PHYSICAL MODELING OF IGBT TURN ON BEHAVIOR

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Abstract – Although IGBT turn on losses can be comparable to turn off losses, IGBT turn on has not been as thoroughly studied in the literature. Under clamped inductive load condition at turn on there is strong interaction between the IGBT and the freewheeling diode undergoing reverse recovery. A physics-based IGBT model is used that has been proved accurate in the simulation of IGBT turn off. Both resistive and inductive turn on are considered. Discrepancies between model predictions and experimental results are discussed.

I. INTRODUCTION

Nowadays IGBTs are widely used in switching power converter applications. New device generations exhibit constantly improving electrical characteristics, and as a result their use is expanding. Higher-voltage IGBTs are used in high-voltage applications where traditionally thyristors were used. Fast IGBTs allow operation at higher switching frequencies and are displacing MOSFETs in many low-voltage applications.

Accurate IGBT models are desirable in order to accurately simulate switching waveforms and estimate device stresses, and switching and conduction losses. A complete physics-based electro-thermal IGBT circuit simulator model has been presented before [1-2]. Its high accuracy has been validated for various structure IGBTs such as punch-through (PT), non-punch-through (NPT), light-punch-through and field-stop (FS). Its usefulness is enhanced by its practical parameterization procedure and reasonable simulation speed [3].

Usually in the study of IGBTs, the attention is focused on the turn off behavior, since the IGBT current tail causes significant losses. Under hard switching conditions, the IGBT turn off happens under clamped inductive load condition. Therefore, validation for the physics-based model was performed under this condition.

Device manufacturers have expended significant effort to reduce current tail losses, using techniques such as lifetime control in the buffer layer to optimize device characteristics. On the other hand, turn-on losses can be significant, due to the diode reverse recovery, and be comparable to turn off losses. It is therefore of interest to simulate the IGBT turn on behavior, which so far has received scarce attention in the literature.

In this work the physics based IGBT model is used to simulate turn on behavior under resistive and clamped inductive switching conditions.

In section II the physics-based IGBT model is briefly reviewed. In section III experimental validation of the model under inductive turn off condition is presented. In section IV the IGBT turn on behavior is briefly discussed. In section V resistive turn on is considered in detail; the turn on process is described and a comparison of simulation and experimental waveforms is presented. The reason for some observed discrepancies is discussed. In section VI experimental and simulation results for inductive turn on are presented and possible reasons for observed discrepancies are given.

II. PHYSICS-BASED IGBT CIRCUIT SIMULATOR MODEL

A. Fourier-based-solution Approach

Among many analytical IGBT modeling approaches presented in the literature, Fourier-based-solution (FBS) approach has a better trade-off between the accuracy and simulation speed, and has been firmly established after extensive experimental validation and parameterization. The key of this approach is the physics-based description of the carrier distribution in the IGBT n drift region.

Like other conductivity-modulated devices, the behavior of an IGBT depends heavily on the carrier distribution in the wide base region, and the ambipolar carrier diffusion equation (ADE) describes the carrier dynamics in this region under high-level injection conditions.

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

where \(D\) is the ambipolar diffusion coefficient, \(\tau\) is the high-level carrier lifetime within the drift region and \(p(x,t)\) is the excess carrier concentration. Therefore, solving the ADE is the key to modeling the IGBT behavior.

![Fig. 1. Equivalent circuit to describe drift region carrier distribution](image)
After applying the Fourier transformation to the ADE, the drift region carrier distribution can be represented with the equivalent RC network shown in Fig. 1. The detailed discussion of the FBS approach and the corresponding equivalent circuit implementation can be found in [4] and [1].

The representation requires the width of the undepleted base region and the hole and electron currents at the boundaries of the region (\(x_1\) and \(x_2\)), from which one can calculate the gradients of the carrier concentrations, \(f(t)\) and \(g(t)\) at \(x_1\) and \(x_2\), respectively. The functions \(f(t)\) and \(g(t)\) are defined as follows:

\[
f(t) = \frac{\partial p(x,t)}{\partial t}
\]

\[
g(t) = \frac{\partial p(x,t)}{\partial t}
\]

\(A\) is the cross-sectional area of the device, \(D_e\) and \(D_p\), the electron and hole diffusion coefficients, \(I_{n1}\) and \(I_{p1}\) the electron and hole currents at \(x = x_1\) (p' side), and \(I_{n2}\) and \(I_{p2}\) the electron and hole currents at \(x = x_2\) (p-body side). The variable definition is shown in Fig. 2. Clearly, the success of the approach now depends solely upon developing the appropriate boundary conditions — hole and electron currents at the edges of the drift region.

**Fig. 2. Boundary condition definition for the PT IGBT**

The different IGBT structures have different boundary current definitions. For example, the electron current at emitter side \((I_{n1})\) of NPT IGBT is given in Equation (4):

\[
I_{n1} = qAh_pP_{10}^2
\]

where \(h_p\) is the recombination parameter, while the hole current at emitter side \((I_{p1})\) of PT IGBT is given by Equation (5):

\[
I_{p1} = \frac{qAD_{ph}}{L_{ph} \sinh\left(\frac{W_H}{L_{ph}}\right)} [P_{H0} - P_{HW} \cosh\left(\frac{W_H}{L_{ph}}\right)] + I_{QH}
\]

where \(D_{ph}\) is hole diffusion coefficient in the buffer layer, \(L_{ph}\) is excess carrier diffusion length and the term \(I_{QH}\) represents the capacitive current due to variations in the charge \(Q_H\) stored in the buffer layer.

Once one current component is defined, the other current component can be obtained from the current continuity equation:

\[
I_A = I_{n1} + I_{p1} = I_{n2} + I_{p2}
\]

**B. Validation of Fourier-based-solution Approach**

The Fourier-based solution for the drift region carrier distribution has been validated by comparison with ATLAS finite element simulations. The evolutions of the carrier distribution during IGBT inductive turn off predicted by ATLAS and by the model are shown in Fig. 3. The results show good agreement. The biggest discrepancy is in the oscillations of the charge profile at time 3.0 \(\mu s\). This is due to the truncation of the Fourier series retaining only the first eight terms.

**Fig. 3. Charge profiles during inductive turn off**

(a) ATLAS finite element simulation
(b) circuit simulator model (Note: Collector metal at 0 um)
C. Calculation of Voltage Across IGBT

The voltage drop in the IGBT can be described by the voltage drop across the p-emitter n-base junction, \( V_{J1} \), the voltage across the charge storage region, \( V_b \), and the voltage across the depletion region \( V_d \). In the on-state, with the accumulation of carriers in the region under the gate oxide, \( V_b \) is low and \( V_d \) is eliminated around the MOSFET channel.

The voltage drop in the drift region \( V_b \) is determined from the carrier profile, which must be sampled to calculate its resistance. The choice of the number of sampling points involves a trade-off between accurate determination of the carrier profile by use of a large number of sampling points, and simplicity and simulation speed obtainable by use of a small number of sampling points. Experience has shown that seven sampling points represent a good compromise. Choosing a larger number of sampling points brings only minor variations of the calculated voltage \( V_b \). Also, given the truncation of the Fourier series, there is little genuine gain from sampling at a high resolution. The drift region with seven equally spaced sampling points is shown in Fig. 4. The charge profile is assumed to be linear between adjacent sampling points. It is important to note that the sampling points move with respect to the device, following the charge profile. This gives a better definition of the charge profile when the drift region is partially depleted. If sampling points fixed with respect to the device were used, some of them would fall in the depleted region, where the carrier concentration is zero, and a smaller number of sampling points would be available to reconstruct the carrier profile in the undepleted drift region.

![Discretized carrier profile](image)

Fig. 4. Discretized carrier profile for simulation of \( V_b \) for a pin diode. The IGBT distribution is similar, but it drops to zero at the p-well.

D. Other Features of IGBT Model

The IGBT model has other features described in [1], such as quasi-2D models for the nonlinear device capacitances and temperature-dependent physical parameters.

III. Experimental Validation of Model

The physics-based IGBT model has been validated by comparison with experimental switching waveforms under inductive turn-off. Typical results are shown in Fig. 5 for a PT IGBT and in Fig. 6 for a Carrier Stored Trench Bipolar Transistor (CSTBT) at two different temperatures.

![Comparison between experiment and simulation of turn-off transient](image)

Fig. 5. Current fall and voltage rise during turn-off at 300 K comparing simulated and experimental waveforms. The horizontal scale is 200 ns/div. The current scale is in Amperes and the voltage scale is in Volts. Simulation results intentionally delayed for legibility

![CSTBT turn-off transient at 600V/100A under inductive load at different temperatures](image)

Fig. 6. CSTBT turn-off transient at 600V/100A under inductive load at different temperatures
IV. IGBT TURN ON BEHAVIOR

Turn on behavior is important since in modern IGBTs turn on losses can be comparable to turn off losses. Diode reverse recovery at turn on causes significant losses and frequently forces the designer to slow down the gate drive in order to mitigate ringing and EMI problems caused by snappy diode reverse recovery. Without diode reverse recovery, IGBT turn on could be made very fast, comparable to MOSFET turn on, and losses would be quite small. In [5] the special test circuit used provides inductive turn on without diode reverse recovery. Under those conditions the reported turn on losses are an order of magnitude smaller than turn off losses. The situation changes under the conventional clamped inductive condition with a real diode. In [6] extensive turn on and turn off losses are reported for both PT and NPT IGBTs. Turn on losses are generally larger than turn off losses, even not including the substantial diode turn off losses that occur during IGBT turn on. The conclusion is that overall switching losses during IGBT turn on can be significantly larger than those during IGBT turn off.

In an IGBT, the turn off behavior is predominantly a minority carrier phenomenon. The di/dt at turn off and the subsequent current tail are in large part determined by the amount of charge stored in the drift region. Turn off losses are only weakly dependent on the gate circuit [7].

On the other hand, IGBT turn on is largely a majority current phenomenon, determined by the MOSFET part of the IGBT. For this reason, turn on losses are very dependent on the gate drive circuit. A fast gate drive can significantly reduce losses. However, other considerations such as diode reverse recovery and short circuit behavior must be considered in the gate circuit design [7].

V. IGBT TURN ON UNDER RESISTIVE LOAD CONDITION

IGBT resistive turn on is described in [8-9]. The resistive load circuit is shown in Fig. 7. Inductance $L_p$ represents the parasitic loop inductance. Waveforms for the resistive turn on process are shown in Fig. 8. Since inductance $L_p$ is small, at all times equation (7) must be satisfied.

$$i_c = \frac{V_{cc} - V_{ce}}{R_{LOAD}}$$  \hspace{1cm} (7)

At time zero, voltage $V_{g0}$ goes high and the gate-emitter capacitance $C_{ge}$ starts charging. This first interval ends at time $t_{th}$, when the gate-emitter voltage reaches the threshold voltage. At this point, the MOSFET inside the IGBT starts conducting and the collector-emitter voltage $V_{ce}$ drops rapidly due to the voltage drop on resistor $R_{LOAD}$. The Miller capacitance $C_{gc}$ acts as a feedback to limit the gradient of the gate-emitter voltage. As a result, the gate-emitter voltage is approximately constant during this interval. The behavior is dominated by the MOSFET inside the IGBT.

After time $t_1$ the bipolar transistor inside the IGBT and the ohmic voltage drop in the drift region have a significant effect on the voltage and current waveforms. The drift region is initially depleted of charge and consequently it has a rather large resistance. Around time $t_1$ this voltage drop becomes a significant part of the device voltage $V_{ce}$ and it slows down the turn on process. Also the nonlinear Miller capacitance $C_{gc}$ becomes larger at lower voltages, contributing to the slowdown. As charge accumulates in the drift region, the drift region resistance drops. This phenomenon is called conductivity modulation and it is one of the main advantages of IGBTs over MOSFETs for high-voltage devices. So the voltage waveform in this interval is dominated by the charge dynamics in the drift region. This explains the slow evolution of the voltage. The gate-emitter voltage remains approximately constant also in this period due to the effect of the Miller capacitance. The gradient of the collector-gate voltage is significantly smaller than in the $t_{th} - t_1$ interval, but, as mentioned above, the Miller capacitance is significantly larger for low voltages. At time $t_2$ the gradient of voltage $V_{ce}$ becomes too small to pin the gate-emitter voltage, which starts charging up to $V_{g0}$.

In conclusion, in the IGBT resistive turn on two stages can be identified: a fast MOSFET stage in which the collector current increases rapidly, and a slow bipolar stage dominated by conductivity modulation of the drift region. In the second stage significant losses may occur.
Fig. 9 shows a comparison of experimental and simulation results for a NPT IGBT. Notice that there is good agreement in the gate-emitter voltage $v_{ge}$. The model captures the Miller effect and voltage $v_{ge}$ is constant in the interval $t_1$ - $t_2$. In the collector-emitter voltage and collector current waveforms some discrepancy can clearly be seen. There is good agreement at the beginning of the transition and at the end, but not in the middle. The discussion reported above on the turn on transition allows to pinpoint what the problem is. The model is accurate during the MOSFET part of the transition and during the final part of the bipolar part of the transition. The problem appears to be how the model calculates the drift region voltage drop in the initial stages of conductivity modulation. After some time, when the drift region resistance becomes small, the model provides a more accurate estimate and the agreement with the experiment is improved.

An attempt was made to improve the drift region voltage drop calculation by including more sampling points for the drift region in order to improve the accuracy of the drift region voltage drop calculation. A comparison of the original and modified model results is shown in Fig. 10a-b. While for the original model the discrepancy in the collector-emitter voltage started around 170V, the modified model remains accurate all the way down to 80V. Unfortunately at that point it starts diverging, and it can be seen that the estimated drift region voltage drop is too large. Fig. 11 shows more details of the simulation results using the modified model. It is clear that the MOSFET voltage drop dominates in the first phase and the drift region voltage drop dominates in the second phase. This accounts for the improvement in the modified model prediction for $v_{ce}$ between 170V and 80V.

A more careful examination of the drift region voltage drop calculation in the model is planned as future work to improve the IGBT model. As explained in the section describing the voltage drop calculation in the model, in order to have an accurate estimate of the drift region voltage drop it is important to retain enough terms of the carrier distribution Fourier series, otherwise the calculated carrier distribution may become oscillatory. This is evident from Fig. 3. Clearly it

Fig. 9. Resistive turn on of NPT IGBT. Comparison of experimental and simulation results. The x-axis scale is in $\mu$s. The y-axis scale on the left hand side is for the collector emitter voltage, the y-scale on the right hand side is for all the other waveforms.

Fig. 10. Resistive turn on of NPT IGBT. Comparison of experimental and simulation results using original model (a) and modified model (b). The x-axis scale is in $\mu$s. The y-axis scale on the left hand side is for the collector emitter voltage, the y-scale on the right hand side is for the collector current.

Fig. 11. Resistive turn on of NPT IGBT. Simulation results using the modified model. The contributions of the MOSFET voltage drop and of the drift region voltage drop to the collector-emitter voltage are shown. The x-axis scale is in $\mu$s. The y-axis scale is in Volt.
is also important to use enough sampling points for the drift region voltage drop calculation.

VI. IGBT TURN ON UNDER INDUCTIVE LOAD CONDITION

The IGBT turn on process under inductive load condition is described in [7]. This process is significantly more complicated than the resistive load case. The load current commutates from the freewheeling diode to the IGBT. The diode reverse recovery current flows through the IGBT causing a significant overcurrent. After the excess carriers in the diode drift region have been removed (or have recombined), the diode recovers its blocking capabilities and the diode voltage quickly increases. Depending on the diode construction, the diode recovery may be abrupt, causing significant overvoltage and EMI noise [10]. Soft recovery diodes are designed to mitigate this problem. Experimental turn on waveforms for a PT IGBT are shown in Fig. 12. The gate-emitter voltage exhibits the Miller effect discussed earlier. The collector current shows the diode reverse recovery current. Notice that the collector-emitter voltage exhibits an oscillation during the IGBT turn on. Comparison of simulation and experimental results is shown in Fig. 13. For the diode, a simple behavioral model is used with the parameters appropriately adjusted to approximate the measured reverse recovery current. Some discrepancies are evident. The biggest discrepancy is in the collector-emitter voltage. The simulation does not exhibit the significant oscillation of the experimental waveform. The reasons for this are under investigation. A possible explanation is that the voltage bump is due to resistive drop in the drift region due to the large collector current during the diode reverse recovery. As explained in the discussion of the resistive load case, the model needs improvement in the drift region voltage drop calculation. Another possible explanation is the one presented in Fig. 8 of [9]. The authors introduce a small (2nH) parasitic inductance \( L_E \) in the emitter of the IGBT as shown in Fig. 14. The reverse recovery current would cause a voltage drop on this inductance, momentarily reducing the gate-emitter voltage of the IGBT. This could create the voltage bump observed. Notice however that capacitances \( C_{ge} \) and \( C_{ce} \) are connected to the other side of inductor \( L_E \). This position of the inductor does not appear to be physically justified and appears to be a "behavioral" fix.

VII. CONCLUSION

The IGBT turn on behavior under resistive and inductive load has been examined in this paper. A physics-based IGBT model has been used for simulation. Simulation results have been compared with experimental results. Some discrepancy has been observed and possible causes identified.

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REFERENCES


