Practical SPICE Model for IGBT and PiN Diode Based on Finite Differential Method

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Abstract

In this paper, a practical SPICE model for an IGBT and a PiN diode is proposed based on the Finite Differential Method (FDM). Other than the conventional Fourier model and the Hefner model, the excess carrier distribution can be accurately solved by a fast FDM in the SPICE simulation tool. In order to improve the accuracy of the SPICE model, the Taguchi method is adopted to calibrate the extracted parameters. This paper presents a numerical modelling approach of an IGBT and a PIN diode, which are also verified by SPICE simulations and experiments.

Key words: Finite differential method, IGBT, Model, PiN diode, SPICE

NOMENCLATURE

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>A</td>
<td>Total die area of a diode (cm²).</td>
</tr>
<tr>
<td>A_i</td>
<td>Total die area of an IGBT (cm²).</td>
</tr>
<tr>
<td>C_{CG}</td>
<td>Gate-collector capacitance (nF).</td>
</tr>
<tr>
<td>C_{dep}</td>
<td>Depletion capacitance (nF).</td>
</tr>
<tr>
<td>C_{GE}</td>
<td>Gate-emitter capacitance (nF).</td>
</tr>
<tr>
<td>C_{ox}</td>
<td>Oxide capacitance (nF).</td>
</tr>
<tr>
<td>C_{res}</td>
<td>Reverse transmission capacitance (nF).</td>
</tr>
<tr>
<td>D</td>
<td>Ambipolar diffusivity (cm²/s).</td>
</tr>
<tr>
<td>D_n</td>
<td>Electron diffusivity (cm²/s).</td>
</tr>
<tr>
<td>D_p</td>
<td>Hole diffusivity (cm²/s).</td>
</tr>
<tr>
<td>e_{si}</td>
<td>Dielectric constant of silicon (F/cm).</td>
</tr>
<tr>
<td>h_n</td>
<td>Electron recombination coefficient (cm⁴/s).</td>
</tr>
<tr>
<td>h_p</td>
<td>Hole recombination coefficient (cm⁴/s).</td>
</tr>
<tr>
<td>I_C</td>
<td>Collector current of an IGBT (A).</td>
</tr>
<tr>
<td>I_{CG}</td>
<td>Collector-gate current in an IGBT (A).</td>
</tr>
<tr>
<td>I_D</td>
<td>Total anode current of a diode (A).</td>
</tr>
<tr>
<td>I_{disp1}</td>
<td>Displacement current at x_l (A).</td>
</tr>
<tr>
<td>I_{disp2}</td>
<td>Displacement current at x_r (A).</td>
</tr>
<tr>
<td>I_G</td>
<td>Drive current of an IGBT (A).</td>
</tr>
<tr>
<td>I_{GE}</td>
<td>Gate-emitter current in an IGBT (A).</td>
</tr>
<tr>
<td>I_{mos}</td>
<td>MOS channel region current (A).</td>
</tr>
<tr>
<td>I_{n1}</td>
<td>Electron current at x_l (A).</td>
</tr>
<tr>
<td>I_{n2}</td>
<td>Electron current at x_r (A).</td>
</tr>
<tr>
<td>I_{p1}</td>
<td>Hole current at x_l (A).</td>
</tr>
<tr>
<td>I_{p2}</td>
<td>Hole current at x_r (A).</td>
</tr>
<tr>
<td>J_T</td>
<td>Current density of a diode (A/cm²).</td>
</tr>
<tr>
<td>K_p</td>
<td>MOS channel transconductance Ω⁻¹.</td>
</tr>
<tr>
<td>N_B</td>
<td>Base impurity doping concentration (cm⁻³).</td>
</tr>
<tr>
<td>p</td>
<td>Carrier density of drift region (cm⁻³).</td>
</tr>
<tr>
<td>p_l</td>
<td>Carrier density at x_l (cm⁻³).</td>
</tr>
<tr>
<td>p_r</td>
<td>Carrier density at x_r (cm⁻³).</td>
</tr>
<tr>
<td>P_{ave}</td>
<td>Average carrier density of drift region (cm⁻³).</td>
</tr>
<tr>
<td>q</td>
<td>Electron charge (e).</td>
</tr>
<tr>
<td>t</td>
<td>Time (us).</td>
</tr>
<tr>
<td>t_{off}</td>
<td>Turn-off delay time (ns).</td>
</tr>
<tr>
<td>t_{on}</td>
<td>Turn-on delay time (ns).</td>
</tr>
</tbody>
</table>
**I. INTRODUCTION**

An insulated gate bipolar transistor (IGBT) can be equivalent to a giant transistor (GTR) driven by a metal-oxide semiconductor field effect transistor (MOSFET) in the physical structure. Therefore, it has the advantages of both GTRs and MOSFETs, such as high input impedance and low on-state voltage. With increments in its voltage and current ratings, the application range can be extending from medium power applications to high power applications [1].

The authors of [2] presented a 6.5 kV commercially available IGBT module in a boost circuit application switching at 9 kHz, and a 7 kA IGBT module was fabricated in [3]. Improvements in packing materials and structure design enable IGBTs to be operated at higher frequencies. Thus, the authors of [4] provided an ultra-thin punch-through IGBT with a blocking voltage of 650V, which is able to drive DC-DC converter at 200 kHz.

Since the invention of the first IGBT in 1982 [5], various of modeling methods have been proposed from different aspects and with different objectives. However, depending on the modeling method used, the models can be divided into two categories, numerical models and behavioral models.

Behavioral models concentrate on IGBT behavior without considering their physical mechanism. A curve-fitting method was used in [6] to calculate switching losses. However, this model cannot demonstrate dynamic behavior. The authors of [7] used a Hammerstein model to describe the static characteristics and the switching characteristics of an IGBT.

There are three approaches based on this method: (1), Hefner model [8], Fourier-based model [9] and Laplace transformation model [10]. The Hefner model assumes a linear distribution of excess carriers and describes them as Equ. (2). This method is appropriate for a planar IGBT. However, it is not appropriate for a trench-field-stop IGBT and a PIN diode. The Fourier-based model can simulate all kinds of IGBTs. However, its iterations are too complex, which can lead to some problems in terms of convergence. The drift region excess carrier distribution of a Fourier-based model can be calculated by Equ. (3). A Laplace transformation model can solve the ADE in the frequency domain. However, it is not suitable for changes of the drift region boundary.

![Fig. 1. Configuration of a Hammerstein model.](image)

**II. SIMULATION OF THE DRIFT REGION**

### A. Finite Differential Method Implementation in SPICE

A time domain based partial differential equation solution was proposed in [11]. According to this method, the drift...
region can be finite differenced in \( k \) equal parts, \( x_1, \ldots, x_j, \ldots x_k \). Hence, the orders of the ADE can be reduced by Equ. (4) and Equ. (5) in part \( j \), and the partial differential equation can be simplified as Equ. (6).

\[
\frac{\partial p_j}{\partial x} = \frac{p_{j+1} - p_j}{\Delta x} \quad (4)
\]
\[
\frac{\partial^2 p_j}{\partial x^2} = \frac{p_{j+1} + p_{j-1} - 2p_j}{\Delta x^2} \quad (5)
\]
\[
\frac{\partial p_j}{\partial t} = \frac{D}{\Delta x^2} p_{j+1} - \left( \frac{2D}{\Delta x^2} + \frac{1}{\tau_{HL}} \right) p_j + \frac{D}{\Delta x^2} p_{j-1} \quad (6)
\]

Equ. (6) can be solved by a SPICE simulation, where it is replaced by a simple network of resistors and capacitors driven by voltage controlled current sources. As showed in Fig. 2, \( p_i \) is replaced by the node voltage \( e_j \), and the differential of \( p_i \) is equivalent to the charge current of \( C \).

The value chosen for \( C \) and \( R \) is unimportant as long as the \( R-C \) time constant is much larger than the largest time of the ADE interest. In addition, a large number (larger than \( 10^{15} \)) or a non-zero small number (less than \( 10^{-6} \)) results in convergence problems in SIMetrix.

Based on this principle, the values of \( R \) and \( C \) are chosen as 1 and \( 10^{-6} \), respectively. Hypothetically, the drift region is finite differenced in \( k \) equal parts. A higher \( k \) can result in a higher calculation accuracy. However, the solving process is more complex and the simulation speed increases. Hence, a trade-off between calculation accuracy and simulation speed should be made. As a consequence, the stage number is chosen as 5.

Taking the boundary conditions of Equ. (7) and Equ. (8) into consideration, the excess carrier distribution can be calculated by Equ. (9).

\[
\frac{-p_{j+1} + p_{j-1}}{\Delta x} + \frac{I_D - I_{disp1}}{2qAD_p} - \frac{h_p}{D} p_j^2 = 0 \quad (7)
\]
\[
\frac{-p_k + p_{k-1}}{\Delta x} + \frac{I_D - I_{disp2}}{2qAD_n} - \frac{h_n}{D} p_k^2 = 0 \quad (8)
\]

\[
\begin{bmatrix}
  0 & d_{11} & d_{12} & 0 & 0 & 0 & \ldots & p_1 & b_{h1} & \ldots \\
  p_2 & a_{21} & a_{22} & a_{23} & 0 & 0 & \ldots & p_5 & 0 & \ldots \\
  p_3 & 0 & a_{32} & a_{33} & a_{34} & 0 & \ldots & p_4 & 0 & \ldots \\
  p_4 & 0 & 0 & a_{43} & a_{44} & a_{45} & \ldots & p_5 & 0 & \ldots \\
  0 & 0 & 0 & d_{54} & d_{55} & d_{56} & \ldots & p_5 & b_{h2} & \ldots
\end{bmatrix} \quad (9)
\]

"
and their drift region contains a large number of electrons and holes which makes it a low resistance state. When a PiN diode is switched from the conduction mode to the reverse-blocking mode, the excess carrier density of the drift region decreases rapidly due to the termination of the injection. The stored charge within the drift region of the PiN diode must be extracted before it can support high voltages. This process is illustrated in Fig. 4 and Fig. 5. This phenomenon is referred to as reverse recovery.

### III. DIODE MODEL IMPLEMENTATION

To implement the FDM model for a PiN diode, the basic one-dimension diode body is divided into three parts, as shown in Fig. 6. The performance characteristics of the PiN diode depend mainly on the chip geometry and the processed semiconductor material in the drift region [13]. When the PiN diode is forward biased, holes and electrons are injected into the drift region. This charge does not recombine instantaneously. Instead it has a finite lifetime (\(\tau_{HL}\)) in the drift region. If the PiN diode is reverse biased, there is no stored charge in the drift region, which behaves like a parallel \(R-C\) network.

#### A. Physics Method for PiN Diode Modeling

The on-state voltage drop of a PiN diode \(V_{AK}\) consists of three parts, which are the junction voltages \(V_j\), the voltage of the drift region \(V_d\) and the voltage of the depletion region \(V_d\).

\[
V_{AK} = V_j + V_d - V_{de}
\]  

(13)

Neglecting the distributional effects in the drift region, \(V_j\) and \(V_d\) can be described by a simple quasi-static model as Eq. (14) and Eq. (15). The voltages across the P+/N- junction and the N+/N- junction are determined by the injected minority carrier density and the majority carrier density, respectively. More importantly, Eq. (14) can be derived under the charge neutrality condition \(p(x)=n(x)\).

In Eq. (15), \(I_{p1}/A_{sat}\) and \(I_{n1}/A_{sat}\) are far less than \(qNB\). Consequently, neglecting them can improve the calculating speed and the convergence property of the FDM model.

\[
V_j = V_T \ln \left( \frac{P_r P_i}{n_i^2} \right)
\]  

(14)

\[
V_d = \frac{qN_B + I_{p1} / A_{sat}}{2e_{sd}} x_f^2 + \frac{qN_B + I_{n2} / A_{sat}}{2e_{sd}} (W_B - x_f)^2
\]  

(15)

\[
V_B = \frac{I_D}{Aq \mu_n N_B + (\mu_n + \mu_p)P_{base}} + 2V_j - \frac{D_p}{D_n + D_p} \ln \left( \frac{P_r}{P_i} \right)
\]  

(16)

#### B. PiN Diode Parameters Extraction

The first parameter that needs to be extraction is the chip area \(A\), which is closely related to the rated current. The approach to extract \(A\) is to directly measure the diode area or to obtain it from a datasheet.

The parameter \(\tau_{HL}\) must be determined from the diode turn-off waveform such as the one shown in Fig. 7. Under this circumstance, \(\tau_{HL}\) can be calculated by Eq. (17) [14].

\[
I_{RM} = \alpha(\tau - \tau_{r})[1 - \exp(\frac{T_I}{T})]
\]  

(17)

Where \(\alpha\) is the current slope from \(T_0\) to \(T_1\). \(T_0\) is the zero-crossing point during the turn-off transient, and \(T_1\) is the moment the reverse recovery current reaches its peak value. \(\tau_{r}\) is the reverse recovery time constant, which can be measured from the current waveform.

Another important parameter is the drift region width \(W_d\), which can influence the breakdown voltage of bipolar semiconductor devices. The approach to extracting \(W_d\) is based on Eq. (18) [15], and the PiN diode parameters extraction results are listed in Table I.

\[
V_{BB} = \frac{b W_B}{\ln(a W_B)}
\]  

(18)

Where \(V_{BB}\) is the avalanche breakdown voltage. In addition, \(a\) and \(b\) represent two constants that are related to the semiconductor technology.
Practical SPICE Model for IGBT and PiN Diode Based on …

Fig. 7. Turn-off waveform of a PiN diode.

TABLE I
PARAMETERS EXTRACTION RESULTS OF A DIODE

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Chip area A</td>
<td>0.376cm²</td>
</tr>
<tr>
<td>High-level lifetime τHL</td>
<td>0.12μs</td>
</tr>
<tr>
<td>Drift region width W_B</td>
<td>139.7μm</td>
</tr>
<tr>
<td>Recombination coefficient h_n, h_p</td>
<td>10⁻¹⁴cm⁴/s</td>
</tr>
<tr>
<td>Drift region doping concentration N_d</td>
<td>10⁴cm²/s</td>
</tr>
</tbody>
</table>

TABLE II
SIMULATION PARAMETERS

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Simulation time</td>
<td>31.5μs</td>
</tr>
<tr>
<td>Real time</td>
<td>203ms</td>
</tr>
<tr>
<td>Time step</td>
<td>178.1μs</td>
</tr>
<tr>
<td>Total points</td>
<td>470</td>
</tr>
<tr>
<td>Total iterations</td>
<td>2.205k</td>
</tr>
</tbody>
</table>

Fig. 9. Double-pulse test with the proposed PiN model (V_gon=15V, V_goff=-9V, R_g=7.5Ω, L=42.3 mH). (a) Under a DC bus voltage of 100V. (b) Under a DC bus voltage of 200V.

C. PiN Diode Model Verification

With all of the above equations, the PiN diode model is simulated in SIMetrix. SIMetrix is a mixed-signal circuit simulator with its core algorithms based on the SPICE program. The simulation parameters are listed in Table II.

To verify the simulation results of SIMetrix, a double-pulse test circuit was developed as shown in Fig. 8, and the DUT (Device Under Test) is an IRD3CH53DB6. A 42.3 mH inductor served as a load and the value of the driver resistance R_g was chosen as 7.5 Ω. In order to avoid the spurious triggering caused by the Miller capacitance, V_goff was set at -9 V. Fig. 9 shows a comparison of simulation and experimental results.

IV. IGBT MODEL IMPLEMENTATION

The basic unit cell structure of an IGBT is shown in Fig. 10. The N⁺ zone is called the source region, and the electrode attached to it is called the source electrode. The N plus area is called the drift region. The control area of the IGBT is the gate area, and the channel is formed near the boundary of the gate.

A. Physics Method for IGBT Modeling

The basic IGBT dynamic model consists of three state equations: the current continuity equation, the voltage drop equations and the IGBT driver equations. The current continuity equation governing the behavior of IGBT is described as Eqn. (19).

\[
I_C = I_n1 + I_p1 = I_{n2} + I_{p2} + I_{disp2} + I_{CG} \tag{19}
\]

Where \(I_{n1}=qA\phi h \gamma_r\), \(I_{n2}=I_{max}\) and the specific descriptions of \(I_{disp2}\) and \(I_{CG}\) are described in [16].
The voltage drop of an IGBT is comprised of three parts: the voltage across the junctions $J_1$, the voltage across the depletion region $V_{d2}$, and the voltage across the drift region $V_B$. Similarly, the three voltages can be describe as Equ. (20), Equ. (21) and Equ. (22).

\[
V_{j1} = 2V_T \ln \left( \frac{p_l}{n_i} \right) \tag{20}
\]

\[
V_{d2} = qN_B + \frac{I_{n2}}{2e_{si}} \left( W_f - x_j \right)^2 \tag{21}
\]

\[
V_B = \frac{I_C}{Aq \mu_n N_B + (\mu_n + \mu_p)P_{base}} + 2V_T \frac{D_p}{D_n + D_p} \ln \left( \frac{p_r}{p_l} \right) \tag{22}
\]

A typical IGBT application circuit is illustrated in Fig. 11. According to Kirchhoff’s law, the current flowing into the gate terminal is equal to the current flowing out of it. Consequently, $I_G$ and $I_{CG}$ can be described as Equ. (23) and Equ. (24), respectively. Combining Equ. (23) with Equ. (24), the IGBT driver equation is derived as Equ. (25). The MOS part of the IGBT can be directly implemented with a PSPICE MOSFET model.

\[
I_G = I_{CG} + I_{GE} = C_{CG} \frac{dV_{GE}}{dt} + C_{GE} \frac{dV_{GE}}{dt} \tag{23}
\]

\[
I_{CG} = C_{CG} \left( \frac{dV_{GE}}{dt} - \frac{dV_{GE}}{dt} \right) \tag{24}
\]

\[
\frac{dV_{GE}}{dt} = \frac{I_G}{C_{GE} + C_{CG}} + \frac{C_{CG}}{C_{GE} + C_{CG}} \frac{dV_{CE}}{dt} \tag{25}
\]

**B. Miller Plateau Implementation**

The Miller plateau caused by Miller capacitance can lead to a loss of control over the turn-on $\text{d}i/\text{d}t$ [17] and snap-off during the turn-off process, which can result in device failures. The Miller capacitance forms from an overlap of the gate metallization and the N-drift region [18]. It can be described as a voltage-controlled capacitance that is not available for SPICE simulations.

In this paper, voltage-controlled current sources are used to simulate the charge process of $I_{CG}$. The moment the Miller Plateau occurs is related to the threshold voltage of the elemental MOSFET. This phenomenon is shown in Fig. 13. In this method, $C_{aux}$ and $C_{dep}$ are equivalent to CAP and DCAP, respectively. By changing the null-bias capacitance CJO and the junction electric potential $V_J$ of the DCAP, different Miller plateau characteristics can be simulated, as shown in Fig. 14. In addition, Fig. 15 shows a comparison of simulation and experimental results, under the same condition with the diode test.

**C. IGBT Parameters Extraction**

The parameters that need to be extracted for the modeling
of an IGBT can be divided into five categories: the MOS part parameters, IGBT structure parameters, IGBT drift region parameters, IGBT lifetime parameters and IGBT recombination coefficients.

The structure parameters include the device area $A$ and the oxide capacitance $C_{ox}$ of the IGBT. The device area can be measured or obtained directly from a datasheet. The oxide capacitance is a part of the Miller capacitance. It can be acquired from the $C_{res}$ curve in the datasheet. When $C_{res}$ reaches its maximum value, the depletion layer has not formed yet ($C_{dep}=\infty$). This means the value of $C_{ox}$ is the same as $C_{res}$.

Other than the PiN diode, the drift region parameter $W_i$ is different between Non-Punch Through (NPT) IGBTs and Punch Through (PT)/Field Stop (FS) IGBTs. Since the electric field in the drift region of an NPT IGBT is a triangular distribution, $W_i$ can be derived by Eqn. (26). As for a PT/FS IGBT, the electric field of the drift region is a trapezoidal distribution, and the drift region width of the IGBT can be calculated by Eqn. (27).

$$W_i = \frac{E_{si}E_c}{qN_B}$$  \hspace{1cm} (26)

$$W_i = \frac{E_{si}}{qN_B(E_c - \frac{E_c^2}{2qN_BV_{BR}/E_{si}})}$$  \hspace{1cm} (27)

Where $E_c$ is the typical value of the silicon-based electric field, and $V_{BR}$ is the avalanche breakdown voltage of the IGBT.

In order to extract a high-level lifetime $\tau_{HL}$ of the IGBT, the current tail in the turn-off transient under an inductive load needs to be measured. According to the method in [19], $\tau_{HL}$ is the time constant of the current tail as shown in Fig. 16.

The MOS part of the IGBT can be directly implemented with a PSPICE MOSFET model, the parameters of which include the MOSFET threshold voltage $V_{th}$ and transconductance $K_p$. $V_{th}$ is the threshold voltage of the MOSFET which means the gate voltage at the critical conduction moment. $K_p$ is the channel transconductance of the MOSFET, which represents the ability of the gate voltage to control the collector current. These two parameters can be extracted from Eqn. (28), and the IGBT parameters extraction results are listed in Table III.

$$V_{th} = \frac{1}{K_p} \left( V_{GS} - V_{th} \right) \left( V_{DS} - \frac{V_{DS}^2}{2} \right) \text{ for } V_{DS} < V_{GS} - V_{th}$$

$$V_{th} = \frac{K_p(V_{GS} - V_{th})^2}{2} \text{ otherwise}$$  \hspace{1cm} (28)

### D. Taguchi Method for Calibration

Due to measurement error, the extracted parameters cannot be completely accurate. Consequently, the Taguchi method was adopted to calibrate the model based on the FDM. The Taguchi method is a statistical method, developed by Genichi Taguchi to improve the quality of manufactured goods. More recently, it has been applied to engineering [20].

In the Taguchi method, quantitative indexes are designed to compare the losses of different experimental groups. The purpose of the FDM model is to simulate the switching performances of an IGBT. As a result, a Taguchi loss function is designed as Eqn. (29).

The calibration parameters are the threshold voltage $V_{th}$, transconductance $K_p$, oxide capacitance $C_{ox}$ and drift region width of the IGBT $W_i$. The extracted parameters are taken as median values. The threshold values are generated by 90% and 110% of median values, respectively. A L9 (3)$^4$ orthogonal table is set as shown in Table IV, and the results of Taguchi experiments are displayed in Table V. Calibration results of the Taguchi method are illustrated in Fig. 17.

$$E_{loss} = \frac{(t_r - t_{r \_exp})^2}{t_{r \_exp}} + \frac{(t_{don} - t_{don \_exp})^2}{t_{don \_exp}} + \frac{(t_{doff} - t_{doff \_exp})^2}{t_{doff \_exp}} + \frac{(t_f - t_{f \_exp})^2}{t_{f \_exp}}$$  \hspace{1cm} (29)
TABLE V
RESULT OF THE TAGUCHI METHOD

<table>
<thead>
<tr>
<th>No</th>
<th>$V_{th}$</th>
<th>$K_p$</th>
<th>$C_{ox}$</th>
<th>$W_i$</th>
<th>$E_{loss}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1</td>
<td>1</td>
<td>1</td>
<td>1</td>
<td>0.065</td>
</tr>
<tr>
<td>2</td>
<td>1</td>
<td>2</td>
<td>2</td>
<td>2</td>
<td>0.062</td>
</tr>
<tr>
<td>3</td>
<td>1</td>
<td>3</td>
<td>3</td>
<td>3</td>
<td>0.075</td>
</tr>
<tr>
<td>4</td>
<td>2</td>
<td>1</td>
<td>2</td>
<td>3</td>
<td>0.014</td>
</tr>
<tr>
<td>5</td>
<td>2</td>
<td>2</td>
<td>3</td>
<td>1</td>
<td>0.051</td>
</tr>
<tr>
<td>6</td>
<td>2</td>
<td>3</td>
<td>1</td>
<td>2</td>
<td>0.104</td>
</tr>
<tr>
<td>7</td>
<td>3</td>
<td>1</td>
<td>3</td>
<td>2</td>
<td>0.008</td>
</tr>
<tr>
<td>8</td>
<td>3</td>
<td>2</td>
<td>1</td>
<td>3</td>
<td>0.046</td>
</tr>
<tr>
<td>9</td>
<td>3</td>
<td>3</td>
<td>2</td>
<td>1</td>
<td>0.086</td>
</tr>
<tr>
<td>K1</td>
<td>0.202</td>
<td>0.087</td>
<td>0.214</td>
<td>0.201</td>
<td></td>
</tr>
<tr>
<td>K2</td>
<td>0.169</td>
<td>0.159</td>
<td>0.163</td>
<td>0.175</td>
<td></td>
</tr>
<tr>
<td>K3</td>
<td>0.139</td>
<td>0.265</td>
<td>0.134</td>
<td>0.135</td>
<td></td>
</tr>
</tbody>
</table>

Fig. 17. Calibration results of the Taguchi method.

Fig. 18. Double-pulse test with the proposed IGBT model.

E. IGBT Model Verification

With all of the above equations and parameters, an IGBT model is simulated in SiMetrix. The simulation conditions are the same as those of the PiN diode. To verify the simulation results of the SiMetrix, a double-pulse test circuit was developed to compare the switching transients and the Miller plateau with the simulation results as shown in Fig. 18. The DUT (Device Under Test) is an IRG8CH97K10F. Fig. 19 illustrates the switching transients of the proposed IGBT.
This paper proposes a practical SPICE model for an IGBT and a PiN diode based on Finite Differential Method. The model relates well to experiment results during turn-on and turn-off transients. The Miller plateau and reverse characteristics of the PiN diode are also included. The key point of this modeling is to solve the Ambipolar Diffusion Equation in a SPICE simulation. Taking into account the charge dynamics within the base region of the device, the presented model provides great improvements in terms of model speed and accuracy.

APPENDIX

The FDM part of the IGBT SPICE model proposed in this paper is:

```
.PARAM C11 = {((N-1)/Wd)**2*D*1.0E2}
.PARAM C12 = {-2*C11-1/Td}
.PARAM C13 = {C11}
.PARAM E11 = {-(N-1)/Wd}
.PARAM E12 = {-E11}
.PARAM E13 = {1/(2*q*Ad*Dp)*1E-16}
.PARAM E14 = {-hp*1E8/D}
.PARAM E15 = {1/(2*q*Ad*Dn)*1E-16}

*******************************************************************************************
*************** Differential Coefficients ****************************
*******************************************************************************************

R1 1 12 RX
G1 12 1 POLY(3)  1 12 2 12 4 12 0 E11 E12 E13 E14

R2 2 12 RX
C2 2 12 CX IC = 0 BRANCH={IF(ANALYSIS=2,1,12)}
G2 12 2 POLY(3)  1 12 2 12 3 12 0 C11 C12 C13

R3 3 12 RX
C3 3 12 CX IC = 0 BRANCH={IF(ANALYSIS=2,1,12)}
G3 12 3 POLY(3)  2 12 3 12 4 12 0 C11 C12 C13

R4 4 12 RX
C4 4 12 CX IC = 0 BRANCH={IF(ANALYSIS=2,1,12)}
G4 12 4 POLY(3)  3 12 4 12 5 12 0 C11 C12 C13

R5 5 12 RX
G5 12 5 POLY(3)  5 12 4 12 15 12 0 E11 E12 E15 E14

******************************************************************************* Voltage of IGBT ****************************
*******************************************************************************

EJ1 11 103 VALUE ={LIMIT( 2*VT*LN(V(1,12)/(ni/1e12) ) ,0,10)}

EJ2 103 104 VALUE = {LIMIT(V(14,12) *0.6E5/(1.4E5 + 1850/N*(V(1,12)+V(2,12)+V(3,12)+V(4,12)+V(5,12)))+2*VT *Dp/(Dn+Dp)*LN(V(5,12)/V(1,12)+10),0,10)}

EJ3 104 105 VALUE = {V(202,12)}

*******************************************************************************************
*************** Current of IGBT ****************************
*******************************************************************************************

EP1 14 12 VALUE={I(EJ1)}
RP1 14 12 1
EP2 15 12 VALUE={I(EJ1)}
RP2 15 12 1

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REFERENCES


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