^{-1}\xi\left(3-4\xi^2\right) \big/ \sqrt{1-\xi^2}\right\}} \qquad \text{(V/s)} \qquad (8.45)$$

when $\cos \omega = \xi(3-4\xi^2)$

for $\xi > ½$

$$\hat{S} = 2\xi\, e_s w_0 \qquad (= e_s R/L) \qquad \text{(V/s)} \qquad (8.46)$$

when $t = 0$

Equations (8.42) to (8.46), after normalisation are shown plotted in figure 8.4 as a function of the snubber circuit damping factor ξ. The power dissipated in the resistor is approximately $Ce_s^2 f_s$.

8.6 Appendix: Turn-off *R-C-D* switching aid circuit analysis

Switch turn-off losses for an unaided switch, assuming the collector voltage rise time is negligible compared with the collector current fall time, are

$$W = ½V_s I_m t_{fi} \qquad (8.47)$$

If τ is the time in figure 8.11 for the snubber capacitor C_s to charge to the supply V_s, and t_{fi} is the switch collector current fall time, assumed linear such that $i_c(t) = I_m(1-t/t_{fi})$, then two capacitor charging conditions can exist

- $\tau \le t_{fi}$
- $\tau \ge t_{fi}$

Let $k = \tau/t_{fi}$ and electrical energy $W = \int_0^t vi\, dt$

Case 1: $\tau \le t_{fi}$, $k \le 1$

Figure 8.11b shows ideal collector voltage and current waveforms during aided turn-off for the condition $t \le t_{fi}$. If, assuming constant maximum load current, I_m, the collector current falls linearly, then the load deficit, $I_m t/t_{fi}$, charges the capacitor C_s, whose voltage therefore increases quadratically. The collector voltage v_c and current i_c are

given by

$$\begin{bmatrix} i_c(t) = I_m(1-\dfrac{t}{t_{fi}}) \\ v_c(t) = V_s\left(\dfrac{t}{\tau}\right)^2 \end{bmatrix},\ 0 \le t \le \tau \qquad \text{and} \qquad \begin{bmatrix} i_c(t) = I_m(1-\dfrac{t}{t_{fi}}) \\ v_c(t) = V_s \end{bmatrix},\ 0 \le t \le t_{fi} \qquad (8.48)$$

The final capacitor charge is given by

$$Q = C_s V_s = \int_0^\tau (I_m - i_c(t))dt = ½I_m t_{fi} k^2 \qquad \text{(C)} \qquad (8.49)$$

The energy stored by the capacitor, W_c, and lost in the switch, W_t, are given by

$$\begin{aligned} W_c &= ½C_s V_s^2 \qquad (= ½QV_s) \\ &= ½V_s I_m t_{fi} \times ½k^2 \qquad \text{(J)} \end{aligned} \qquad (8.50)$$

$$\begin{aligned} W_t &= \int_0^\tau V_s I_m (t/\tau)dt + \int_0^{t_{fi}} V_s I_m (1 - t/t_{fi})dt \\ &= ½V_s I_m t_{fi}\left(1 - \tfrac{4}{3}k + ½k^2\right) \qquad \text{(J)} \end{aligned} \qquad (8.51)$$

Case 2: $\tau \ge t_{fi}$, $k \ge 1$

Figure 8.11c shows the ideal collector voltage and current switch-off waveforms for the case when $k \ge 1$. When the collector current falls to zero the snubber capacitor has charged to a voltage, v_0, where

$$\begin{aligned} v_o &= \frac{1}{C_s}\int_0^{t_{fi}} i\, dt \\ &= \frac{1}{C_s} \times ½I_m t_{fi} \qquad \text{(V)} \end{aligned} \qquad (8.52)$$

The collector voltage v_c and current i_c are given by

$$\begin{bmatrix} i_c(t) = I_m(1-\dfrac{t}{t_{fi}}) \\ v_c(t) = V_s\left(\dfrac{t}{t_{fi}}\right)^2 \end{bmatrix},\ 0 \le t \le t_{fi} \ \text{and} \ \begin{bmatrix} i_c(t) = 0 \\ v_c(t) = \dfrac{1}{t_{fi}}\dfrac{\left(V_s - v_0\right)t}{k-1} + \dfrac{kv_0 - V_s}{k-1} \end{bmatrix},\ t_{fi} \le t \le \tau \qquad (8.53)$$

The final capacitor charge is given by

$$\begin{aligned} Q = C_s V_s &= \int_0^{t_{fi}} (I_m - i_c(t))dt + \int_{t_{fi}}^\tau I_m\, dt \\ &= I_m t_{fi}\left(k - ½\right) \qquad \text{(C)} \end{aligned} \qquad (8.54)$$

The energy stored by the capacitor W_c, and lost in the switch W_t, are given by

$$\begin{aligned} W_c &= \tfrac{1}{2} C_s V_s^2 \quad (= \tfrac{1}{2} Q V_s) \\ &= \tfrac{1}{2} V_s I_m t_{fi} \times \left(k - \tfrac{1}{2}\right) \qquad \text{(J)} \\ W_t &= \int_0^{t_{fi}} v_0 I_m (1 - t/t_{fi}) \left(t/t_{fi}\right)^2 dt \\ &= \tfrac{1}{12} v_0 I_m t_{fi} \qquad \text{(J)} \end{aligned} \tag{8.55}$$

Using equations (8.52) and (8.54) to eliminate v_0 yields

$$W_t = \tfrac{1}{2} V_s I_m t_{fi} \times \frac{1}{6(2k-1)} \qquad \text{(J)} \tag{8.56}$$

The total circuit losses W_{tot}, are

$$\begin{aligned} W_{total} &= W_t + W_c \\ W_{total} &= \tfrac{1}{2} V_s I_m t_{fi} \times \left(1 - \tfrac{4}{3} k + \tfrac{1}{2} k^2\right), \ \ k \le 1 \quad \text{(J)} \\ W_{total} &= \tfrac{1}{2} V_s I_m t_{fi} \times \frac{\left(k^2 - k + \tfrac{1}{3}\right)}{\left(k - \tfrac{1}{2}\right)}, \quad k \ge 1 \quad \text{(J)} \end{aligned} \tag{8.57}$$

The equations (8.50), (8.51), and (8.55) to (8.57) have been plotted, normalised, in figure 8.13.

Reading list

International Rectifier, *HEXFET Data Book,* HDB-5, 1987.

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

Siliconix Inc., *Mospower Design Catalog,* January 1983.

Graffiam, D. R. *et al., SCR Manual,* General Electric Company, 6th Edition, 1979.

Problems

8.1. The figure 8.25 shows GTO thyristor turn-off anode *I-V* characteristics. Calculate

i. turn-off power loss at 1 kHz. What percentage of the total loss does the tail current account for?

ii. losses when a capacitive turn-off snubber is used and the anode voltage rises quadratically to 600V in 0.5μs. What percentage of the total losses does the tail current account for? What is the necessary capacitance?

iii. losses when a capacitive turn-off snubber is used and the anode voltage rises quadratically to 600V in 2μs. What percentage of the total losses does the tail current account for? What is the necessary capacitance?

[10.5 W]

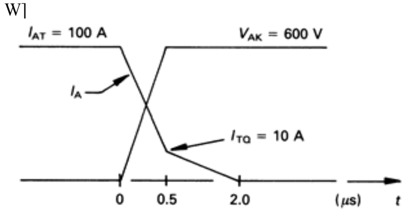

Figure 8.25. *Problem 8.1, GTO thyristor tail current characteristics.*

8.2. Prove that the minimum total losses (switch plus snubber resistor), associated with a switch which utilises a capacitive turn-off switching-aid circuit, occur if the snubber capacitor is fully charged when the collector current has fallen to ⅓ its original value. Derive an expression for this optimal snubber capacitance.

$$C_s = \tfrac{2}{9} \frac{I_m t_{fi}}{V_s} \qquad \text{(F)}$$

8.3. Derive an expression for the optimal turn-on switching-aid circuit inductance, assuming the collector current rise time in the unaided circuit is very short compared with the collector voltage fall time.

$$L_s = \tfrac{2}{9} \frac{V_s t_{fv}}{I_m} \qquad \text{(H)}$$

8.4. A ferrite toroid has *B-H* characteristics as shown in figure 8.26 and a cross-sectional area, *A,* of 10 mm^2 and effective length, L_{eff} of 50 mm.

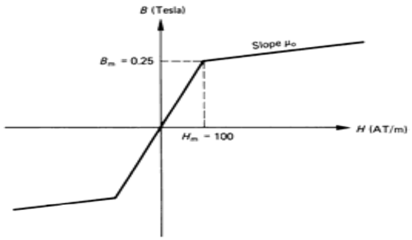

Figure 8.26. *Problem 8.4, B-H characteristics.*

A number of such toroid cores are to be stacked to form a core for a saturable inductor turn-on snubber in a switching circuit. The circuit supply voltage is V and the switch voltage fall time at turn-on is t_{fv}. Assume t_{fv} is independent of supply voltage and falls linearly from V to 0 V.

i. Using Faraday's Law, show that if the ferrite inductor is to saturate just as the switch collector voltage falls to zero at turn-on, then the number of turns N for n cores is given by

$$N = \frac{Vt_{fv}}{2B_m An}$$

ii. Derive an expression for the inductance before saturation.

iii. It is required that the maximum magnetising current before saturation does not exceed 1 A. If only 10 turns can be accommodated through the core window, what is the minimum number of cores required if V = 200 V and t_{fv} = 1 μs?

iv. How many cores are required if the supply V is increased to the peak voltage of the three-phase rectified 415 V ac mains, and the load power requirements are the same as in part (c)?

v. Calculate the percentage change in the non-saturated inductance between parts iii and iv.

vi. What are the advantages of saturable inductance over linear non-saturable inductance in turn-on snubber applications? What happens to the inductance and stored energy after saturation?

[$\ell = N^2/R$, n =5, n =4, 1:9]

8.5. Prove, for an inductive turn-on snubber, where the voltage fall is assumed linear with time, that

$$k = \sqrt{\frac{2L_s I_m}{V_s t_{fv}}} \quad \text{for } k \ge 1$$

$$k = \frac{L_s I_m}{V_s t_{fv}} + ½ \quad \text{for } k \le 1$$

where $k = t_{fv}/\tau$ (see figures 8.17 and 8.20).

8.6. Derive the expressions in table 8.1 for a turn-off snubber assuming a cosinusoidal current fall. Prove equation (8.21), the optimal capacitance value.

8.7 Show that when designing a capacitive turn-off snubber using the dimensionally correct equation $i = Cdv/dt$, as in example 8.4b, the capacitor charges to $½V_s$ when the switch current reaches zero.

8.8 Show that when designing an inductive turn-on snubber using the dimensionally correct equation $v = Ldi/dt$, as in example 8.5b, the inductor current reaches $½I_m$ when the switch voltage reaches zero.

8.9 Reset of inductive turn-on snubber energy $½L_s I_m^2$ can be effected through a resistor, R, as in figure 8.19a or through a Zener diode, D_z, as in figure 8.19b.
Show that for the same reset voltage, namely $V_z = I_m R$, in each case, Zener diode reset is n times faster the resistor reset when $nR_s C_s \le \check{t}_{on}$.