Approximate Loss Formulae for Estimation of IGBT Switching Losses through EMTP-type Simulations

A.D. Rajapakse, *Member, IEEE*, A.M. Gole, *Member, IEEE*, and P. L. Wilson, *Member, IEEE*

***Abstract*-- This paper presents a method for simulating switching and conduction losses of an Insulated Gate Bipolar Transistor (IGBT) device in an electromagnetic transient program (emtp) without recourse to an unreasonably small time-step. A set of equations are derived for calculating switching losses of an IGBT using the device switching characteristics approximated with piece-wise linear functions. These loss equations are integrated to a power electronic switch model of an emtp-type program and used for the simulation of losses in the device. This approach can be used to determine the heat generation of IGBT devices in a large class of Voltage Sourced Converter (VSC) systems.**

Keywords: Insulated gate bipolar transistors, Semiconductor device switching losses, Estimation of switching losses, Simulation of power electronic systems, Pulse width modulated power converters.

I. INTRODUCTION

THE insulated gate bipolar transistor (IGBT) is popularly used in high power, high frequency power-electronic applications such as pulse width modulated (PWM) inverters. These applications require well designed thermal management systems to ensure the protection of IGBT s, which operate with smaller safety margins due to economic considerations. Hence, tools for accurate prediction of device power dissipation and junction temperature become important in achieving optimized designs. At high switching frequencies, switching losses constitute a significant portion of the device power dissipation. Therefore, accurate calculation of switching losses is an important step in the thermal management system design [1].

Most electromagnetic transient programs (emtp-type) model the power electronic devices in a circuit as on-off switches or two state resistances [2]. This simple representation is sufficiently accurate to simulate the system-level electrical behaviour. However, determination of switching losses requires the consideration of the physics of the switching process, which lasts only about several hundreds of nanoseconds. Simulating this process considering the detailed physics as in the number of models described in [3] requires a very small time-step and iterations within a time-step. This could result in an unacceptably large CPU time when simulating large multi-device power electronic systems (such as FACTS devices) using fixed time-step emtp-type simulation programs.

One approach proposed for accurate estimation of power losses is the use pre-defined scalable switching functions obtained through measurements to guide the simulation during switching transients [4]. However, this method still requires the simulation to be carried out with very small time steps.

Another approach that has been proposed for estimation of switching losses is the use of simple functions derived for losses based on the typical switching waveforms [5],[6]. This method was extended in [1] by deriving a set of formulae for switching losses based on the predicted device's current and voltage waveforms. The method differs from the simple use of lookup tables or fitted curves [7],[8] because the predicted waveforms conform to the physics of the switching process and take into account the dependency of the switching losses on various factors such as the switching voltage, switching current, stray inductance and the reverse recovery process of the freewheeling diode.

This paper further refines the approach of [1] and develops a simpler set of loss formulae that uses less number of parameters that need to be extracted from the published data sheets (or test waveforms). Validation of the model was conducted using a simple laboratory set-up and by comparison with published loss curves by the manufacturer. The paper also presents an application example of loss determination in a PWM inverter.

II. DERIVATION OF SWITCHING LOSS FORMULAE

A. Approach Used for Modelling of Device Losses

The losses in a power-switching device constitute of conduction losses, off-state blocking losses, turn-on switching losses, and turn-off switching losses. The conduction loss is calculated in a straightforward manner as the product of the device current and the forward saturation voltage; and the blocking loss is the product of the blocking voltage and the leakage current [1].

The device switching losses depend on the application circuit. Thus, a circuit configuration of a voltage source

This work was supported by the Manitoba HVDC Research Centre and the Natural Sciences and Engineering Research Council (NSERC), Canada.

A.D. Rajapakse and A. M. Gole are with the Department of Electrical and Computer Engineering, University of Manitoba, Winnipeg, MB, Canada, R3T 5V6 (email: athula @ee.umanitoba.ca, gole @ee.umanitoba.ca)

P. L. Wilson is with the Manitoba HVDC Research Center, Winnipeg, MB, Canada, R3J 3W1 (email: plwilson@hvdc.ca)

Presented at the International Conference on Power Systems Transients (IPST'05) in Montreal, Canada on June 19-23, 2005
Paper No. IPST05 - 184

converter, which is the most common application of IGBT, is considered. In this configuration, each switch consists of an IGBT in anti-parallel with a freewheeling diode. It is assumed that the switching devices are subject to 'hard switching' and the load is inductive. The turn-on of the IGBTs is significantly affected by the reverse recovery behaviour of the freewheeling diode and the parasitic inductances. A test circuit of Fig 1 closely represents one leg of a voltage source converter.

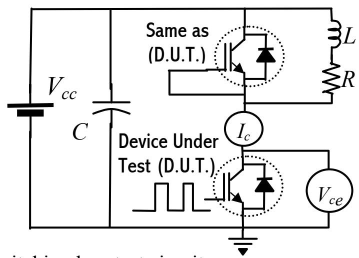


Fig. 1. Inductive switching loss test circuit

Figure 2 shows measured current and voltage waveforms of a hard switched IGBT device during the (a) turn-on and (b) turn-off transients. Switching process completes within a few hundred nanoseconds, and therefore, simulation of switching losses is difficult without resorting to extremely small simulation time steps. The type of simulation programs considered in this paper, that is emtp-type programs, uses a much larger time-step (tens of microseconds) in simulating large systems.

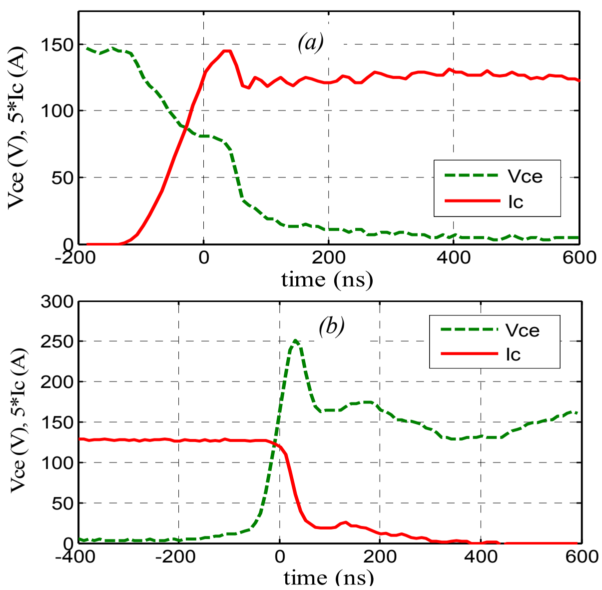


Fig 2. Measured IGBT (a) turn-on and (b) turn-off waveforms.

A method to meet these apparently contradictory requirements was proposed in [1] by developing algebraic equations that represent the voltage and current waveforms during the switching event. These algebraic equations are based on the fact that current and voltage waveforms during the switching are principally a function of the pre- and post-switching voltages and currents [3]. A formula for switching energy can be obtained by integrating the product of the above voltage and current equations that 'fill-in' the intermediate sub-microsecond values of voltage and current within the larger emtp time-step of several microseconds. Thus the emtp-type simulation can be conducted with a larger time-step, with a formula providing an estimate of the loss after each switching event.

The loss equations proposed in [1] are complex and involve several parameters that need to be found using test waveforms. In the absence of test waveforms, these parameters must be simultaneously tuned using the data sheet specified switching losses at the rated conditions. This may be a difficult task for an inexperienced user. Therefore, more simplified set of loss formulae is presented in this paper. Moreover, the number of tuning parameters involved with each formula is restricted to one. This is achieved by following the general modeling approach of [1], but using piece-wise linear waveforms to approximate the variations of the current and voltage waveforms during the switching events.

B. Diode Turnoff Loss Formula

In modern fast recovery diodes used with IGBTs, the turn-on loss is negligible (less than 1%) compared to the turn-off loss. However, the reverse recovery during the turn-off causes appreciable amount of losses in the diode and also affects the turn-on losses of the incoming IGBT.

Idealized approximation of diode turn-off waveforms are shown in Fig. 3. In a free-wheeling diode, the initial rate of fall of the diode current dI_d/dt, is determined by the turn-on speed of the IGBT in the opposite leg. During the initial phase of reverse recovery, negative current through the diode increases and reaches its peak value, I_{rrm}. At this point, the diode starts to rapidly gain the reverse voltage.

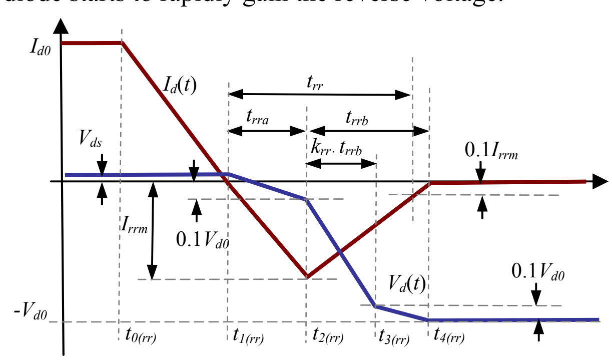


Fig. 3. Approximated waveforms of diode turn-off transient

The waveforms can be characterized using the parameters I_{rrm} (peak reverse recovery current), and t_{rr} (the reverse recovery time). In the interval $[t_{0(rr)}, t_{2(rr)}]$, I_d is linearly decreasing, with $V_d = V_{ds}$ (on state forward voltage drop). Knowing the initial dI_d/dt , (which depends on the IGBT turn-on rise time), the parameters t_{rra}, and t_{rrb} are estimated as:

$$t_{rra} = I_{rrm} \Big/ \left(\left. \frac{dI_d}{dt} \right|_{t_{1(rr)}} \right) \tag{1}$$

$$t_{rrb} = 1.11(t_{rr} - t_{rra}) \tag{2}$$

It is also assumed that diode voltage drops to 90% of its reverse blocking voltage during $[t_{2(rr)}, t_{3(rr)}] = k_{rr}.t_{rrb}$.

An energy loss occurs in the diode during the reverse recovery, particularly during building up of reverse voltage. This can now be computed as:

$$W_{rec} = \int_{t_{0(rr)}}^{t_{4(rr)}} V_d(t) I_d(t) dt$$

$$W_{rec} = 0.5 V_{ds} I_{d0} \frac{I_{d0}}{\left(\left.\frac{dI_d}{dt}\right|_{t_{1(rr)}}\right)} + 0.033 V_{d0} I_{rrm} t_{rra} + V_{d0} I_{rrm} (0.467 - 0.433 k_{rr} + 0.15 k_{rr}^2) t_{rrb} \quad . \quad (3)$$

The parameter k_{rr} [0.0-1.0] can be determined by substituting data sheet value of W_{rec} and the corresponding measuring conditions to (3).

C. IGBT Turn-on Loss Formula

Fig. 4 shows approximate waveforms for the hard turn-on transient of an IGBT. In data sheets, the turn-on behaviour is characterized by the turn on delay time, $t_{d(on)}$, the rise time, t_r, and the turn-on energy, W_{on}.

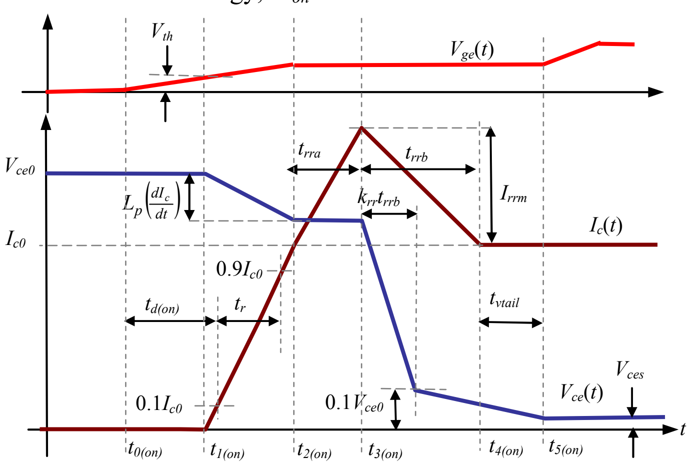


Fig. 4. Hard switching turn-on transient waveforms for loss calculation

The turn-on gate pulse applied at $t_{0(on)}$ raises the gate voltage V_{ge} gradually, with a rate of rise determined by the input capacitance of the IGBT and gate drive resistance. After a time $t_{d(on)}$, when V_{ge} reaches a threshold voltage V_{th}, the collector current, I_c, starts to rapidly rise and the load current in the freewheeling diode (of the opposite leg) gradually transfers to the IGBT. The rate of rise of collector current can be determined using the definition of rise time:

$$\frac{dI_c}{dt} = \frac{0.8 I_{c0}}{t_r} \quad (4)$$

During this current rise, the device (collector-emitter) voltage $V_{ce}(t)$ experiences a drop due to parasitic inductance (L_p). The voltage drop due to parasitic inductance is not instantaneous due to parasitic capacitance effects and the nonlinearities in the IGBT transconductance. It is assumed that the voltage V_{ce} drops linearly over the period [$t_{1(on)}$,$t_{2(on)}$] and reaches a plateau in the voltage waveform. This plateau voltage, V'_{ce}, can be found as

$$V'_{ce0} = V_{ce0} - L_p \frac{dI_c}{dt} = V_{ce0} - \frac{0.8 I_{c0}}{t_r} L_p \,. \quad (5)$$

When the load current is fully transferred to the IGBT at $t_{2(on)}$, the outgoing freewheeling diode starts to turn off and forces its reverse recovery current through the IGBT. This current appears as an overshoot in I_c during [$t_{2(on)}$,$t_{4(on)}$]. When the diode reverse recovery current reaches its peak value at $t_{3(on)}$, the collector-emitter voltage V_{ce} begins to fall (with the diode gaining its reverse voltage). The duration, during which the reverse recovery current increases, can be computed by substituting (4) into (1):

$$t_{rra} = \frac{I_{rrm}}{0.8 I_c} t_r \quad (6)$$

The time taken for current to fallback to on state value after the peak is equal to t_{rrb} defined in (2). It is assumed that collector-emitter voltage rapidly falls to 10% of its off-state value within a time of $k_{rr}.t_{rrb}$. The remaining voltage gradually drops and reaches the on-state saturation voltage, V_{ces} at time $t_{5(on)}$. The turn on energy loss can be computed as

$$W_{on} = \int_{t_{0(on)}}^{t_{5(on)}} V_{ce}(t) I_c(t) dt$$

$$\begin{aligned} W_{on} = {} & 0.05 V_{ce0} I_{c0} t_{d(on)} + 0.225 V_{ce0} I_{c0} t_r \\ & + V'_{ce0} I_{c0} \left[0.394 t_r + t_{rra} + 0.55 k_{rr} t_{rrb}\right] \\ & + 0.5 V'_{ce0} I_{rrm} t_{rra} + 0.5 V_{ces} I_{c0} t_{vtail} \\ & + V'_{ce0} I_{rrm} (0.033 + 0.483 k_{rr} - 0.167 k_{rr}^2) t_{rrb} \\ & + V'_{ce0} I_{c0} k_{vtail} [0.5(1-k_{rr}) t_{rrb} + 0.05 t_{vtail}] \\ & + 0.167 V'_{ce0} I_{rrm} k_{vtail} (1-k_{rr}) t_{rrb} \end{aligned} \quad (7)$$

where

$$k_{vtail} = \frac{t_{vtail}}{t_{vtail} + (1-k_{rr}) t_{rrb}} \,. \quad (8)$$

D. IGBT Turn-off Loss Formula

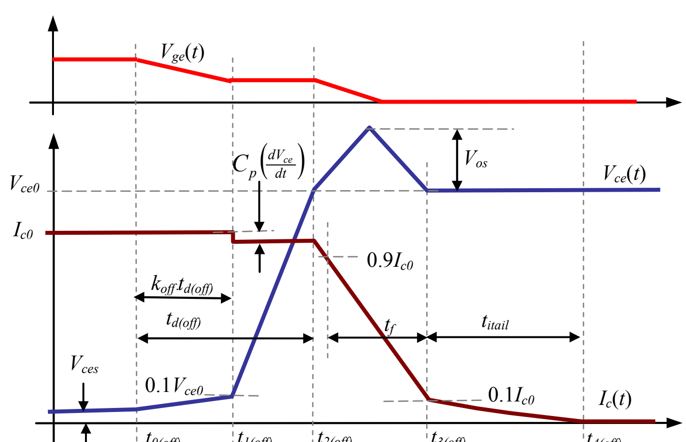


Fig. 5. Approximated hard switching turn-off transient waveforms

The IGBT's turn-off behaviour shown in Fig. 5 is characterized in data sheets by the turn-off delay time, $t_{d(off)}$, fall time, t_f, and turn-off energy, W_{off}. The turn-off process starts on the application of negative gate voltage at time $t_{0(off)}$. The input capacitance of the IGBT discharges gradually reducing gate-emitter voltage, V_{ge}, but the collector-emitter voltage, V_{ce}, remains essentially unchanged until V_{ge} drops sufficiently to drive the IGBT out of saturation. This initial period $[t_{0(off)}, t_{1(off)}]$ is assumed to be equal to $k_{off} \cdot t_{d(off)}$ as indicated in Fig. 5. Thereafter, the collector-emitter voltage rises rapidly at a rate of

$$\frac{dV_{ce}}{dt} = \frac{0.9V_{ce0}}{(1-k_{off})t_{d(off)}} \quad . \tag{9}$$

There could be a drop in the collector current due to the rate of rise of collector voltage, if the parasitic capacitance is significant. Considering the idealized waveforms, the resulting plateau in the collector current, I'_{c0}, can be approximately determined as

$$I'_{c0} = I_{c0} - C_p \frac{dV_{ce}}{dt} = 1 - \frac{0.9V_{ce0}}{(1-k_{off})t_{d(off)}} C_p \tag{10}$$

When V_{ce} reaches the forward blocking voltage V_{ce0}, at $t_{2(off)}$, the freewheeling diode become forward biased and starts to take over the load current. The IGBTs internal construction includes a MOSFET driving a bipolar transistor, and due to the mechanisms involved in these devices, the collector current I_c initially has a rapid fall; followed by a more gentle drop towards extinction at time $t_{4(off)}$. The rapid drop in current through the parasitic inductance produces an overshoot in the voltage V_{ce}. The peak overshoot voltage is determined from

$$V_{os} = L_p \frac{dI_c}{dt} = \frac{0.8I'_{c0}}{t_f} L_p \tag{11}$$

It is assumed that the peak of the overshoot occurs at the midpoint of rapid current fall. The tail current, which is assumed as 10% of the on state collector current, dies down after a period of t_{itail} at $t_{4(off)}$.

Based on the idealized waveform, an approximate expression for the turn-off energy can be developed as

$$W_{off} = \int_{t_{0(off)}}^{t_{4(off)}} V_{ce}(t) I_c(t) dt$$

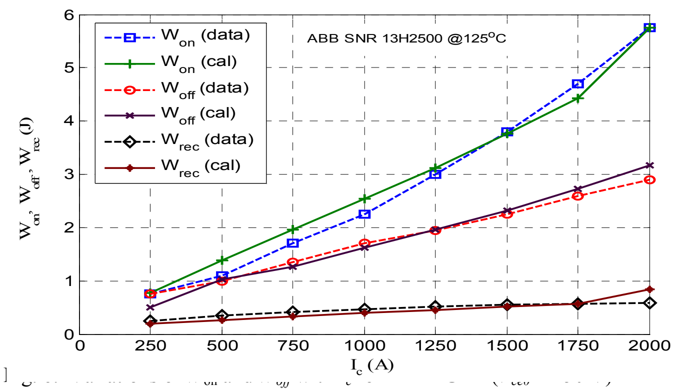
$$\begin{aligned} W_{off} &= 0.5V_{ces} I_{c0} k_{off} t_{d(off)} + 0.05V_{ce0} I_{c0} k_{off} t_{d(off)} \\ &+ 0.275V_{os} I'_{c0} t_f + 0.55V_{ce0} I'_{c0} (1-k_{off}) t_{d(off)} \\ &+ V_{ce0} I'_{c0} \left[0.55t_f + 0.05t_{tail}\right] \end{aligned} \tag{12}$$

E. Validation of Switching Energy Models

The approach developed above was validated by comparison with published results from manufactures' data sheets as well as with a laboratory setup of the circuit in Fig. 1.

1) Comparison with Manufactures Data Sheets

The possibility of using the derived expression for switching energies with data sheet parameters was studied by computing switching energy versus collector current curves for several IGBTs. The parameters that are not available in the data sheets were estimated by adjusting their values to match with the switching energy at the rated current. Those parameter values are then used to predict the turn-on and turn-off losses at different collector currents. Two commercially available IGBTs (with anti-parallel diodes) from different manufacturers were considered, the SNR13H2500 from ABB, rated at 2.5 kV, 1300 A and the IXGK50N60BD1 from IXYS rated at 600 V, 50 A. The variation of calculated reverse recovery, turn-on and turn-off energy losses (as in(3), (7) and (12)) for these two devices are shown in Figs. 6 and 7 respectively superimposed on the loss curves from the manufacturers' data sheets. As can be seen, the equations developed in this paper capture with reasonable accuracy, the variation of losses with current.



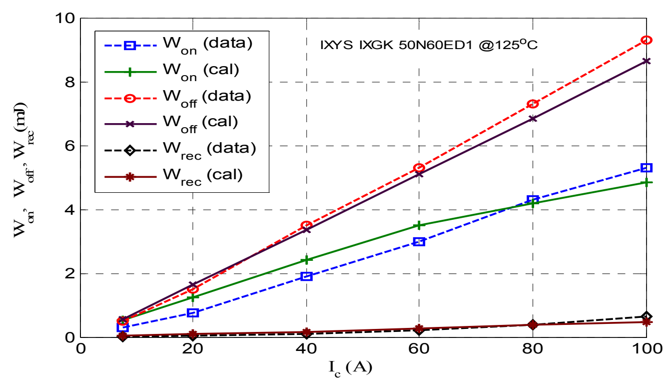


Fig 7 Variations of E_{on} and E_{off} with I_c for IXYS IGBT (V_{ce0}=480 V)

2) Experimental Confirmation

An IGBT-diode package rated at 600 V, 25 A (International Rectifier IRG4PC40KD) was used for the experiment. The test circuit as in Fig. 1 was constructed with the parameters as used in [1] (L=20 mH, C= 4500 μF). The on-state current in the device was adjusted by selecting the resistance R appropriately.

The current and voltage waveforms for IGBT turn-on and turn-off transients were recorded using a high speed oscilloscope. The tests were repeated at various combinations of several different voltages (150 V and 120 V) and currents (15 A, 20A and 25 A). The experimental energy losses were obtained by integrating the product of the measured voltage and current during switching. Table 1 compares these experimental energy losses with the corresponding calculated energy losses from (7) and (12). It also gives the energy losses calculated using the formulae proposed in [1]. It can be seen that the new formulae proposed in this paper are more accurate with the maximum error less than ±10% for all cases.

The number of non-standard parameters (which are not provided in manufacturers datasheets) necessary for the loss equations (k_{rr} in (3), t_{vtail} in (7) and k_{off} and t_{itail} in (12)) are less than number of such parameters used in the equations given in [1]. These parameters can be readily determined if test waveforms are available or otherwise determined approximately as explained above, using the switching energy values given at the rated conditions.

TABLE I: COMPARISON OF MEASURED AND CALCULATED POWER LOSSES FOR IRG4PC40KD IGBT

V_{ce0} (V)	I_{c0} (A)	W_{on} (mJ) Meas	*Cal[1]	Error (%)	Cal	Error (%)
150	15	0.193	0.216	11.9	0.194	6.4
150	20	0.274	0.312	13.9	0.279	1.7
150	25	0.353	0.417	18.1	0.375	0.6
120	15	0.155	0.173	11.6	0.155	0.0
120	20	0.221	0.242	9.5	0.217	-1.7
120	25	0.310	0.325	4.8	0.295	-4.8
V_{ce0} (V)	I_{c0} (A)	W_{off} (mJ) Meas	*Cal[1]	Error (%)	Cal	Error (%)
150	15	0.183	0.203	10.9	0.177	-3.3
150	20	0.289	0.318	10.0	0.287	-0.5
150	25	0.394	0.478	21.3	0.425	7.9
120	15	0.112	0.125	11.6	0.120	7.0
120	20	0.183	0.205	12.0	0.184	0.4
120	25	0.289	0.321	11.1	0.285	-1.4

*Cal[1]: Calculated using the loss formulae given in [1]

III. REPRESENTATION OF SWITCHING LOSSES IN EMTP-TYPE SIMULATIONS

The general switch model of the emtp-type program considered (PSCAD/EMTDC^{TM}) consists of an ideal switch in series with a resistance whose value dependent on the switch state and a constant dc source representing the forward voltage drop. A loss estimation module is integrated to the existing switch model as an additional calculation layer. Thus its traditional structure is retained in the network simulation. The losses in the device (including conduction, blocking and switching losses) are estimated at each time step after the network solution by observation of the pre- and post-switching currents and voltages.

Several manufacturers provide information on the temperature dependence of certain switching parameters (i.e., V_{ces} and switching times) and this information is also represented in the developed model. Once the losses in the device are known, its internal temperature can be found using a dynamic model of the thermal path, such as the model proposed in [1]. The computed device temperature is then used to change the parameters of the switch loss model for the next time-step.

Because the losses are analytically computed separately in the loss calculation module they do not appear as losses in the electrical circuit simulation. Since the network simulation time step is much larger than the switching times, it is impossible to represent instantaneous power losses due to switching in the network. However, the average power losses due to switching can be represented in the network simulation by slightly modifying the simple switch model as shown in Fig. 8. The values of the voltage source representing the forward voltage drop and the current source representing the leakage current can be appropriately changed so that the average power loss due to switching is 'felt' in the network.

The accumulated switching energy loss in the device at time t , $W(t)$, is computed at each time step as

$$W(t) = [W(t-\Delta t)e^{-\Delta t/T_W} + W_{on}(t) + W_{off}(t)] \tag{13}$$

assuming that the accumulated energy decays exponentially with a time constant of T_W. Δt is the simulation time-step. The decaying energy is reflected in the electric network as power dissipation in the device. In order to achieve this, if the IGBT is in conduction, its forward voltage drop is adjusted according to

$$V_{ces} = \frac{W_{cond}(t) + W(t)(1-e^{-\Delta t/T_W})}{\Delta t \cdot I_c} \ , \tag{14}$$

and if the IGBT is off, its leakage current is adjusted so that

$$I_{cs} = \frac{W_{blok}(t) + W(t)(1-e^{-\Delta t/T_W})}{\Delta t \cdot V_{ce}} . \tag{15}$$

A similar approach is adapted for the diode too.

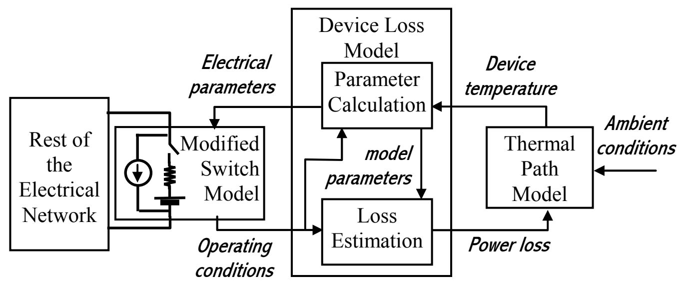


Fig. 8. Interface between the device model and the network model

IV. APPLICATIONS

A. Example: VSC with Hysteresis Current Control

The voltage source converter (VSC) example given in [1] was simulated with the loss estimation methods developed above. The inverter shown in Fig 9 uses six IGBT/Diode modules (parameters used in the model corresponds to IXER 35N120D1 IGBT/diode module). The converter is controlled using hysteresis current control, which maintains the current in any phase within a specified tolerance band around the reference current setting. Unlike conventional sinusoidal PWM, the switching events and their frequency are highly load dependent and so an a-priori estimation of switching losses is not possible.

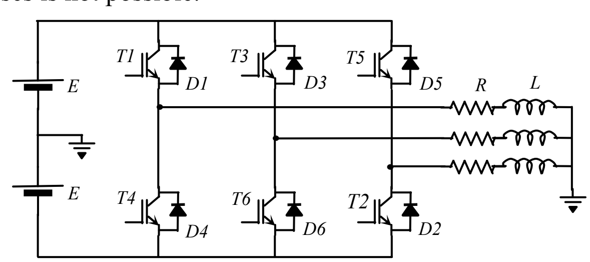


Fig. 9. Hysteresis controlled voltage source inverter

Figure 10 shows the load (L=5mH, R=5Ω) currents in three phases when the hysteresis band is set to 5A. The currents and voltages across the IGBT T1 and diode D1 are shown in Fig. 11. Figure 12 shows the total power losses in IGBT T1 and diode D1 as functions of time. The curves P_{T1cal} and P_{D1cal} show the instantaneous power losses calculated using the methods described in Section II. In IGBT, the conduction loss is proportional to the magnitude of the current. The spikes at the beginning and end of conduction periods correspond to turn-on and turn-off losses respectively. In the diode, reverse recovery energy losses appear as spikes at the end of conduction periods. The curves denoted as P_{T1meas} and P_{D1meas} show the respective device power dissipations as measured in the network. Due to the averaging, the peaks in the power dissipation (due to switching) are smoothen and distributed over a period of time. The time constant T_W should be selected so that it facilitates energy dissipation without buildup in the accumulated energy $W(t)$. However, a too small time constant can cause overly large values of V_{ces} and I_{cs}.

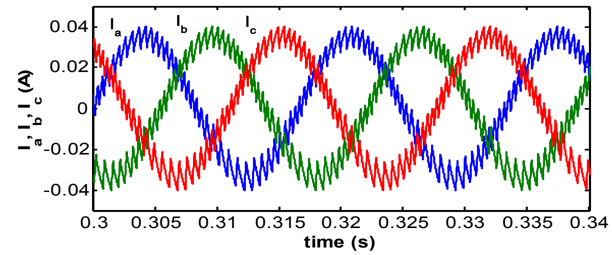


Fig. 10. Load currents

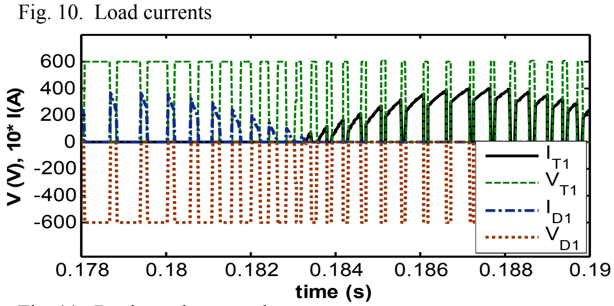


Fig. 11. Device voltages and currents

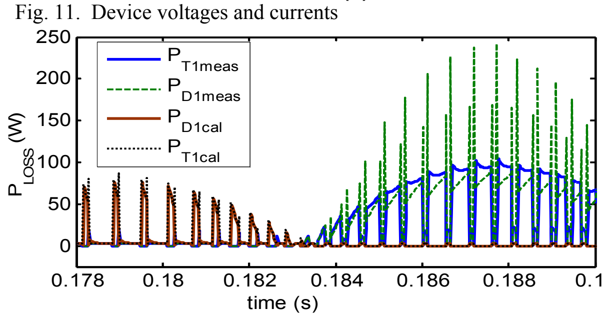


Fig. 12. Power losses in T1 and D1

The estimated total losses (using loss formulae) in the IGBT T1 is 24.39W (conduction=17.62W, turn-on=3.12W, turn-off=2.43W, and blocking=1.23W) and that of diode D1 is 7.10W (conduction= 5.63W, turn-off=1.20W, and blocking=0.26W). The average power dissipations measured in the network are: IGBT T1=23.26 W and diode D1=5.67W. The above results show that the device dissipations are fairly closely reflected in the network simulation. This example also shows that the presented method can be used for analyzing the contribution of different type losses to device heating and determining the device losses under different loads, different control settings and with different device types.

As indicated in Sections II*C* and II*D*, parasitic inductances have significant effect on the switching losses and thus they are included in the proposed loss equations. Determination of the values of these components is discussed in [9] and [10].

V. Conclusions

This paper presented an improvement to a method developed for estimating switching losses of an IGBT subjected to hard switching conditions without resorting to very small time step simulations. This is achieved by deriving a set of formulae for approximately estimating the turn-on and turn-off switching losses using predicted trajectories of the device current and voltage variations between their respective pre- and post-switching values.

According to test results, the formulae presented in this are more accurate than those of [1] and at the same time, reduced the number of non-standard parameters to be extracted from manufacturer's data sheets. The method suggested for representing the average effect of switching losses in the network works fairly well as demonstrated in the example of a PWM inverter simulation.

VI. References

[1] A.D. Rajapakse, A.M. Gole and P.L. Wilson, "Electromagnetic transient simulation models for accurate representation of switching losses and thermal performance in power electronic systems," *IEEE Trans. Power Delivery*, vol. 20, No.1, pp. 319-327, Jan. 2005.

[2] H. Selhi, and C. Christopoulos, "Generalized TLM switch model for power electronics applications," *IEE Proc. Scientific Measuring Technology*, vol. 145, no.3, May 1998.

[3] K. Sheng, B.W. Williams and S.J. Finney,"A review of IGBT models,"*IEEE Trans. Power Electronics*, vol.15, no. 6, Nov. 2000, pp 1250-1266.

[4] C. Wong, "EMTP modeling of IGBT dynamic performance for power dissipation estimation," *IEEE Trans. Industry Applications*, vol. 33, pp 64 –71, Jan.-Feb. 1997.

[5] J. Qian, A. Khan, A. I. Batarseh, " Turn-off switching loss model and analysis of IGBT under different switching operation modes," in *Proc. 21st International Conference on Industrial Electronics, Control, and Instrumentation*, 6-10 Nov. 1995, vol.1, pp. 240 –245.

[6] M. Bland, P. Wheeler, J. Clare and L. Empringham, "Comparison of calculated and measured losses in direct AC-AC converters," in *Proc. IEEE 32nd Annual Power Electronics Specialists Conference*, 17-21 June 2001, vol.2, pp. 1096 –1101.

[7] S. Azuma, M. Kimata, M. Seto, X. Jiang, H. Lu, D. Xu, and L. Huang, "Research on the power loss and junction temperature of power semiconductor devices for inverter," in *Proc. of the IEEE International Vehicle Electronics Conference*, 6-9 Sept. 1999, vol.1, pp. 183 -187

[8] F. Blaabjerg, U. Jaeger, S. Munk-Nielsen, and J.K. Pedersen, "Comparison of NPT and PT IGBT-devices for hard switching applications," in *Conference Record of the 1994 IEEE Industry Applications Society Annual Meeting*, vol.2, pp. 1174 –1181.

[9] M.C. Caponet, F. Profumo, R.W. De Doncker, and A. Tenconi, "Low stray inductance bus bar design and construction for good EMC performance in power electronic circuits", *IEEE Trans. Power Electronics*, vol.17 , no. 2, Mar. 2002, pp. 225 – 231.

[10] K. Xing, F.C. Lee, and D. Boroyevich, "Extraction of parasitics within wire-bond IGBT modules," *Proc. of Applied Power Electronics Conference and Exposition*, 15-19 Feb. 1998, pp 497 – 503.