INTERNATIONAL JOURNAL OF NUMERICAL MODELLING: ELECTRONIC NETWORKS, DEVICES AND FIELDS
Int. J. Numer. Model. 2006; **19**:33–46
Published online in Wiley InterScience (www.interscience.wiley.com). DOI: 10.1002/jnm.597

A fast power loss calculation method for long real time thermal simulation of IGBT modules for a three-phase inverter system

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SUMMARY

A fast power losses calculation method for long real time thermal simulation of IGBT module for a three-phase inverter system is presented in this paper. The speed-up is obtained by simplifying the representation of the three-phase inverter at the system modelling stage. This allows the inverter system to be simulated predicting the effective voltages and currents whilst using large time-step. An average power losses is calculated during each clock period, using a pre-defined look-up table, which stores the switching and on-state losses generated by either direct measurement or automatically based upon compact models for the semiconductor devices. This simulation methodology brings together accurate models of the electrical systems performance, state of the art-device compact models and a realistic simulation of the thermal performance in a usable period of CPU time and is suitable for a long real time thermal simulation of inverter power devices with arbitrary load. Thermal simulation results show that with the same IGBT characteristics applied, the proposed model can give the almost same thermal performance compared to the full physically based device modelling approach. Copyright © 2006 John Wiley & Sons, Ltd.

KEY WORDS: IGBT; voltage source inverters (VSI); thermal design; simulation; modelling

1. INTRODUCTION

The high power density found in modern IGBT power modules subject the power devices to high thermal stress, especially in the very harsh electrical environments found in hybrid vehicles [1]. Since the reliability of power modules depends strongly on the device junction operating temperature, effective power losses and thermal simulation techniques are required to aid in the design of these modules.

To determine the losses, an accurate approach is to simulate a circuit by using physically based device model [2–4], however, due to the complicated physical switching process, and the

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Contract/grant sponsor: Toyota Motor Corporation

Received 1 September 2005
Revised 1 October 2005
Accepted 1 October 2005

very fast transient period, the determination of power losses requires very small simulation time steps (in the order of nanoseconds), this results in an unacceptably large CPU times and memory storage; for a multi-device power electronics systems the problem is even more intractable. At the other end of the scale it is common to use device data-sheets, provided by manufacturers, to estimate losses [5]. This method is simple, but can be very inaccurate.

It is clear that a method is needed which is fast and yet accurate at the system level. In this paper a fast power losses calculation method suitable for long real-time thermal simulations of a three phase IGBT module is presented.

The approach taken consists of two stages. Initially the conducting and switching losses are calculated, for both the IGBTs and diodes, over a range of temperatures and a number of operating currents. These can either be measured or predicted using advanced compact models, but must be produced using the typical operating environment of the inverter load and parasitic elements, which includes a fixed dc-link voltage. These results are then stored in a look-up table which effectively represents the pre-characterized device behaviour.

Next, a system level inverter electrical simulation is carried out, in which, the inverter is very much simplified, so that the simulation can be run with a large simulation time-step which lead to a fast simulation speed. The electrical simulation gives the required electrical characteristics such as the effective voltage and current for power loss calculation. The PWM related signals can then be derived from the effective voltage and current waveforms. The average power loss over each simulation time step has been mathematically calculated using the defined device characteristics in the look-up table. By feeding this average power loss into a compact thermal network, the junction temperatures of devices are calculated. This method is suitable for a long real time thermal simulation for inverter power devices with arbitrary load operation. Throughout this paper, the simulation time step is set equal to the PWM switching period (1 ms).

2. EXPRESSION OF IGBT CHARACTERISTICS FOR POWER LOSSES CALCULATION

The power loss of IGBT devices can be divided into conduction losses and switching losses (turn-on and turn-off loss), as illustrated by a typical inductive load turn-on and turn-off process as shown in Figure 1. The corresponding power losses waveform is shown in Figure 2, and the areas filled with light black denote the energy losses.

The conduction energy loss in the switching cycle can be calculated by integrating the power over the on-time period, which can be expressed as

$$E_{\mathrm{cond\_loss}}(k) = \int_{T_{\mathrm{on}}(k)} V_{\mathrm{ce}}(t)\, i_{\mathrm{c}}(t)\, \mathrm{d}t \tag{1}$$

where, V_{ce}, i_{c}, represent the forward saturation voltage and collector current, respectively, k represents the kth switching cycle, and $T_{\mathrm{on}}(k)$ is the on-time period of devices for the kth switching cycle. The relationship between V_{ce} and i_{c} for a given dc link voltage is given by

$$V_{\mathrm{ce}}(t) = f_{\mathrm{cont}}(i_{\mathrm{c}}(t), T_j(t)) \tag{2}$$

where T_j represents the junction operating temperature.

Switching energy losses can be calculated by integrating the product of device transient voltage and current over the transient period. A key assumption needs to be made at this point,

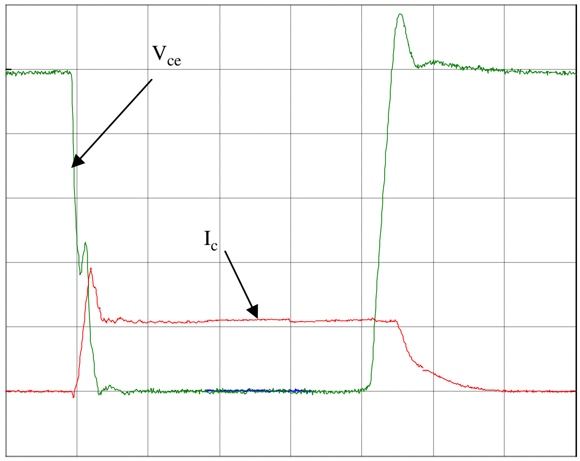

Figure 1. Switching current and voltage.

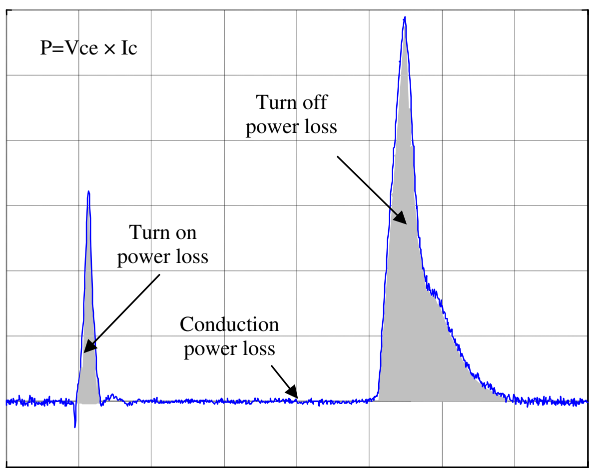

Figure 2. Power and switching energy losses.

which is that the time constant of the inductive load is very large compared with the time constant of PWM waveforms. This means that the current through the load is virtually constant over one switching cycle. This key assumption means that the switching losses are a function of the current and temperature alone. Therefore, we can define the switching losses as follows:

$$E_{\text{sw-on}}(k) = f_{\text{sw-on}}(i_c(k), T_j(k)) \tag{3}$$

$$E_{\text{sw-off}}(k) = f_{\text{sw-off}}(i_c(k), T_j(k)) \tag{4}$$

and the total switching loss of each switching cycle is then calculated by

$$E_{\text{sw}}(k) = E_{\text{sw-on}}(k) + E_{\text{sw-off}}(k) \tag{5}$$

and represented by

$$E_{\mathrm{sw}}(k) = f_{\mathrm{sw}}(i_{\mathrm{c}}(k), T_j(k)) \tag{6}$$

To speed-up the simulation, two 2-D lookup tables corresponding to Equations (2) and (6), respectively, are pre-setup for a given operating range of currents and temperatures by measurement or software simulation. The input–output from of the lookup tables are shown in Figures 3 and 4, respectively.

By inputting current and temperature, the corresponding conduction saturation voltage and switching energy losses can be obtained from the lookup tables, respectively. The average power loss over a simulation time step (a PWM switching cycle) can then be calculated by

$$P_{\mathrm{ave\_loss}}(k) = \frac{1}{T_{\mathrm{sw}}}(E_{\mathrm{cond\_loss}}(k) + E_{\mathrm{sw}}(k)) \tag{7}$$

where, k represents the kth switching cycle. If the dc link voltage or the device are changed, the lookup table needs to be changed to reflect these changes and a new table will be required.

3. INVERTER CIRCUIT ELECTRICAL SIMULATION FOR POWER LOSSES CALCULATION

The simplest approach to this problem is to assume that the semiconductors are ideal switches and diodes [6–8], however, this still results in an unacceptably large CPU time and memory

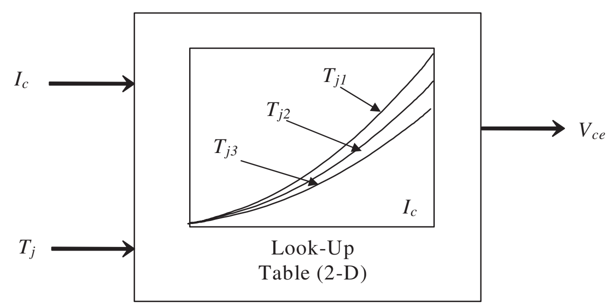

Figure 3. Lookup table of $V_{\mathrm{ce}}-I_{\mathrm{c}}$ characteristics.

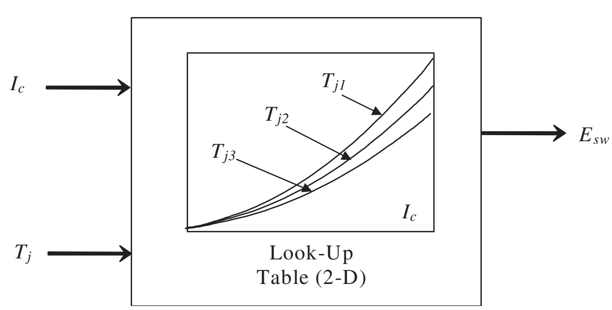

Figure 4. Lookup table of switching energy losses.

usage. In this paper a new methodology is proposed, initially, a system level inverter electrical simulation is carried out, in which, the inverter is modelled as a unity gain power amplifier, so that the simulation is run with a large simulation time-step that leads to a fast simulation speed. The electrical simulation gives the required electrical characteristics, such as the effective voltage and current. The PWM related signals, can then be mathematically derived from the effective voltage and current waveforms. The block diagram of the inverter electrical simulation is shown in Figure 5, where the *torque* and *speed* are the specified operating conditions as a function of time: V_{m} and I_{m} represent the 3-phase continuous effective motor voltage and current vectors.

4. PWM AND DEVICE CURRENT RECONSTRUCTION FOR POWER LOSSES CALCULATION

The simulation in Figure 5 gives the effective motor voltages and currents, which are continuous quantities (rather than the actual PWM switching waveforms) and cannot directly be used for power losses calculation. Therefore, a PWM reconstruction technique is proposed for the power losses simulation. By using PWM reconstruction technique, device current and the *on-time* of device can be mathematically reconstructed for power losses calculation. Figure 6 shows the block diagram of the PWM and device current reconstruction. where, V_{m} and I_{m} are the effective motor terminal voltage and current vectors. The mathematical method to reconstruct the PWM signals will depend on the actual PWM strategy applied. Based on the reconstructed PWM and device conduction current, the power losses calculation and the thermal simulation can then be carried out, the block diagram is shown in Figure 7, where I_{c}, T_j represents the device current and junction temperature. The average power losses of each PWM switching cycle is then calculated using Equation (7). Feeding this average power losses into the thermal network, the junction temperature can be calculated. The temperature predicted at the end of

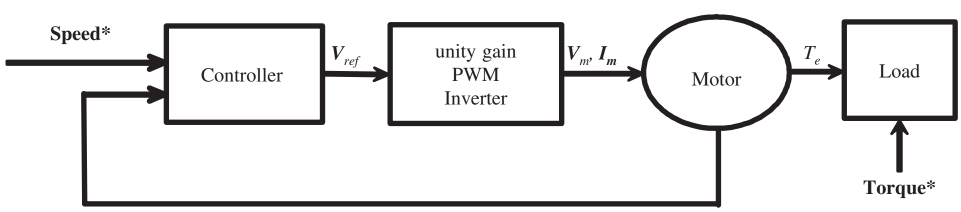

Figure 5. Block diagram of a continuous-time domain inverter system simulation model.

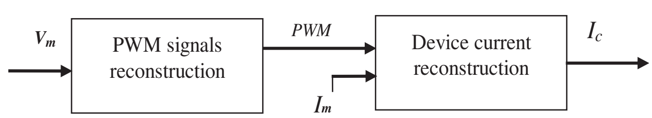

Figure 6. PWM signals and device conduction current reconstruction process using effective smooth motor stator voltage vector.

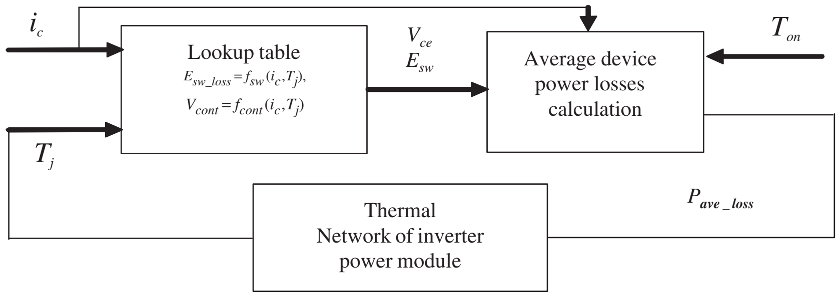

Figure 7. Block diagram of the power losses and thermal simulation model.

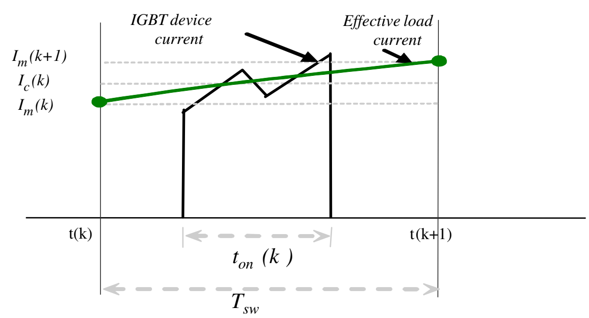

Figure 8. The switching current and the effective current waveforms.

one time step is assumed to be the temperature for the next time step, this means that there is a small delay which is insignificant at the switching frequencies used. In an inverter drive system, where significant inductance exists in load circuit, the current in the load circuit is continuous, since the time constant of the load circuit is large compared with the PWM switching waveform. However, when using a low PWM switching frequency such as 1 kHz, harmonics may need to be taken into account (see Figure 8).

To obtain an accurate output from the lookup table, in this paper, the current is given by Equation (8)

$$I_c(k) = \frac{1}{2}(I_m(k) + I_m(k+1)) \tag{8}$$

where $I_m(k)$, $I_m(k+1)$ are the sampled effective currents at kth and $(k+1)$th sampling points, respectively.

5. EXAMPLE OF THE POWER LOSSES CALCULATION METHOD AND THERMAL SIMULATION

5.1. Simulation circuits

A simulation program for proposed fast power losses and thermal simulation model has been built in MATLAB/SIMULINK. To verify the accuracy of the proposed model, a simulation of an inverter circuit using compact model has been carried out using SABER. The circuits are shown in Figures 9 and 10. Both the inverters work with the same reference three-phase voltage V_{ref} the amplitude of V_{ref} is 150 V, and the frequency is 50 Hz. The same three-phase inductive-resistive load are applied for both circuits, and $L = 3.2\,\text{mH}$, $R = 1.3\,\Omega$. The PWM switching frequency is 1 kHz, dc link voltage is 500 V. The IGBT with a blue-circle is to be analysed in this paper.

5.2. Physically based IGBT compact model

Physically based compact model [9, 10] has been selected as a switching device in the inverter circuit. This model used the full ambipolar diffusion equation (ADE) solution and solving numerically the basic drift-diffusion equations. The power devices model uses one-dimensional approximation to the carrier distribution in the drift region. It includes conductivity modulation and non-quasistatic charge storage effect, which will describe correctly in static and dynamic behaviour of power bipolar devices. These models have been implemented in the SABER circuit simulator for today modelling environment.

The behaviour of power devices depends heavily on the excess carrier distribution in the low-doped drift region during the high-level injection mechanism. However, quasi-neutrality ($p \approx n$)

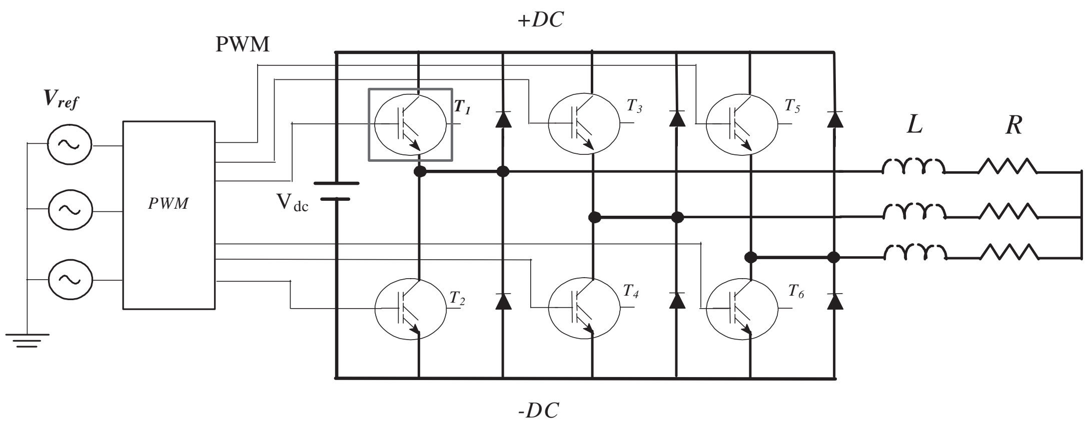

Figure 9. Physically compact model based inverter system (Model-1).

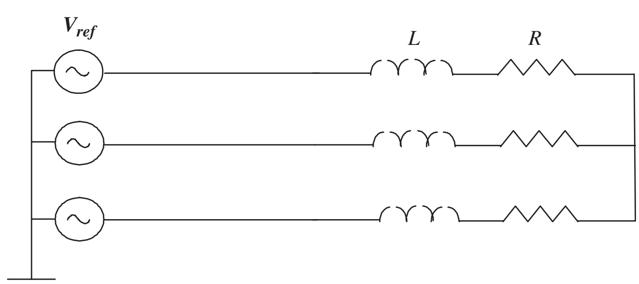

Figure 10. Unity gain power amplifier inverter system (Model-2).

is valid because the mobility will not undergo velocity saturation as a consequence of a high electric field. Under this condition, the ADE determines the carrier transport

$$D_{\mathrm{a}}\frac{\partial^2 p(x,t)}{\partial x^2} = \frac{p(x,t)}{\tau_{\mathrm{A}}} + \frac{\partial p(x,t)}{\partial t} \tag{9}$$

where D_{a} represent the ambipolar diffusion constant and τ_{A} represent the ambipolar carrier lifetime. Our approach [9, 10] is to model the ADE describing the internal plasma charge distribution using two exponential functions based on the ambipolar diffusion length

$$p = A\mathrm{e}^{x/L} + B\mathrm{e}^{-x/L} \tag{10}$$

where the diffusion length is given by $L = \sqrt{D\tau}$, and A and B are arbitrary constants determined by boundary conditions. These functions are fitted using weighting factors, which are adjusted using a numerical fitting algorithm. Although this approach accurately represents the evolution of the plasma with time, it requires considerable effort to ensure robustness of the model. Our approach does not make any assumptions about the shape of the plasma and so is more generic with a sound physical basis. Indeed compared with a full numerical model our approach shows excellent agreement over a wide range of operating conditions. In order to model heavily doped n^+ buffer layer existing in the IGBT device, the width of the undepleted region and the carrier current at the boundaries (x_l and x_r) of the region are required. Now, the boundary conditions for the n-base carrier storage region are defined by (11) and (12) as

$$\left.\frac{\partial p}{\partial x}\right|_{x_l} = \frac{1}{2qA}\left(\frac{I_{\mathrm{p}} - I_{\mathrm{pr,b}}}{D_{\mathrm{n}}} - \frac{I_{\mathrm{pr,b}}}{D_{\mathrm{p}}}\right) \tag{11}$$

$$\left.\frac{\partial p}{\partial x}\right|_{x_r} = \frac{1}{2qA}\left(\frac{I_{\mathrm{ch}}}{D_{\mathrm{n}}} - \frac{I_{\mathrm{p}} - I_{\mathrm{ch}}}{D_{\mathrm{p}}}\right) \tag{12}$$

A is the cross sectional area of the device, D_{p} and D_{n} the hole and electron diffusion coefficients, $I_{\mathrm{pr,b}}$ is the minority carriers in the buffer layer and the hole current at the n-base end (x_l). I_{ch} the channel current in the inversion layer and I_{p} the total plasma current flow through the drift region.

The parameter definitions of physically based IGBT model have been shown in the nomenclature. However, the default parameters will be used in the model as a comparison with the MATLAB model. Due to the aim of this paper to show fast power losses simulation method, the parameter variation with temperature is not considered in the physically based IGBT compact model. The behaviour diode mode that come from the standard SABER library is employed in the model-1 circuit simulation. The conduction $V_{\mathrm{ce}}-I_{\mathrm{c}}$ and switching energy loss characteristics of the physically based compact IGBT model are shown in Figures 11 and 12, respectively.

5.3. Power losses simulation results and comparisons

For the purpose of verifying the accuracy of the proposed model, simulation of 40 ms is carried out with both models under same operating conditions. The time-step for Model-1 is 8e-8 s, and for Model-2, is le-3 s. The two device characteristics lookup tables used for model-2 simulation are calculated using the physically based IGBT compact model characteristics in Figures 11 and

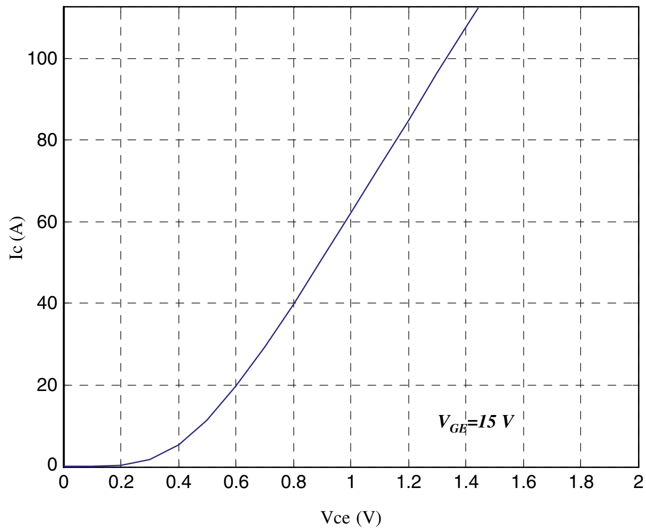

Figure 11. $V_{ce}-I_c$ characteristics of the compact IGBT model.

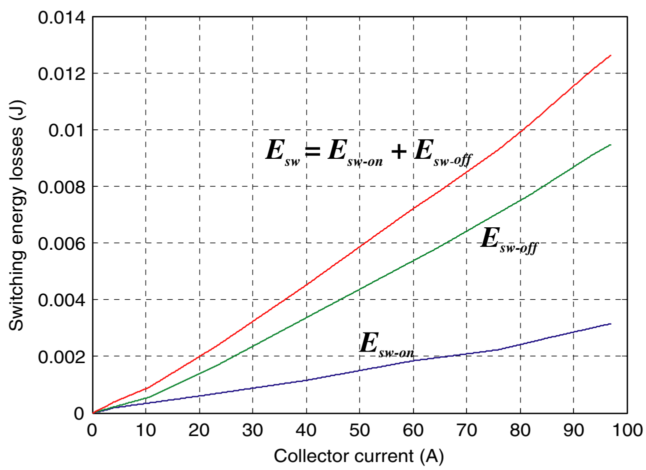

Figure 12. Switching energy loss characteristics of the compact IGBT model.

12. The processes of the simulations show that the speed for Model-2 is thousands of times faster than the Model-1.

Figure 13(a) shows the reference three phase voltages applied for models and the triangular signals used to generate the PWM, and Figure 13(b) gives the load current of phase-A. It can be seen that the almost same currents are generated, except that some harmonics exist with the Model-1. This confirms that the inverter modelled in a unity gain amplifier in Model-2 plays the same role compared to the inverter based on physically compact device model, and also means that using the effective smooth voltage to reconstruct the PWM is reasonable. Figure 14(a) and (b) give the IGBT switching current and power losses of the blue-circled IGBT in Figure 9 with the simulation of Model-1.

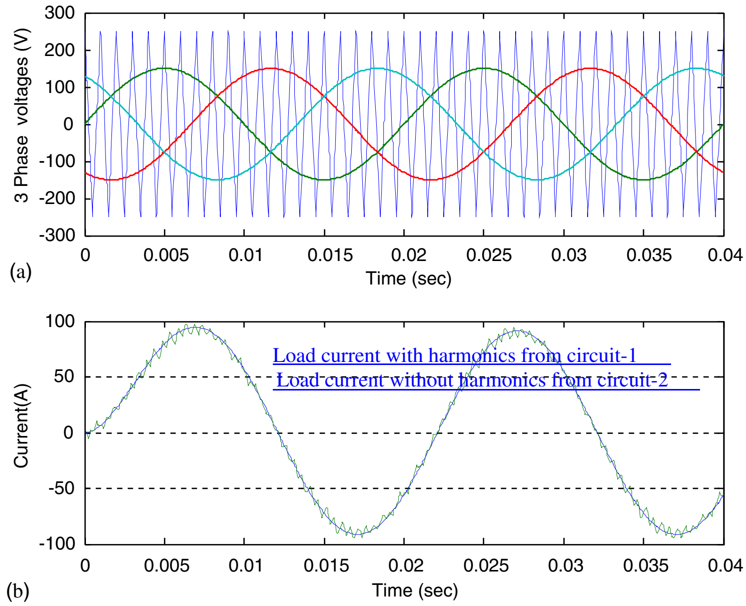

Figure 13. (a) Reference voltage; and (b) load currents.

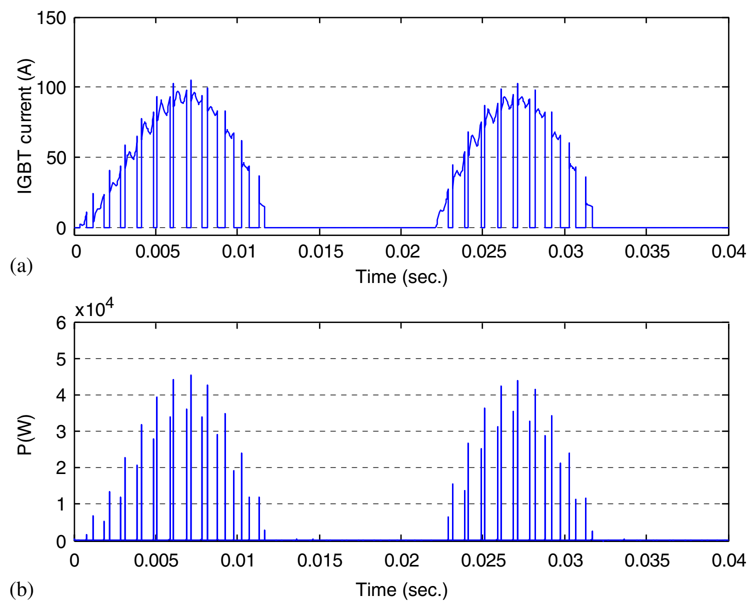

Figure 14. (a) IGBT device current; and (b) power losses.

The conduction, turn-on and turn-off energy losses have been calculated at each PWM switching cycle using the lookup table method, and the three portions of the energy loss at each switching cycle are given in Figure 15. It can be seen that for 1 kHz PWM switching frequency,

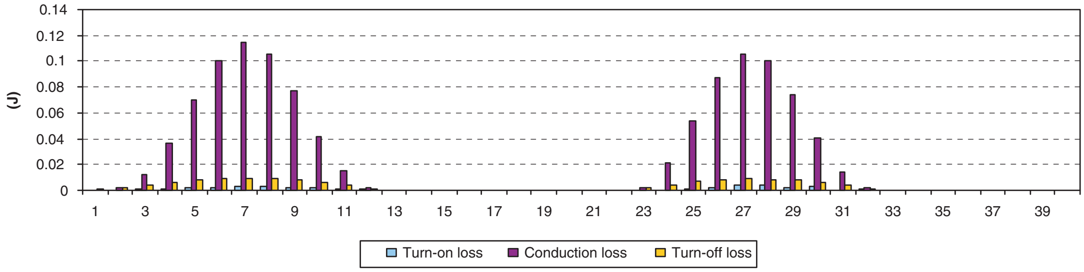

Figure 15. Conduction, turn-on and turn-off energy losses chart at each switching cycle.

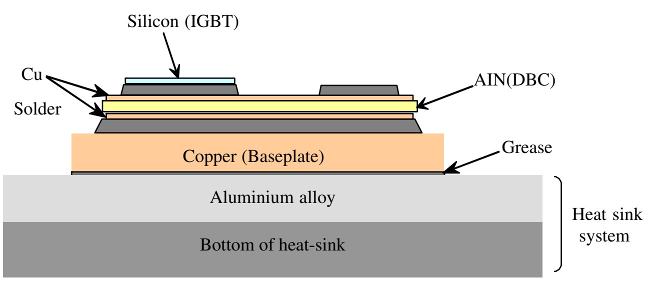

Figure 16. The cross-section of the power module with heat-sink.

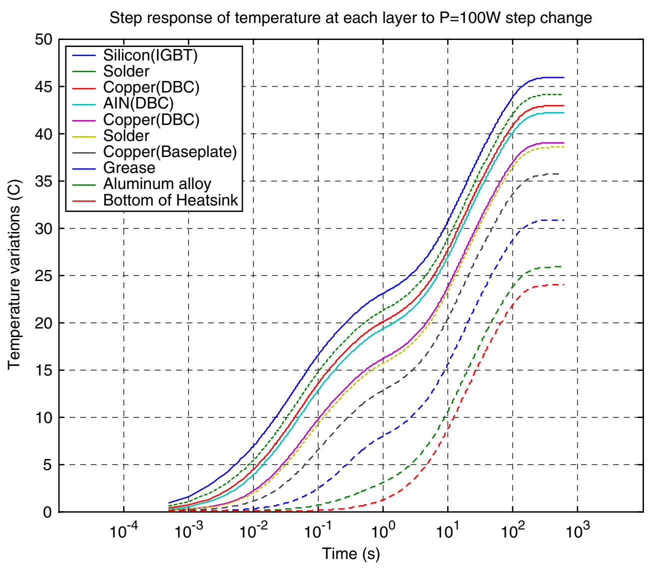

Figure 17. Step response of temperature at each layer of the power module (FEM).

the conduction energy losses is bigger than the switching energy losses in most of switching cycles.

5.4. Thermal simulation and comparison

For the purpose of accuracy comparison, thermal simulations are carried out for IGBT module and heat-sink system. The cross-sectional view of the IGBT module is shown in Figure 16. Silicon IGBT are mounted on the top of an insulator layer, and that is fixed on a base-plate, the IGBT module is mounted on a laboratory heat-sink. A compact thermal network composed of thermal resistance and thermal capacitance shown in Figure 18, is employed as a thermal model for carrying out the thermal simulation, because the compact thermal model can be implemented using a circuit simulator easily.

A transient heat 3D-FEM simulation of the IGBT module with a laboratory heat sink is carried out. The transient response of temperature at each layer of a single IGBT module with heat-sink is shown in Figure 17 [2], where, a 100 W step changed heat source is applied, based on

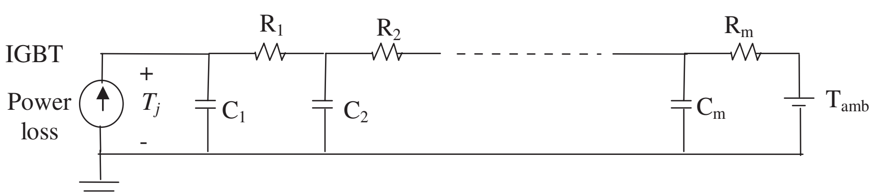

Figure 18. Equivalent compact thermal network of IGBT module and heat sink system.

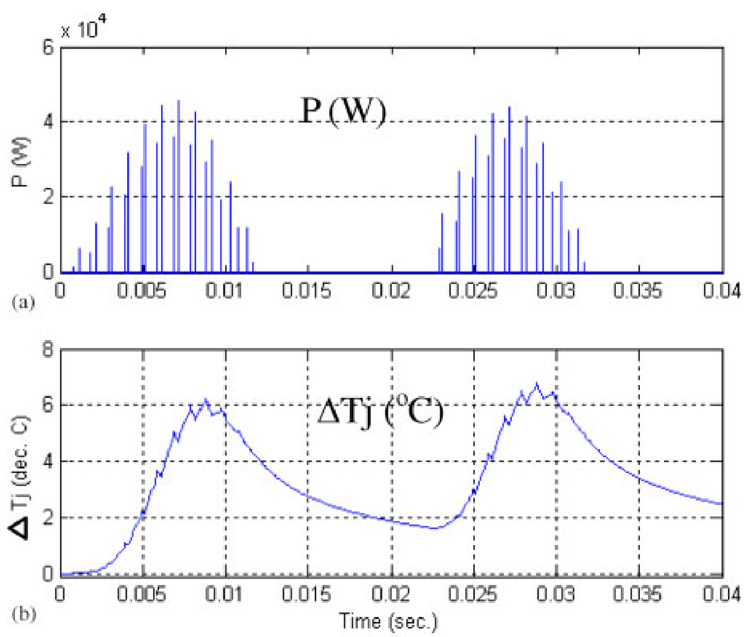

Figure 19. Power losses and variation of junction temperature (Model-1).

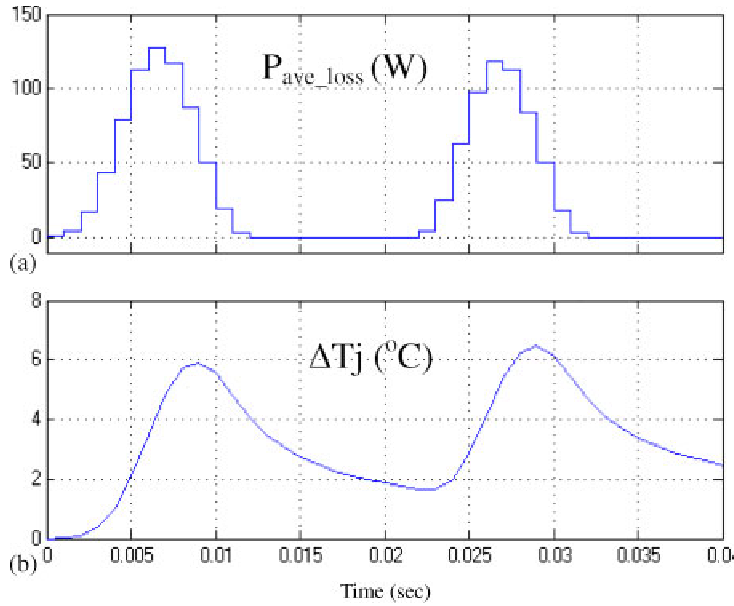

Figure 20. Power losses and variation of junction temperature (Model-2).

the transient response of temperature at each layer, the corresponding thermal impedance of the compact model can be obtained.

The thermal network shown in Figure 18 is assumed to be connected with the Phase-A upper side IGBT, so power loss of phase-A upper side IGBT is considered as input heat source to the thermal network, the ambient temperature T_{amb} is set to be 25°C. With Model-1 thermal simulation, the power losses are directly from the circuit simulation and are shown in Figure 19(a). With Model-2 thermal simulation, the average power losses over each PWM switching cycle (1 ms) is calculated based on the compact model characteristics in Figures 11 and 12 using proposed lookup table method and PWM reconstruction technique and is shown in Figure 20(a).

Applying the heat sources shown in Figures 19(a) and 20(a), respectively, to the thermal network in Figure 18, the variations of IGBT junction temperature ΔT_j are simulated with the two different type of heat sources, and the results are shown in Figures 19(b) and 20(b), respectively. It can be seen that the junction temperatures are almost the same. The advantages of the Model-2 is that due to the large simulation time-step applied (1 ms), the simulation speed is thousands of times faster than Model-1 where simulation time step is le-8 s.

6. CONCLUSIONS

A fast power losses simulation model for a three-phase inverter power module (IPM) thermal simulation is proposed in this paper. Based on the unity gain power amplifier inverter model, device characteristics, PWM and devices current reconstruction, and lookup table techniques, the fast accurate power losses simulation method is implemented for power device thermal simulation. Due to large simulation time steps applied, this allows power losses and thermal performance of devices in an inverter to be predicted over long period of real time. This simulation methodology brings together accurate models of the electrical systems performance, state of the art-device compact models and a realistic simulation of the thermal performance in a usable period of CPU time.

ACKNOWLEDGEMENTS

The authors would like to thank Toyota Motor Corporation for their financial support and technical advice.

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