FACTA UNIVERSITATIS
Series:**Electronics and Energetics** Vol. 27, N^o 1, March 2014, pp. 13 - 23
DOI: 10.2298/FUEE1401013I

REVIEW OF ADVANCED IGBT COMPACT MODELS DEDICATED TO CIRCUIT SIMULATION

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Abstract. *The paper aims to review the research area of the IGBT compact modelling and to introduce different device models. The models are separated in two groups, one that solves ambipolar diffusion equation (ADE) and one that does not. Both types of compact models have been successfully used in the past for power electronic circuit design.*

Key words: *IGBT, compact, model, power, inverter, circuit*

1. INTRODUCTION

Insulated Gate Bipolar Transistors (IGBTs) are devices of choice in modern power converter systems targeting medium to high voltage and current applications, such as hybrid or electric vehicles [1], [2]. During the power circuitry early design stages, one could consider an IGBT to be a binary on-off switch, thus achieving very fast simulation of the converter operation. However, this modelling approach cannot be used to analyze some key aspects of the device and converter performance such as heat dissipation for example, very important design parameter especially during operation at high switching frequencies [3]-[5]. Obviously, this will not lead to robust equipment design, as it does not provide any information regarding switch failure mechanisms.

To overcome all the above issues, one could develop and use the IGBT models based on full internal physics of the device. These would be typically 2D or 3D finite-element (FE) models developed and run in some of the commercially available simulation tools. This modelling approach will provide designers with detailed knowledge of the IGBT devices, but it requires very long simulation time, and it is numerically prohibitive if one would like to study complex circuits containing multiple power devices requiring many switching events [6].

The compact modelling approach is placed between these two extremes [7]-[27]. Compact models are lower complexity, but yet fully physically based and very accurate,

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Received December 18, 2013
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models of the power devices dedicated to circuit simulation. This physical modelling approach could be based on certain mathematical simplifications of the fundamental semiconductor charge transport equations, for example [9], [20], [21], [24]. In order to develop an IGBT model that will describe correctly its static and dynamic behaviour, the main challenge is to incorporate into a device model conductivity modulation and non-quasistatic charge storage effects [22], [23].

The absence of an industry-accepted IGBT model and the pronounced industry-need for more accurate IGBT compact model have triggered very intensive research in this area for more than a decade. The distinct challenge in developing IGBT compact model for circuit simulation lays in the fact that model needs to satisfy some refuting requirements. It needs to provide high quantitative accuracy, short CPU time, and physical, yet easy to determine model parameters. As a result different IGBT compact models have been developed and presented in the literature, some of those suitable for long time inverter simulations (minutes) [15]. The aim of this paper is to review the research area of the IGBT compact modelling and to introduce different models, such as IGBT models based on the Ambipolar Differential Equation (ADE) solutions [18]-[24] and the ones which are not solving ADE, typically physics-based sub-circuit models [14], [16], [17], [25]-[27].

2. ADE Solution Based Models

To describe IGBT's static and dynamic behaviour, the incorporation of conductivity modulation and non-quasistatic charge storage effects into the device model is vital [7], [22]. When the excess carrier density overcomes the IGBT's n^- base doping level by several orders of magnitude within the carrier storage region, the assumption that the excess electron concentration, n, and excess hole concentration, p, are equal is valid [21]. Then, the carrier transport is determined by the Ambipolar Diffusion Equation (ADE):

$$D\frac{\partial^2 p(x,t)}{\partial x^2} = \frac{p(x,t)}{\tau} + \frac{\partial p(x,t)}{\partial t} \tag{1}$$

where D represents the ambipolar diffusion constant and τ stands for the ambipolar carrier lifetime. The boundary conditions for the above equation are determined by the current at the left (x_l) and right (x_r) ends of the carrier storage region. At the left end of the carrier storage region, the electron and hole currents are given by:

$$I_{electron}(x_l) = I_{nl} = Anq\mu_n E(x_l) + AqD_n \left.\frac{\partial n}{\partial x}\right|_{x_l} \tag{2}$$

$$I_{hole}(x_l) = I_{pl} = Apq\mu_p E(x_l) - AqD_p \left.\frac{\partial p}{\partial x}\right|_{x_l} . \tag{3}$$

In the above equations, q stands for the electron unity charge, A is the cross sectional area of the carrier storage region, D_n and D_p stand for the electron and hole diffusion constants respectively, μ_n represents the electron mobility, μ_p is the mobility of the holes, and E stands for the electric field. Dividing equation (2) with D_n and equation (3) with D_p and then subtracting (3) from (2) (under condition $n \approx p$ and $\mu_n / D_n = \mu_p / D_p$) gives a derivative boundary condition on p for the ADE at the left end of the carrier storage region [18]-[22]:

$$\left.\frac{\partial p}{\partial x}\right|_{x_l} = \frac{1}{2qA}\left(\frac{I_{nl}}{D_n} - \frac{I_{pl}}{D_p}\right). \tag{4}$$

A similar expression is obtained for the right end of the carrier storage region:

$$\left.\frac{\partial p}{\partial x}\right|_{x_r} = \frac{1}{2qA}\left(\frac{I_{nr}}{D_n} - \frac{I_{pr}}{D_p}\right), \tag{5}$$

where I_{nr} and I_{pr} represent electron and hole currents respectively at the cathode end (see Fig. 1).

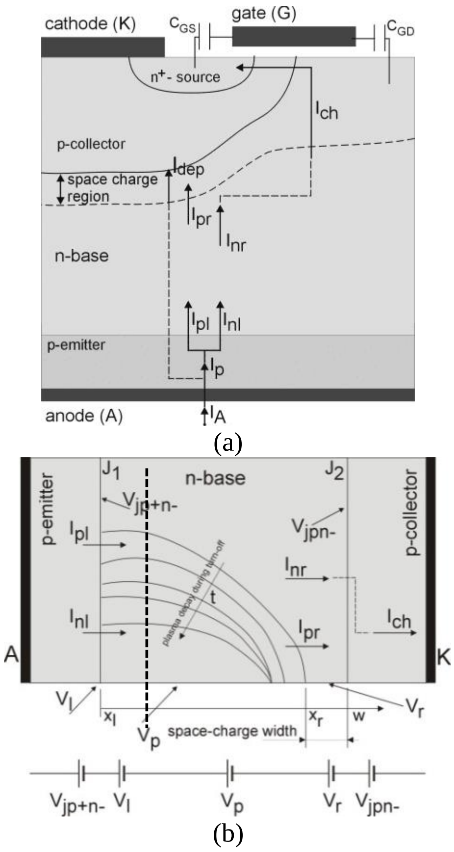


Fig. 1 Schematic representation of the NPTIGBT structure (a) and bipolar part of the structure (b)

2.1. Exponential solution based models

An exponential approximation based solution for this equation has been developed. To model the plasma carrier distribution, set of exponential shape functions is used [21]. These shape functions are found to model the shape of the plasma correctly, without oscillations in the internal distribution. The slopes to the carrier distribution at the boundaries are also physically correct. In steady state forward bias operation the plasma carrier concentration has a distribution of catenary form requiring just two exponential basis functions giving [21]:

$$p = Ae^{x/L} + Be^{-x/L} \tag{6}$$

where L is the diffusion length. In transient operation, more complex profiles can be approximated using a number of exponential basis functions with a range of decay length parameters, shorter than the steady state ones. The models reported in [21], [22] actually uses up to seven exponential basis functions to model the plasma distribution during transient operation. To implement model and make it functional, one needs to determine forward junction voltage between p+ emitter and n- base (see Fig. 1), the ohmic voltage drop across the plasma region, the depletion voltage, depletion capacitance, and depletion current at the anode end. The depletion current is a small extra current component that exists under high speed transient conditions as described in [21], [22]. This model has been used successfully to predict switching characteristics of different commercially available IGBT devices; one example is given in Fig. 2.

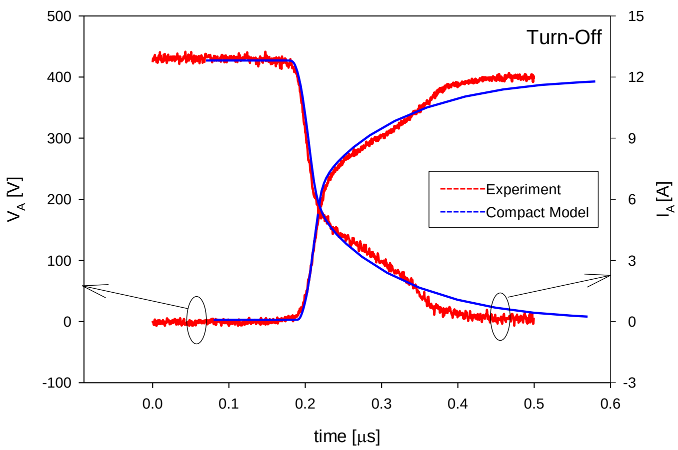


Fig. 2 IGBT turn-off characteristic - experimental results vs. compact model

The downside of this modelling approach is model complexity [20], large number of model parameters [19], [22], difficult model implementation in circuit simulators such as PSPICE, long simulation time. Many modern IGBT devices include localize life time control (LLC) region in order to reduce current tail during device turn-off and increase operating frequency [11]. The above discussed model does not have the ability to directly include LLC region. In order to include this feature, it needs some alterations. If, for example, LLC region is inserted between the p-emitter and arbitrary dashed line shown in Fig. 1b, the plasma carrier distribution model will need to be consider separately across these two regions(each having different lifetimes) left and right from the arbitrary dashed line. This would introduce another set of equations needed to describe boundary between the LLC region and rest of the n- base region, thus making model even more complex.

2.2. Fourier series solution based models

A Fourier series solution based model has been developed and described in [24]. It has been based on the research results showing that the diffusion equation could be solved by means of an electrical analogy [23]. The plasma carrier concentration has a distribution of a sum of Fourier series components in space:

$$p = p_0(t) + \sum_{k=1}^{\infty} p_k(t)\cos\left(\frac{\pi k(x - x_1)}{x_2 - x_1}\right) \tag{7}$$

where k represents the harmonic number. Set of equations described in (7) can be represented in the form of two RC lines corresponding to the even and odd values of k. The RC lines are driven by currents defined by the boundary conditions as described in [24].

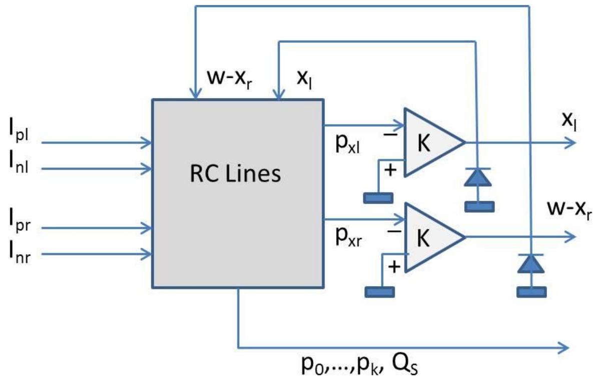


Fig. 3 Analogue solution to the ADE [24]

Fig. 3 shows analogue solution to the ADE with fixed or mobile boundaries. In Fig. 3, Q_S represents total carrier stored charge, $p_0,\ldots,p_k$ stand for Fourier series coefficients, w is the width of the n-base region, x_l and x_r are the positions of the plasma region left and right boundaries (see Fig. 1b), p_{xl} and p_{xr} corresponds to the excess carrier concentration at x_l and x_r respectively, and currents $I_{pl,r}$ and $I_{nl,r}$ are as depicted in Fig. 1a. The model could be implemented in any general purpose simulation software having non-linear elements and variable parameters [24].

3. Physics-Based Sub-Circuit Compact Models

Common feature of all sub-circuit compact models is that they are not trying to solve ambipolar diffusion equation in order to reproduce measurement data or predict device characteristics. Recently, HiSIM-IGBT compact model has been developed and presented [25]-[27]. Model is based on the consistency of the potential distribution within the IGBT device by considering in great details the MOSFET surface potentials and the BJT junction potentials, as described in [25], [27]. The model has been originally developed for Trench IGBT device. HiSIM-IGBT equivalent circuit is shown in Fig. 4. The IGBT's

MOSFET part is described with a conventional model, and the main model development effort has been put into extending the BJT shown in Fig. 4, since IGBT output current is managed by the bipolar transistor theory. In this model, the IGBT characteristics are determined by three parameters, the trench-bottom MOSFET gate charge, Q_{TB}, the base resistance, R_B, and the NQS IGBT base charge model, Q_B.

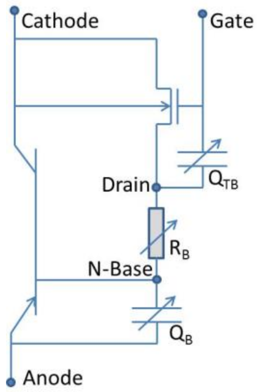


Fig. 4 HiSIM-IGBT equivalent circuit [27]

Another popular IGBT model has been presented by Jankovic *et al.* in [14], [16], [17]. In [14], the physics-based IGBT sub-circuit model which successfully included the effects of localised lifetime control (LLC) on device electrical performance has been described. In particular, the model depicts the non-punch trough IGBTs with different locations of LLC region. In what follows, the description of model implementation in SPICE will be given with attention to the modifications performed to include the lifetime control effects. The equivalent sub-circuit of the IGBT model implemented in SPICE is shown in Fig. 5.

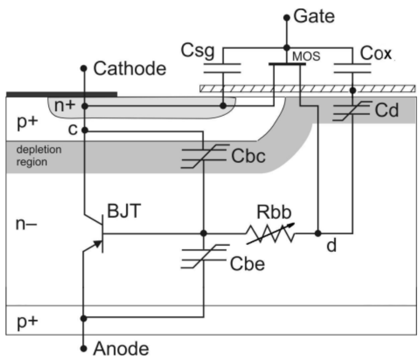


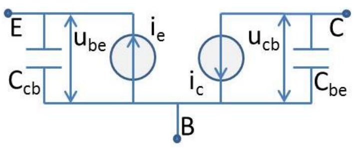
Fig. 5 LLC IGBT equivalent circuit

It includes a N-channel MOSFET, a wide-base PNP bipolar transistor (BJT), the voltage-dependent base resistor R_{bb}, the p-n junction capacitances, C_{bc} and C_{be}, the gate overlapping source capacitance C_{gs}, and the drain-gate overlapping capacitance C_{gd} (the gate-overlap capacitor C_{ox} in series with the gate induced depletion capacitance C_d). The

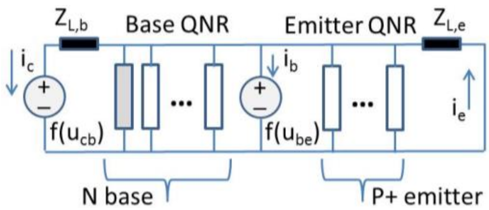
N-channel MOSFET part of the IGBT is modelled using a SPICE Level 5 model. PNP BJT is low efficient and its operation is fully affected by the LLC technique. Fig. 6a shows the schematic of the PNP BJT circuit model consisting of two voltage-controlled current sources (i_e and i_c) and the junction capacitances C_{be} and C_{cb}. The current sources mirror the input/output currents of separately developed sub-circuit shown in Fig. 3b. The carrier transport trough the emitter and the base Quasi-Neutral Regions (QNRs) are described with two equivalent loosy transmission lines (TLs) consisting of identical RC-cells shown in Figure 6c. The input voltage generators $f(u_{be})$ and $f(u_{cb})$ perform the voltage transformations $\{\exp(u_{be}/V_t) - 1\}$ and $\{\exp(u_{cb}/V_t) - 1\}$, respectively. The RC-cell elements are non-linear conductance, capacitance, resistance and load impedance denoted as G_k, C_k, R_k and Z_l, in Fig. 6. Their values are calculated by the following formulas [21]:

$$
\begin{aligned}
C_k &= \frac{C_2}{\sqrt{1+C_1 \; u_{2k}}} \\
G_k &= \frac{1}{C_3(1+C_1 u_{2k})+C_4} + C_5(-1+\sqrt{1+C_{1u_{2k}}}) + C_6(1+\sqrt{1+C_1 u_{2k}}) \\
R_k &= C_7(1+\sqrt{1+C_1 u_{2k}})\sqrt{1+C_8\left[\frac{u_{2k-1}-u_{2k+1}}{\sqrt{1+C_1 u_{2k}}(1+\sqrt{1+C_1 u_{2k}})}\right]^2} \\
Z_l &= C_9(1+\sqrt{1+C_1 u_{2n+1}})
\end{aligned}
\tag{8}
$$

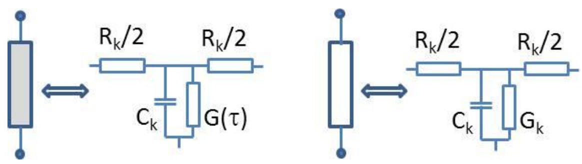
where u_{2k-1}, u_{2k+1} and u_{2k} are the input, the output and the middle node voltage, respectively, in the k-th RC-cell.



(a)



(b)



(c)

Fig. 6 LLC IGBT model details

The parameters C_1-C_9 are related to the physical and technology parameters of the emitter or the base QNRs as:

$$C_1 = \frac{4n_{ie}^2}{N_D^2},\quad C_4 = \frac{2\tau_0}{qn_{ie}w},\qquad C_7 = \frac{N_D\,w}{2qV_T n_{ie}^2 \mu_0},$$
$$C_2 = \frac{qn_{ie}^2 w}{N_D}, C_5 = \frac{qn_{ie}^2 C_n N_D w}{2}, C_8 = \left(\frac{\mu_0}{v_{sat}}\frac{C_1 V_T}{2w}\right)^2, \tag{9}$$
$$C_3 = \frac{\tau_0 N_D}{qn_{ie}^2 w}, C_6 = \frac{qn_{ie}^2 C_p N_D w}{2}, C_9 = \frac{N_{D,end}}{2\,q\,n_{ie,end}^2\,v_s}$$

where N_D is the doping concentration, μ_0 is the doping-dependent low-field mobility, v_{sat} represents the drift saturation velocity, $\tau_0$stands for the low-injection level doping-dependent minority carrier lifetime, n_{ie} is the effective intrinsic carrier concentration incorporating band-gap-narrowing effects , and C_n, C_p are the Auger's recombination constants. The parameter*w*is a physical width of single RC-cell, which is obtained by dividing a zero-bias QNR width W with the chosen number n of RC-cells ($w=W/n$). In [14], the emitter QNR is represented with three RC-cells. Since the LLC substantially decreasesτ_0 of particular device area, it follows from eqs. (8) and (9) that the RC-cells of controlled recombination region must have different G_k element. It is illustrated in Fig. 6c where the RC-cell of the controlled region is shown separately with different conductance $G(\tau)$. Note that the BJT with the first (from left to right) RC-cell shaded shown in Fig. 6b corresponds to the location of the LLC region within the IGBT.

The model, as described above, has been used successfully for the prediction of the inverter circuit power losses [16] and also to investigate IGBT tail current characteristics at different temperatures as shown in Fig. 7.

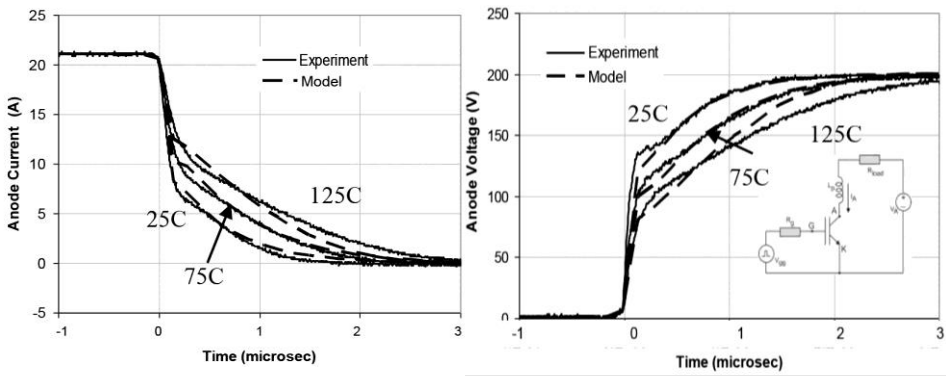


Fig. 7 Simulated and measured anode tail current and anode voltage of PT IGBT during the device turn-off at 25°C, 75°C and 125°C

4. ELECTRO-THERMAL (ET) MODELLING STRATEGY

Thermal compact model of an IGBT is equally important as its electrical counterpart to accurately predict circuit performance [3]-[5]. The work presented in [3] describes an ET modelling strategy that has been widely accepted by compact modelling research community and successfully applied since. It could be described in what follows. Adding an extra node, thermal node, to the electrical compact model of the IGBT device an electro-thermal (ET) models can be formulated. This thermal node has information regarding junction temperature of the device T_j and it represents a connection between the active devices and rest of the circuit thermal network [3]. This is schematically represented in the Fig. 8. A structure diagram of the ET compact device model which shows the interaction between thermal and electrical networks through the electrical and thermal nodes is shown in Fig. 9. As can be seen from the Fig. 9, the instantaneous value of the device temperature estimated by the thermal network is used for the calculation of the temperature dependent model parameters and temperature dependent silicon properties. Then, these temperature dependent values are used by the ET compact device models to calculate instantaneous electrical characteristics as well as instantaneous dissipated power. Finally, the dissipated power is used as an input parameter by the thermal network, and the device electrical characteristics are transferred to the electrical network.

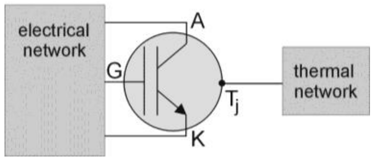


Fig. 8 IGBT ET compact model – electrical contacts (G, A, K) as well as thermal node (T_j) are shown

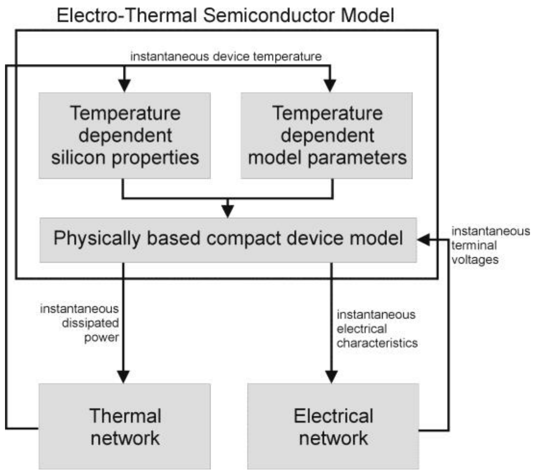


Fig. 9 Structure diagram of the ET compact model – interaction with the electrical and thermal network is shown

The thermal parts of the compact models are represented using a thermal RC network due to an electrical analogy [4]: thermal resistance is represented by an electrical resistance, thermal capacitance by an electrical capacitance, and dissipated power by current source [29]. Either Foster or Cauer RC networks can be used for this purpose [28], [29]. Since the Foster network is not directly suitable for the heat-flow path identification (because of the node-to-node heat capacitances), the Cauer RC network is preferred choice for thermal device characterisation. Cauer network includes only node-to-ground capacitances and it represents a discretised image of the real heat-flow structure. Network elements can be determined by using a deconvolution method for extraction of the RC thermal network parameters from the thermal transient response of the device for a step function excitation [28]. Namely, applying an abrupt dissipation step onto the chip, the time-function of the rise of the chip temperature has to be determined. Either experimental method or 3D Finite Element Model could be employed to obtain these thermal transient responses [29].

5. CONCLUSIONS

The research area of the IGBT compact modelling has been reviewed and different device models have been introduced. The models could be separated in two groups, ones that solve ambipolar diffusion equation (ADE) and others that do not. The models based on ADE solution, one could claim, are more physically based, but they are more complex to include in standard circuit simulator, need longer CPU time, might have convergence problems when simulating the circuits with larger number of IGBTs. Both types of compact models have been successfully used in the past for power electronic circuit design.

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