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Snubber Considerations for IGBT Applications

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Abstract - Snubber circuits can be used to protect fast switching IGBTs from turn-on and turn-off voltage transients. Snubbers are available in various configurations and a clear understanding of their operation is necessary to make the appropriate selection. This paper will discuss pros and cons of these circuits. Decoupling capacitors, RCD voltage clamp circuit, RCD charge-discharge snubbers are included in the discussion.

Introduction

When a power device is abruptly turned off, trapped energy in the circuit stray inductance is dissipated in the switching device, causing a voltage overshoot across the device. The magnitude of this transient voltage is proportional to the amount of stray inductance and the rate of fall or turn-off current. The situation is at its worst for fast switching IGBT modules. These devices switch at a high magnitude of currrents in a short duration of time, giving rise to potentially destructive voltage transients. The higher current modules normally consist of several IGBT chips in parallel. Each individual chip switches its share of the load currrent at a di/dt that is determined by the gate drive circuit. The total current and di/dt seen by the external power circuit is the sum of currents and di/dts through each IGBT chip. The di/dts produced could easily be a few thousand A/us. Proper attention needs to be paid to protect these devices from destruction.

It is determined that the snubbers offer optimized protection against voltage transients during the normal turn-on and turn-off switching. Usage of such protection circuits allow faster yet safer operation by containing the operating loci within the boundaries of the rated Safe Operating Area. This paper discusses various snubber circuits. Pros and cons of decoupling capacitors, PCD snubber clamp circuits, RCD charge-discharge snubbers are discussed. Circuit operations are analyzed and the test results are illustrated.

Note that the fault current transients are more effectively protected by considerably slowing the rate of fall of fault current. Fault current turn-off protection through electronic gate control is convered in detail elsewhere[1].

Decoupling Capacitors

As mentioned earlier, the magnitude of transient voltage depends on the trapped energy in the circuit stray inductance, also call "DC loop" inductance L_S. As a preventive measure, steps should be taken to improve the circuit layout. Usage of laminated copper plates, minimizing the size of the "DC loop" and choosing source capacitance with inherently low self inductance are ways to lower stray inductance [2]. Decoupling capacitors, connected across the module's bus terminals, as shown in Fig.1a, are found to be useful for low/medium current applications. High frequency polypropolene film capacitors available today are designed to fit IGBT terminal spacing for direct mounting. The resultant internal inductances are drastically lower than conventional leaded capacitors.

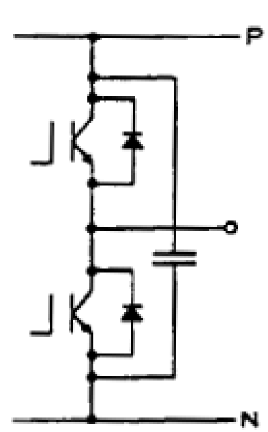


Figure 1a.
Decoupling capacitor

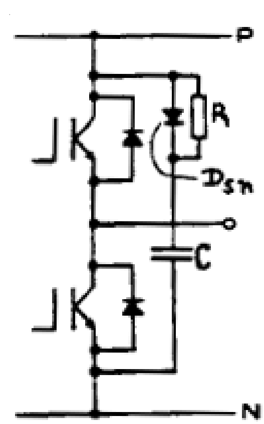
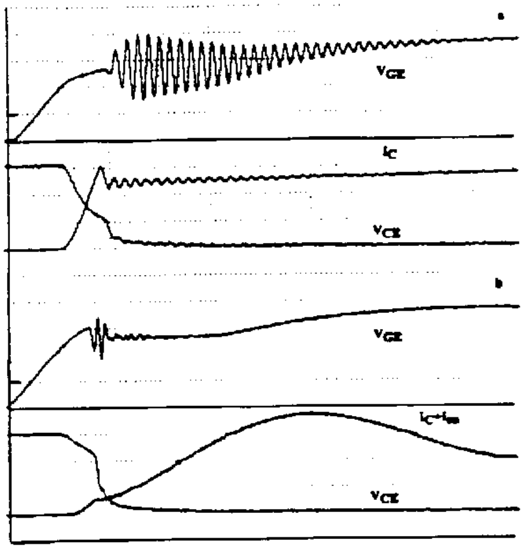
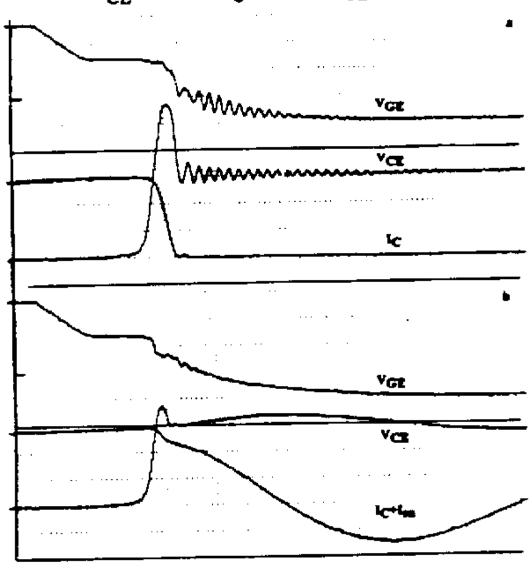


Figure 1b.
Discharge restricted decoupling capacitor.

Figs. 2a, 2b and 3a, 3b show turn-on and turn-off switching waveforms, with and without a 1μF module-compatible decoupling capacitor.



400ns/div
Figures 2a,b. Turn-on waveforms without & with decoupling capacitor, respectively.
Tested at 300V, 125A, 15V/39Ω, C_{sn}=1μF
V_{CE}: 100V/div, I_C: 50A/div, V_{GE}: 5V/div



400ns/div
Figures 3a,b. Turn-off waveforms without & with decoupling capacitor, respectively.
Tested at 300V, 150A, 15V/39Ω, C_{sn}=1μF
V_{CE}: 100V/div, I_C: 50A/div, V_{GE}: 5V/div

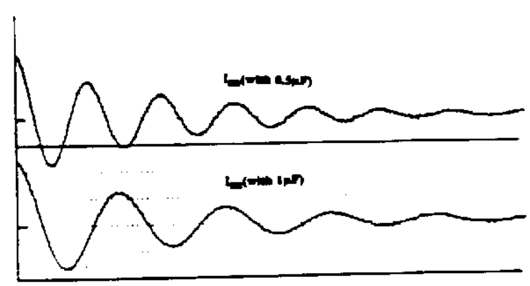
These figures show voltage across the IGBT (V_{CE}), the DC bus current (i.e. I_C in Figs. 2a and 3a and I_C+I_{sn} in Fig. 2b and 3b), and the gate voltage (V_{GE}). The actual current through IGBT is not recorded in Figs. 2b and 3b as the decoupling capacitor was directly screwed on to the module terminals and the path to the IGBT was not accessible for the current measurements. However, since the IGBT switching current is only marginally dependent on the effective external inductances, L_s (the principal difference between the two cases), it is assumed to be almost identical in the both cases. It can also be shown that under a given set of conditions, the total switching losses are independent of the external stray inductances[3].

It is clear from the above figures that the usage of decoupling capacitors, by providing "non" inductive path during the switching operation, eliminate severe voltage transients during switching and help smooth out the circuit waveforms. Note that in Fig. 3b the voltage transient during the turn-off is reduced to 370V from its dangerously high value of 570V in Fig. 3a (rated voltage was 600V).

Due to the fear of destroying the device by over voltage (caused by the recovering diode), the IGBT turn-on switching speed is limited by usage of higher gate resistors. With the decoupling capacitor in place, this constraint is removed. The gate resistor can now be reduced to a lower value, thereby allowing the IGBT to turn-on faster, and reducing overall switching losses.

The snubber capacitance value can be approximated from the following, based on the circuit stray inductance L_s, maximum switching current I_o, DC rail voltage V_{CC} and allowable peak voltage V_{pk} (Appendix II).

$$C_{sn} = L_s \cdot I_o^2 / (V_{pk} - V_{CC})^2 \qquad (1)$$



1μs/div
Figure 4. Snubber current waveforms at IGBT turn-off (with decoupling capacitor). With .5μF and 1μF capacitors.
Tested at 300V, 125A, 15V/39Ω
I_{sn}: 50A/div

The overriding criteria in selecting the capacitor may be the RMS current limit of the decoupling capacitor. IGBT switching sets off oscillations in the "DC loop", between the decoupling capacitor and the DC source. The oscillations are damped by the resistive stray elements in the loop. The resulting snubber currents are shown in Fig. 4, and given by the following (Appendix I).

$$i_{sn} = I_o\, e^{-\alpha t} \cos(\beta t) \qquad (2)$$

Where $\alpha = R/2L_s$ and $\beta = [4/L_sC_{sn} - (R/2L_s)^2]^{1/2}$, R being the stray resistance in the loop that include capacitor's ESR. From Appendix II I_{RMS} in the capacitor are given by:

$$I_{RMS} = I_o\, (f_{sw} \cdot L_s / R_s)^{-1/2} \qquad (3)$$

Since the capacitor's ESR is the dominant part of the total stray resistance, the worst case losses in the capacitor can be approximated as follows:

$$ESR \cdot I_{RMS}^2 = 2 \cdot (1/2\, L_s\, I_o^2) \cdot f_{sw} \quad (4)$$

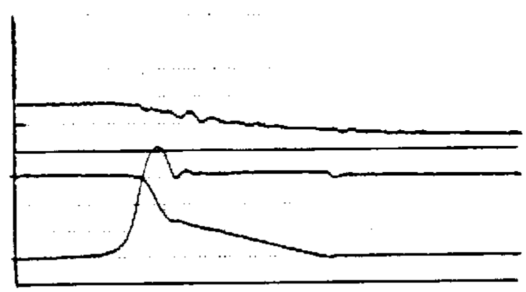
At higher frequencies and higher device currents, the oscillatory snubber currents may cause excessive heating in this high frequency capacitors.

The capacitor should therefore be selected based on the limiting values given by (1) and (3). For the load currents up to 150A or so, and switching frequency of up to a few kHz, the decoupling capacitors seem to provide the optimal protection against the normal switching transients. While this method provide low inductive path for the currents during switching, it is not a panacea for bad circuit layout, as higher L_s result in greater RMS currents in the capacitors (as seen form 3).

The discharge resistant circuit of figure 1b

The circuit in Fig. 1b operates by the same principal as the decoupling capacitor, but only during the turn-off switching. As the IGBT is turned off the energy trapped in the stray "DC loop" inductance is transferred to the capacitor. The diode D_{sn} however blocks off any oscillations from occurring. See Fig 5. The excess charge on the capacitor is gradually discharged through the snubber resistor.

The disadvantage of this circuit is that the added element in the protection circuit (i.e. the blocking diode) increases the overall snubber inductance. Also direct mounted, low inductance snubber capacitor can not be used in this case. This causes an increase in the V_{CE} overshoot as seen from Fig. 5, as compared to the one shown in Fig. 3b.



400ns/div

Figure 5. Turn-off waveforms with **Discharge Restricted decoupling capacitor.**

Tested at 300V, 150A, 15V/39Ω, C_{sn}=1μF/20Ω

V_{CE}: 100V/div, I_C: 50A/div, V_{GE}: 10V/div,

The turn-on operation of this circuit is exactly similar to the one for the RCD clamp circuit; to be discussed next. The snubber losses are given by (1) and (5), also in the next section.

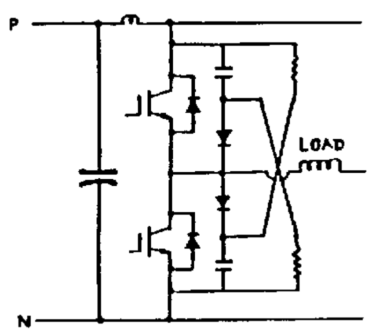


Figure 6a. RCD Clamp-Snubber

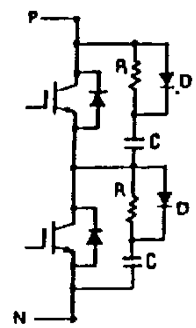


Figure 6b. RCD Charge-Discharge Snubber

RCD Snubber and clamp circuits

Figs. 6a and 6b are two principal examples of RCD snubbers for high current IGBT applications. While both circuits reduce transient voltages across switching devices, the charge-discharge snubber circuit in Fig. 6b is also targeted for reducing IGBT turn off losses. First, operation of the circuit in Fig. 6a will be considered.

The "clamp" circuit of figure 6a

The operation of this RCD circuit during turn-on and turn-off switchings are described separately below.

Turn-off: The function of this circuit is somewhat similar to a voltage clamp. During IGBT conduction time the snubber capacitors are charged to the bus voltage. As the IGBT is turned off the voltage across it, V_{CE}, rises rapidly. The circuit "DC loop" stray inductance, L_s, may cause V_{CE} to rise above the bus voltage. As this occurs, the snubber diode is forward biased and the snubber is activated. The energy trapped in the stray inductance now is diverted to the snubber capacitor which absorbs this incremental energy without substantial rise in its voltage. The waveforms shown in Fig. 7 clearly illustrate the turn-off behavior with and without the RCD clamp. The voltage overshoot has been substantially reduced from 210 Volts to only 50 volts. Initially a small stray inductance in the snubber circuit causes V_{CE} to peak slightly above V_{CC}.

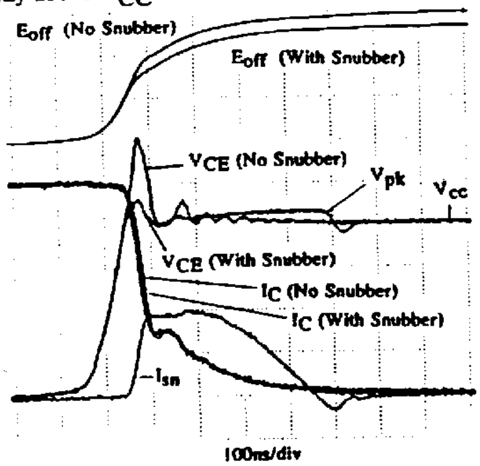


Figure 7. Turn-off waveforms with & without RCD snubber.
Tested at: 400V, 100A, 25°C ; L_s = 100nH,
$V_G/R_{G(off)}$ = -8V/33Ω
C_{sn} = 0.22µF, R_{sn} = 12Ω
V_{CE}: 100V/div, I_C/I_{sn}: 20A/div, E_{off}: 2mJ/div

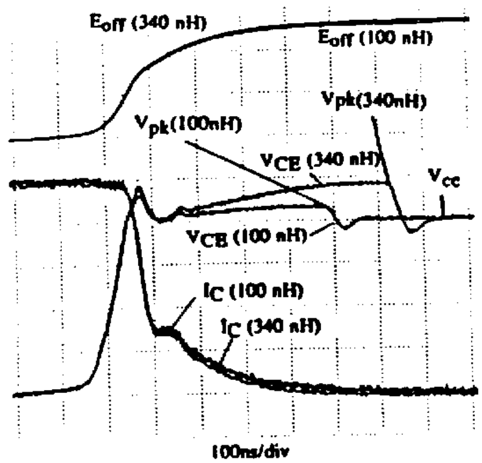


Figure 8. Turn-off waveforms with RCD snubber for two different L_s values (100nH, 340nH).
Tested at: 400V, 100A, 25°C; $V_G/R_{G(off)}$ = -8V/33Ω
C_{sn} = 0.22µF, R_{sn} = 12Ω
V_{CE}: 100V/div, I_C: 20A/div, E_{off}: 2mJ/div

Fig. 8 displays the waveforms generated for two different stray inductances (100nH, 340nH). As illustrated in the figure the initial V_{CE} peak - which is dependent on the stray inductance within the snubber circuitry - is the same for the two cases. The final voltage peak (V_{pk}) for the higher inductance does reach a higher value as expected since there is more trapped energy ($\frac{1}{2} \cdot L_s \cdot I^2$)diverted to the same snubber capacitor. This value however is well within the voltage rating of the device and only marginally influences the losses in the IGBT, since it occurs when the current has reached to a smaller value. The V_{pk} magnitude can be calculated from the formulae given in the following section

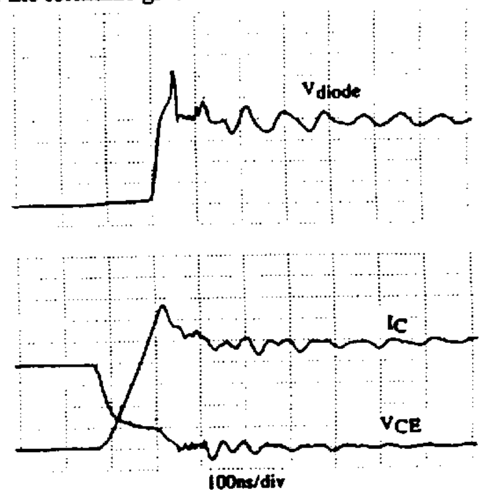


Figure 9. Turn-on waveforms for an IGBT with no RCD snubber protection .
Tested at: 400V, 100A, 25°C ; L_s = 240nH,
$V_G/R_{G\,(on)}$ = 15V/5.1Ω
V_{CE}: 100V/div, I_C: 20A/div, V_{diode}: 100V/div

Turn-on: Fig. 9 displays the turn on waveforms for an unprotected IGBT with a gate resistor (R_G) of 5.1 Ohms. The rapid rise in the IGBT current (1200 A/µs) combined with the circuit stray inductance (300 nH) caused the FWD to go through severe reverse recovery process. As seen in the figure, the FWD recovery voltage (≈630V) actually exceeded the rated voltage of the module.

In order to bring this voltage down to a safe value, the turn on di/dt was lowered by using a higher R_G. The results are shown in Fig. 10. The increase in R_G, however had profound effect in increasing the switching losses, as expected [3], [4].

The RCD clamp shown in Fig. 6a is also effective in reducing turn-on voltage transients. As the IGBT current rises, the $L_s \cdot di/dt$ voltage loss causes the voltage across the positive and negative terminals of the module, V_{ab}, to drop by the same amount (i.e. to V_{CC} - L_s·di/dt). The snubber capacitors that were fully charged to V_{CC}, now find a discharge path through the forward biased Free Wheel Diode (note that the FWD is on, freewheeling the load current), the IGBT and the snubber resistors. Fig. 11 shows the equivalent circuit during turn on. The snubber diodes are reverse biased and therefore not shown. The current paths are shown in the figure. This snubber discharge current (I_{sn}) partially provides for the reverse recovery charge of the FWD, thus the total current seen by L_s is modified. This has a favorable effect on the magnitude of the reverse recovery voltage transient.

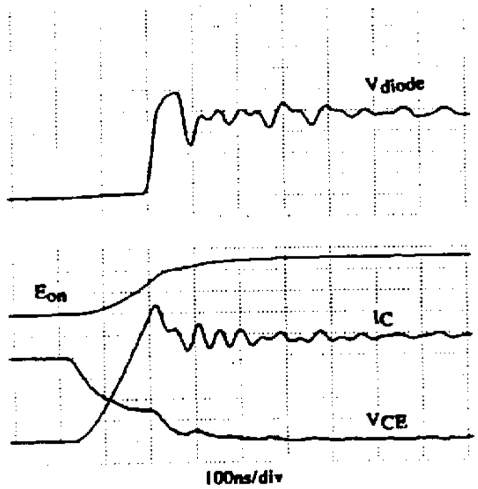


Figure 10. Turn-on waveforms for an IGBT with no RCD snubber protection .
Tested at: 400V, 100A, 25°C ; L_s = 240nH ,
V_G/R_G (on) = 15V/33Ω
V_{CE}: 100V/div, I_C: 20A/div, V_{diode}: 100V/div, E_{on}: 1mJ/div

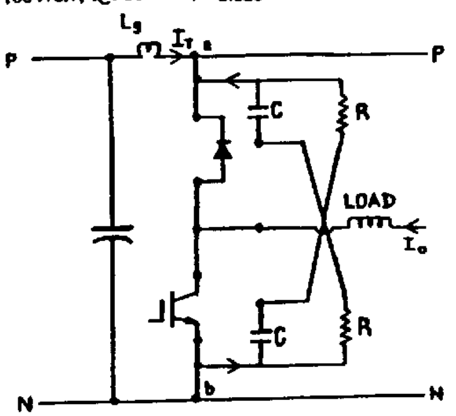


Figure 11. Equivalent circuit under turn-on conditions. Low-side IGBT is switched on.

The waveforms shown in Fig. 12 illustrate the snubber operation. Notice the complete elimination of the voltage transient and also reduction in the oscillations following turn on. Another interesting fact is that this waveform was generated with R_G of 0.5 Ohm which reduced the energy losses from 2.41 mJ in Fig. 10 to 1.25 mJ, a savings of almost 50%! Therefore this snubber, not only clamps the turn on voltage transient, but also enables the user to choose a value of R_G that produces minimal turn on losses.

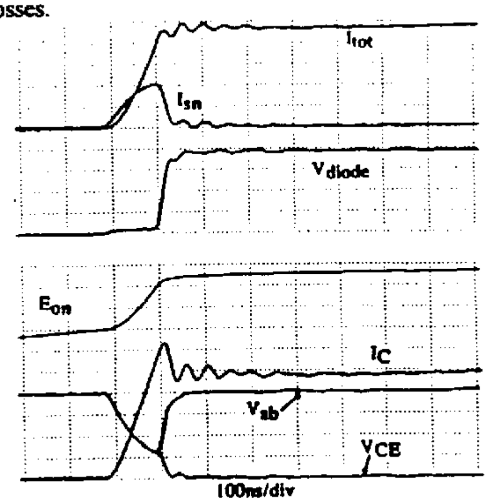


Figure 12. Turn-on waveforms with RCD snubber.
Tested at: 400V, 100A, 25°C ; L_s = 240nH,
$V_G/R_{G(on)}$ = 15V/0.5Ω
C_{sn} = 0.22μF, R_{sn} = 12Ω
V_{CE}/V_{diode}: 100V/div, $I_C/I_{sn}/I_{total}$: 20A/div, E_{on}: 0.5mJ/div
V_{sb}: 100V/Div

Fig. 13 Shows the effect of changing the snubber resistor (R_{sn}) on turn on waveforms. Lower R_{sn}s provide for better snubbing action.

The value for the snubber components can be approximated from the expressions given below, based on circuit stray inductance L_s, switching frequency f_{sw}, maximum switching current I_o, turn-on current rise time t_r, DC rail voltage V_{CC} and allowable peak voltage V_{pk} The derivations are shown in Appendix II.

Snubber capacitor :

$$C_{sn} = L_s \cdot I_o^2 / (V_{pk} - V_{CC})^2 \qquad (1)$$

Snubber resistor:

$$R_{sn} = 1 / (6 \cdot C_{sn} \cdot f_{sw}) \qquad (5)$$

Losses in Snubber resistor :

$$P_R = \tfrac{1}{2} \cdot C_{sn} \cdot (V_{pk}^2 - V_{CC}^2) \cdot f_{sw} \qquad (6)$$

The snubber diode should be of fast and soft recovery type to avoid severe oscillations following V_{pk} at turn off. The resistor should be of non-inductive type to avoid oscillations at turn-on.

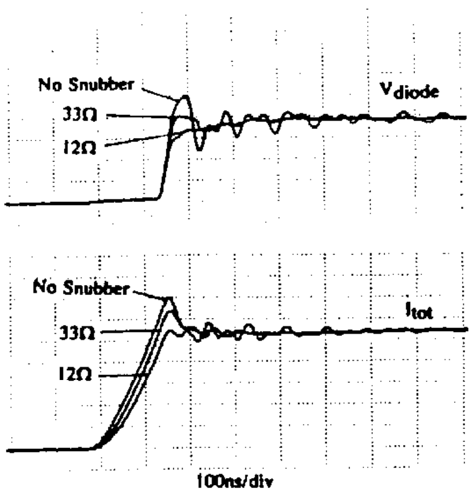


Figure 13. Effect of changing R_{sn} values.
Tested at: 400V, 100A, 25°C; L_s = 240nH,
$V_G/R_{G(on)}$ = 15V/33Ω
C_{sn} = 0.22µF, R_{sn} = 12Ω, 33Ω
V_{diode}: 100V/div, I_{total}: 20A/div

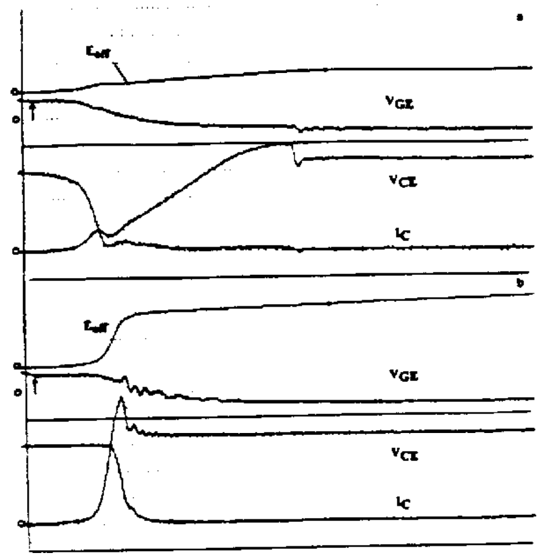
The charge-discharge circuit of figure 6b

The charge-discharge snubber circuit in Fig. 6b can be targeted for reducing turn-off dissipation in IGBT. During IGBT turn on, the snubber capacitor is fully discharged and during turn off, it is fully charged. This circuit, unlike the one in Fig. - 6a which essentially acts as a clamp - reduces the rate of rise of voltage across the IGBT at turn-off, imposing softer switching and thereby reducing losses in the IGBT. The losses in the snubber are:

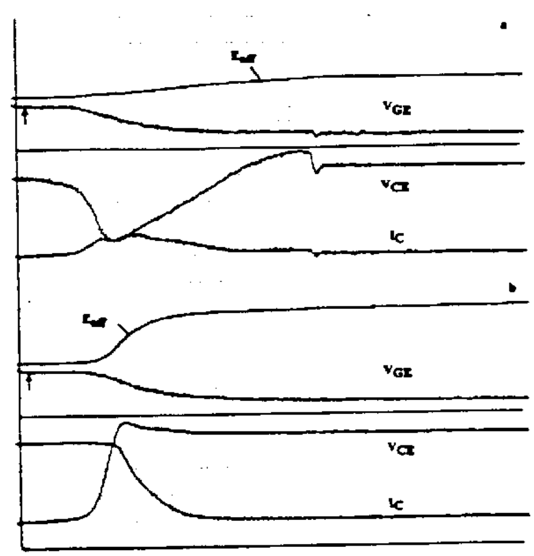
$$P_R = \tfrac{1}{2} \cdot C_{sn} \cdot V_{pk}^2 \cdot f_{sw} \qquad (7)$$

As compared to (6) this losses are substantially higher.

Figs. 14a and 14b show results of testing a 150A, 600V Ultra-Fast IGBT module, with and without a charge-discharge snubber. With a 0.5µF capacitor, the turn-off dv/dt was lowered from 3500V/µs to 300V/µs, as observed from this figures. As a result, turn-off losses were brought down from 11.2mJ to 3.4mJ (down 70%). The test was repeated on a 150A, 600V low V_{CE}(sat) type module. The results are shown in Figs. 15a and 15b. Here again, savings were substantial as the turn-off losses decreased from 20mj to 6.5mJ (down 68%), by due to slowing of the rate of rise of voltage. Low V_{CE}(sat) type of IGBTs due to their longer minority carrier life-time continue to incur higher turn-off losses under soft switching conditions, just as in hard switch conditions.

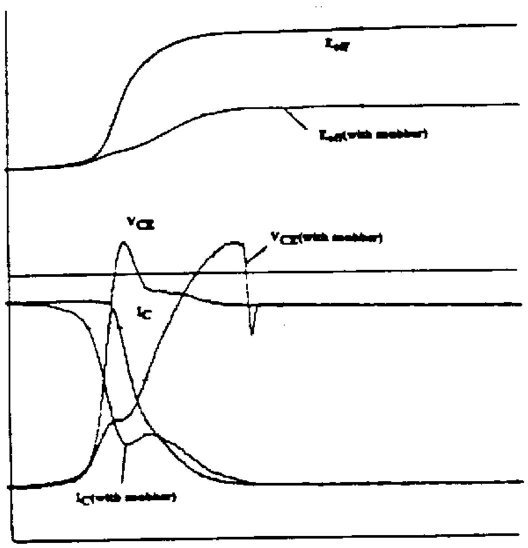


Figures 14a,b. 150A/600V Ultra Fast IGBT turn-off waveforms, With & without Charge-Discharge RCD snubber, respectively.
Tested at 340V, 150A, 125°C, 15V/39Ω, C_{sn}/R_{sn}: 0.5µF/20Ω
V_{CE}: 100V/div, I_C: 50A/div, V_{GE}: 5V/div
E_{off}: 5mJ/div, Time scale: 400ns/div



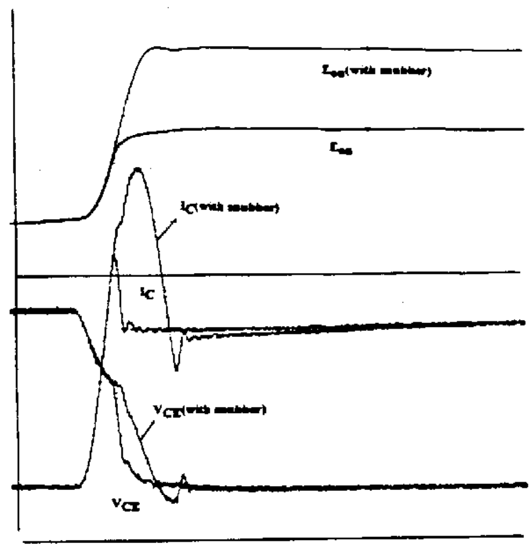
Figures 15a,b. 150A/600V Fast IGBT turn-off waveforms. With & without Charge-Discharge RCD snubber, respectively.
Tested at 340V, 150A, 125°C, 15V/39Ω, C_{sn}/R_{sn}: 0.5µF/20Ω
V_{CE}: 100V/div, I_C: 50A/div, V_{GE}: 5V/div
E_{off}: 5mJ/div, Time scale: 400ns/div

The reduction in turn-off losses directly depends on the load current, I_0. On the other hand, snubber losses depend predominantly on values of C_{sn} and V_{CC}, varying only marginally as the load current is increased (see 7). This is best demonstrated by the test results in Fig. 16 which shows IGBT turn-off at much higher current (350A); once again, with and without snubber. In terms of mJs per switching, the savings here were much more substantial (30mJ) than at 150A (13.5mJ) in Fig. 15a. The snubber losses as given by (7) did not change much. Therefore at higher currents, the savings in IGBT turn-off losses start to overshadow the losses in snubber circuit, making this circuit more applicable. The efficiency is enhanced manifolds if an active feedback circuit is employed to return the trapped energy back to the source.



Figures 16. 150A/600V Fast IGBT turn-off waveforms. With & without Charge-Discharge RCD snubber.
Tested at 340V, 350A, 125°C, 15V/39Ω, C_{sn}/R_{sn}: 0.5μF/33Ω
V_{CE}: 100V/div, I_C: 100A/div, E_{off}: 5mJ/div.

In bridge applications, snubber capacitor across the opposite IGBT could produce extremely high shoot-through currents in the IGBT that is being turned on. The magnitude of this additional current could be as high as $V_{CC} \cdot [C_{sn}/L_s]^{1/2}$, with the width of $\pi[C_{sn} \cdot L_s]^{1/2}$. Fig. 17 shows turn-on waveforms with a 0.1μF snubber capacitor. The snubber inflicted current transients caused turn-on losses at 150A to increase from 9mJ to 16mJ. At impact of this additional current during turn-on becomes less severe as the load current is increased to a higher level. In applications not involving bridge configurations (chopper circuits for an example), this charge-discharge snubber can be used with the considerations of turn-off stresses only. The free-wheel diode can be effectively protected by a simpler RC snubber in such circuits[5].



Figures 17. 150A/600V Fast IGBT turn-on waveforms. With & without Charge-Discharge RCD snubber.
Tested at 340V, 150A, 125°C, 15V/39Ω, C_{sn}/R_{sn}: 0.1μF/33Ω
V_{CE}: 100V/div, I_C: 50A/div, E_{off}: 5mJ/div.

Voltage Transients during Fault Current Turn-off

The short circuit current generated during fault conditions can be up to five to ten times the rated current. Shutting off such high currents too quickly can produce extremely high di/dts that are potentially detrimental to the IGBTs[6].

The snubber circuits discussed in the above section are not as practical when it comes to protecting transient voltages generated during short circuit conditions. As seen from (1), the required snubber capacitor value is proportional to the square of the device current. This means that the capacitor required will have to be 25 to 100 times larger than in normal switching operation. High capacity, high voltage snubber capacitors are large and expensive (as for snubber capacitors required for GTO thyristors!), making the RCD scheme unattractive for high current IGBT modules. Equally importantly, the snubber circuit connected at the terminals do not address the problem of high current module's internal L·di/dt voltage spike.

More practical methods involve slowing down the turnoff of the IGBTs under fault conditions as discussed elsewhere.[1]

Conclusions

The problem of switching voltage transients is an important subject that can not be ignored, especially in the applications using fast switching IGBTs. The paper discussed principal protection circuits. Circuit operations were analyzed and the test results were illustrated. The following table summarizes the results.

Type	Advantages	Disadvantages
Decoupling capacitor	Low snubber losses.	
	Directly and favorably effects turn-off and turn-on voltage stresses.	
	Newly available, module-compatible capacitors are more effective in limiting the voltage transients.	Produces voltage and current oscillations in the DC bus, forcing usage of snubber capacitors with high RMS current limit.
Note: More practical for the lower current range.		
Discharge resistant decoupling capacitor	Low snubber losses.	
	Directly reduces the turn-off voltage overshoots. Also has a favorable effect on the turn-on voltage transients.	Additional eliminates increase snubber inductance, making protection less effective.
	Much quieter switching as the snubber diode blocks off oscillations.	Snappy diode could produce high recovery voltage spikes and dv/dts across IGBT/diode pair.
Note: More practical for the medium current ranges.		
RCD charge-discharge snubber circuit		Very high snubber losses.
	Reduces turn-off voltage overshoots.	Requires more components.
	Could substantially reduce turn-off losses in the transistor.	Increases turn-on losses in the bridge configurations.
	No oscillations in the DC bus.	More complicated component selection.
Note: More Practical for the high current, low bus voltage, chopper applications.		
RCD snubber-clamp circuit	Low snubber losses.	
	Directly reduces the turn-off voltage overshoots. Also has a favorable effect on the turn-on voltage transients.	
	No oscillations in the DC bus.	Requires more components.
Note: Most practical circuit, especially for medium and high current applications.		

To recap, the module-compatible, high-frequency decoupling capacitors were found to offer an optimal solution for the low current applications. The RCD snubber-clamp circuit was found to be the most effective and practical tool for the medium and high current applications.

Appendix I

During short switching interval inductive load current remains unchanged at Io, as the main DC source and the decoupling capacitor supply to it.

i.e. isn + isp = Io during switching. (a)

Where isn is the decoupling capacitor current and isp is the current from the DC source.

Assuming ideal turn-on as the wort case for calculating snubber current, isw(0-)=0, isw(0+)=Io. Where isw is the IGBT current.

Since isp can not change immediately (due to Ls), isn supplies full load current initially. Under ensuing steady-state conditions, isp supplies 100% of the load current.

That is:

isn(0+) = Io, isp(0+) =0 (b)

isn(inf.) = 0, isp(inf.) = Io (c)

The equations governing various circuit variables are:

Vcc=Ls · (disp/dt) + Vsn+ isn · ESR (d)

disn/dt + disp/dt = 0 [from (a)] (e)

isn= - C · dVsn/dt (f)

Assuming that ESR of the high frequency decoupling capacitor constitute most of the circuit stray resistance.

Upon solving (d), (e) & (f) with the initial condition (b) and (c), we get,

$$i_{sn} = e^{-\alpha t} \cdot I_o \cos(\beta t) \qquad (2)$$

At the point of turn off, we have:

isn(0+) = -Io, isp(0+) = Io (g)

isn(inf.) = 0, isp(inf.) = 0 (h)

assuming ideal switching.

The equation remain the same as before and we have,

$$i_{sn} = - e^{-\alpha t} \cdot I_o \cos(\beta t) \quad \text{(note the sign change)}$$

i.e. isn(off) = -isn(on),

The total capacitor RMS current is:

$$Irms = \sqrt{2T/Tsw \quad (1/T) \cdot \int (isn^2)dt}$$

$$= \sqrt{2/Tsw \cdot \int [\, Io * e^{-\alpha t} \cos(\beta t)]^2 dt}$$

$$= Io \sqrt{1/Tsw \cdot \int \{e^{-2\alpha t} (1+\cos(2\beta t))dt\}}$$

$$= Io \sqrt{1/(2\alpha Tsw) \cdot [1+\alpha^2/(\alpha^2+\beta^2)]}$$

where $\alpha = ESR/2L_s$ and $\beta = \sqrt{[4/L_s C_{sn} - (ESR/2L_s)^2]}$

$Irms = Io\sqrt{[fsw \cdot (Ls/R+RC/4)]}$, $fsw = 1/T$,

As Ls/R >> RC/4

$$I_{RMS} = I_o \sqrt{(f_{sw} \cdot L_s / R_s)} \qquad (3)$$

The power loss in the ESR are:

$$Pr = ESR \cdot Irms^2 = fsw \cdot Ls Io^2$$

$$= 2 \cdot (\tfrac{1}{2} Ls \cdot Io^2) \cdot fsw \qquad (4)$$

[$\frac{1}{2} Ls \cdot Io^2$ being the energy stored in L_s]

Appendix II

The expressions (1), (5) and (6) are derived as follows:

At turn-off (see Fig. 6a) as one of the two conducting IGBTs is gated off, collector to emitter voltage $v_{ce}(t)$, rises to the DC bus voltage, V_{CC}. Beyond this point, load current freewheels through the diode across the other IGBT. The stray inductances (L_s) however prolong flow of current in the "DC loop". Two components of currents make up for the current $i(t)$ in L_s. They are IGBT turn-off current ($i_C(t)$) and the snubber current ($i_{sn}(t)$), as marked in Fig. 7.

For simplicity of calculations it is assumed that the IGBT turns off "instantly". i.e. $i_{sn}(t)$ is equal to $i(t)$. This assumption is justified on the grounds that it only renders somewhat conservative estimate of snubber capacitor value.

The equations governing various circuit variables are:

$v_{ce}(t) = V_{CC} + 1/C \int i(t)\, dt$

i.e. $V_{CC} - v_{ce}(t) = -1/C \cdot \int i(t)\, dt$ ——————(a)

$v_{ce}(t) = V_{CC} - L \cdot di(t)/dt$

i.e. $di(t)/dt = [V_{CC} - v_{ce}(t)] / L_s$ ——————(b)

From (a) & (b),

$di(t)/dt = -1/L_sC \cdot \int i(t)\, dt$ ——————(c)

The following is the solution to the above equation:

$i(t) = I_o \cdot \cos(t / \sqrt{L_sC})$ ——————(d)

Where I_o is the load current at the turn-off. The current is thus a cosine function. In Fig. 7, it can be observed that the combination of $i_C(t)$ and $i_{sn}(t)$ does follow cosine wave shape.

Differentiating (d),

$di(t)/dt = -I_o/L_sC \cdot \sin(t / \sqrt{L_sC})$ ——————(e)

From (b) & (e),

$v_{ce}(t) = V_{CC} + I_o \cdot \sqrt{L_s/C} \cdot \sin(t / \sqrt{L_sC})$ ——————(f)

v_{ce} is at maximum when $t / \sqrt{L_sC} = \pi/2$.

Therefore,

V_{CM} = maximum desired voltage across IGBT

$V_{CM} = V_{CC} + I_o \cdot \sqrt{L_s/C}$ ——————(g)

From (g),

Snubber capacitor

$C_{sn} = L_s \cdot I_o^2 / [V_{CM} - V_{CC}]^2$ ——————(1)

The snubber capacitor is therefore charged to V_{CM} at the end of turn-off. Before the next turn-off event, i.e. $1/f_{sw}$ later, C_{sn} should discharge back to its initial value of V_{CC}. The snubber resistor (R_{sn}) selected according to the following Expression fulfills above requirement.

$R_{sn} = 1 / (6 \cdot C_{sn} \cdot f_{sw})$ ——————(5)

The losses in R_{sn} at turn-off are therefore given by:

$P_R(\text{off}) = [\tfrac{1}{2} \cdot C_{sn} \cdot (V_{pk}^2 - V_{CC}^2)] \cdot f_{sw}$ ——————(f)

At turn-on (see Fig. 11 & 12) as an IGBT is turned on, the switching $di(t)/dt$ causes voltage at module terminals $v_{ab}(t)$ to drop from its initial value of V_{CC} to an amount equal to $L_s \cdot di(t)/dt$. As explained in the article, this causes the snubber capacitors to discharge through R_{sn}s.

For simplicity it is assumed that the turn-on di/dt is linear. It is further assumed that the peak turn-on current is 25% above the load current I_o at the switching instant. These assumptions will result in conservative estimate of snubber losses.

The snubber discharge current is:

$i_{sn}(t) = [V_{CC} - V_{ab}] / R_{sn}$ ——————(g)

where $V_{ab} = V_{CC} - L_s \cdot di/dt$ ——————(h)

where di/dt (a constant) = $di/dt = .9\, I_o/t_r$ ——————(i)

t_r is the rise time, specified under inductive load conditions. From (g), (h) & (i),

$i_{sn}(t) = I_{sn} = [L_s \cdot 0.9\, I_o / t_r] / R_{sn}$

$I_{sn} = 0.9 \cdot L_s \cdot I_o / (t_r \cdot R_{sn})$ ——————(j)

The losses in R_{sn} at turn-on are,

$P_R(\text{on}) = [0.9 \cdot L_s \cdot I_o/(t_r \cdot R_{sn})]^2 \cdot R_{sn} \cdot T \cdot f_{sw}$ ——————(k)

where T, the snubber discharge time, is approximated to be interval between the beginning of the current rise and the point where current reaches its peak value ($=1.25 I_o$).

$T = 1.25\, I_o/[.9 I_o/t_r] = 1.39 t_r$ ——————(l)

Combining (k) & (l),

$P_R(\text{on}) = [1.125 \cdot L_s^2 \cdot I_o^2/(t_r \cdot R_{sn})] \cdot f_{sw}$ ——————(m)

From (f) & (m),

$$P_R = [\tfrac{1}{2} \cdot C_{sn} \cdot (V_{pk}^2 - V_{CC}^2) + 1.125 \cdot L_s^2 \cdot I_o^2/(t_r \cdot R_{sn})] \cdot f_{sw}$$

The second half of the above expression is a insignificant part of the total and can be deleted. The total losses in Snubber resistor are therefore,

$P_R = \tfrac{1}{2} \cdot C_{sn} \cdot (V_{pk}^2 - V_{CC}^2) \cdot f_{sw}$ ——————(6)

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