A Parametric Device Study for SiC Power Electronics

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Abstract: Materials and device researchers build switching devices for the circuits researchers to use in their circuits, but they rarely know how and where the devices are going to be used. The circuits people, including power electronics researchers, take the devices as black boxes and use them in their circuits not knowing much about the inside of the devices. The best way to design optimum devices is an interactive design where people designing and building the devices have a close interaction with the people who use them.

This study covers the circuit aspects of the SiC power device development. As a contribution to the above-mentioned interactive design, in this paper, the device parameters, which need to be improved in order to design better devices, will be discussed.

I. INTRODUCTION

Typically, power electronics researchers have to choose off-the-shelf power devices with the specifications best fit for their applications. They, usually, do not have a say about how they would like the device parameters be changed.

Materials and device researchers build switching devices for the power electronics researchers to use in their circuits, but they rarely know how and where the devices are going to be used.

As represented in Fig. 1, a “barrier” exists between the people who design and build power devices and the people who use them in their circuits and systems. Close interaction between the both sides of the barrier is needed to obtain the most performance for devices and systems. With this interaction, the design loop will be closed and the possibility for building application specific optimum power devices will arise.

Recently, a significant increase in the interest of silicon carbide (SiC) power devices has occurred because of their system level benefits. In the literature, SiC research is mainly concentrated on the materials and devices aspects [1, 2]. Recently, more circuit applications [3, 4] are being published.

Moreover, the system level benefits of SiC are also being evaluated in some recent papers [5-7]. However, SiC power devices are still in their development stage; therefore, this is a good opportunity at this time to close the loop.

At Oak Ridge National Laboratory (ORNL), a team of materials, device, and power electronics researchers are working together with the University of Tennessee, Auburn University, and Vanderbilt University to build application specific optimum SiC power MOSFETs. This paper will summarize some of this work.

II. APPLICATIONS

This paper is a part of a study where system impact of SiC power electronics on hybrid electric vehicle (HEV) applications was investigated [5-8]. In the mentioned study, two HEV power converters were identified, modeled, and simulated to show the system level benefits of SiC power electronics quantitatively. The two selected applications were a dc-dc power supply and a traction drive.

The dc-dc power supply shown in Fig. 2 is an isolated full-bridge dc-dc converter, which is selected mostly because of its high frequency transformer, which provides isolation and additional taps in the secondary to feed more than one converter.

The main traction drive shown in Fig. 3 uses most of the power in an HEV when the vehicle is in motion. A traction drive consists of a battery feeding a three-phase induction machine through a three-phase inverter. Because of the cooling requirements of the power devices in the inverter, usually a large heatsink is required.

In an HEV, any reduction in volume and weight of any component will benefit the efficiency of the vehicle. Because SiC devices can operate at higher temperatures and they have...
lower losses, the heatsink volume and weight can be reduced if all SiC devices are used in all HEV power converters.

The simulation results of these converters have shown on average 30% decrease in weight and volume of the heatsink and a 5-10% increase in the efficiency. Improving the related device parameters can increase these further.

In the next two sections, these parameters will be identified for SiC Schottky diodes and MOSFETs and then necessary suggestions for improvement will be stated.

Note that all these modification suggestions also apply to Si devices, but the main focus of this study is given to SiC power devices.

III. DIODES

Some important diode parameters for power electronics systems are the breakdown voltage, on resistance, built-in voltage drop, peak reverse recovery current, and reverse recovery time.

A. Conduction Loss Parameters

1) Traction drive

A diode conduction loss expression for a traction drive inverter shown in Fig. 3 has been derived in [5], and it is repeated below for convenience.

$$P_{\text{cond,04}} = I^2 \cdot R_D \cdot \frac{\pi}{3} \cdot \frac{1}{\pi} \cdot M \cos \phi \phi + I \cdot V_D \cdot \frac{\pi}{2} \cdot \frac{1}{\pi} \cdot M \cos \phi \phi$$

(1)

where

- $I$ is the current through the diode,
- $M$ is the modulation index for sinusoidal PWM,
- $\phi$ is the power factor angle,
- $R_D$ is the diode series resistance, and
- $V_D$ is the diode built-in voltage.

This equation consists of two parts, loss associated with the on resistance, $R_D$ and loss associated with the built-in voltage drop, $V_D$. Diodes with lower $R_D$ and $V_D$ would be preferable, but these parameters depend on similar device parameters e.g. both of these parameters depend on the doping densities. Higher doping density means lower $R_D$ but higher $V_D$ and lower breakdown voltage, $BV$; therefore, both $R_D$ and $V_D$ cannot be lowered at the same time, i.e. a trade-off is required.

Consider a 4H-SiC Schottky diode with a $BV$ of more than 500V for a traction drive.

$$BV = \frac{\varepsilon_c E_c^2}{2qN_d} = 1.3511 \times 10^{11} \frac{N_d}{N_d} > 500V$$

(2)

where $BV$ is the breakdown voltage

- $\varepsilon_c$ is the permittivity
- $E_c$ is the critical electric breakdown field
- $q$ is the electron charge
- $N_d$ is the doping density

The maximum doping density value to sustain the chosen $BV$ is calculated above. The resistance value corresponding to this $N_d$ is the minimum $R_D$. It cannot be decreased with doping any further; however, the doping density can still be selected lower than this value, which would increase $BV$ and $R_D$, and decrease $V_D$. Then, the question is: Can modifying $V_D$ and $R_D$ decrease the conduction losses?

To answer this question, it is required to find how much changes in $R_D$ and/or $V_D$ will affect the conduction losses.

$$I^2 \cdot R_D \cdot \frac{\pi}{3} \cdot \frac{1}{\pi} \cdot M \cos \phi \phi > I \cdot V_D \cdot \frac{\pi}{2} \cdot \frac{1}{\pi} \cdot M \cos \phi \phi$$

(3)

Rearranging terms and assuming $I \neq 0$,

$$I \cdot \frac{\pi}{3} \cdot \frac{1}{\pi} \cdot M \cos \phi \phi > \frac{V_D}{R_D}$$

(4)

$$I \cdot f(M \cos \phi) > \frac{V_D}{R_D}$$

where

$$f(M \cos \phi) = \frac{\pi}{3} \cdot \frac{1}{\pi} \cdot M \cos \phi \phi$$

$M$ is the modulation index, which varies between 0 and $4/\pi$ (square wave operation), and $\cos \phi$ is the power factor, which varies between 0 and 1. The power factor of an induction machine is always lagging; for this example calculation, it is assumed to be 0.9 at rated load.

$$0 \leq M \leq \frac{4}{\pi} \quad \text{and} \quad 0 \leq \cos \phi < 0.9$$

Then,
\[ 0 \leq M \cos \phi < \frac{3.6}{\pi} \]  
and \( f(M \cos \phi) \) varies between 0.787 (no-load) and 0.215 (rated load) as shown in Fig. 4.

At first glance, it might seem that because the \( I \cdot V_D \) multiplier is larger than the \( I^2 \cdot R_D \) multiplier at all \( M \cos \phi \) values in Fig. 4, the \( V_D \) losses should always be higher. This observation would have been true if and only if \( V_D \) and the \( I \cdot R_D \) product were equal. This, however, is not the case, and that is why, all three of these variables are included in (4) to find under what conditions, what part of the conduction losses is higher. The following example illustrates how to make use of (4).

For a particular hybrid electric vehicle traction drive, the rated peak machine current is 136.28A, which makes \( I \cdot f(M \cos \phi) = 136.28 \cdot 0.215 = 29.3A \). Ignoring the off condition, the minimum device current is the magnetizing current, which is 71A. During the magnetizing current operation, the phase angle is almost \( \pi/2 \) radians and the power factor is almost zero, then \( I \cdot f(M \cos \phi) = 71 \cdot 0.787 = 55.9A \)

Considering (4), the following are some recommendations to maximize the efficiency of a SiC diode in a traction drive application:

1) If \( \frac{V_D}{R_D} \) is greater than 29.3, then the \( R_D \) losses are higher at all times, keep the doping density and \( R_D \) constant because decreasing \( R_D \) means decreasing \( B_V \), which would limit the device’s application.

2) If \( \frac{V_D}{R_D} \) is less than 55.9, then the \( V_D \) losses are higher at all times, decrease the doping density so that \( V_D \) will be smaller.

3) If \( \frac{V_D}{R_D} < 55.9A \), then the average current of operation will determine the recommended doping density as follows:

a) A drive working close to its rated current value uses the condition
\[ 29.3A < \frac{V_D}{R_D} \] where \( V_D \) losses are higher, decrease the doping density so that \( V_D \) will be smaller.

b) A drive working at light current loads uses the condition
\[ \frac{V_D}{R_D} < 55.9A \] where \( R_D \) losses are higher, keep the doping as it is because decreasing \( R_D \) means decreasing \( B_V \), which would decrease the voltage blocking capability of the device.

Fig. 5 displays the above statements on an \( R_D - V_D \) plane. A commercial SiC Schottky diode \( I-V \) characteristics are obtained at different temperatures. From these characteristics, \( V_D \) and \( R_D \) values of the diode are calculated. These values are tabulated in Table I and shown as a small rectangular area in

![Fig. 4. The \( R_D - V_D \) plane for the traction drive.](image)
TABLE I
SiC DIODE PWL MODEL PARAMETERS AND \( V_{D}/R_{D} \) RATIO

<table>
<thead>
<tr>
<th>( T_{oven} ) °C</th>
<th>( R_{D} ) in Ω</th>
<th>( V_{D} ) in V</th>
<th>( V_{D}/R_{D} ) in A</th>
</tr>
</thead>
<tbody>
<tr>
<td>27</td>
<td>4.2</td>
<td>1.07</td>
<td>254</td>
</tr>
<tr>
<td>61</td>
<td>9.4</td>
<td>0.63</td>
<td>67</td>
</tr>
<tr>
<td>82</td>
<td>10.3</td>
<td>0.56</td>
<td>55</td>
</tr>
<tr>
<td>106</td>
<td>8.9</td>
<td>0.68</td>
<td>76</td>
</tr>
<tr>
<td>129</td>
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<td>0.59</td>
<td>59</td>
</tr>
<tr>
<td>150</td>
<td>11.5</td>
<td>0.55</td>
<td>48</td>
</tr>
<tr>
<td>174</td>
<td>11.7</td>
<td>0.55</td>
<td>48</td>
</tr>
<tr>
<td>200</td>
<td>11.8</td>
<td>0.50</td>
<td>42</td>
</tr>
<tr>
<td>250</td>
<td>12.1</td>
<td>0.48</td>
<td>40</td>
</tr>
</tbody>
</table>

Fig. 5. Also shown in Table I is the corresponding \( V_{D}/R_{D} \) ratios at different operating temperatures.

At temperatures up to and including 129°C, the \( V_{D}/R_{D} \) ratio is greater than 55.9 A, therefore \( V_{D} \) losses are higher. At the other temperatures, the ratio is between 29.3 A and 55.9 A. The traction drive will operate close to the rated operation of the induction machine; therefore, consider the comparison with 29.3 A. For all the other temperatures, the ratio is greater than 29.3 A; thus, the \( V_{D} \) losses are higher again.

As a conclusion for this case, if the doping concentration, \( N_{d} \) for the SiC diodes in this study is decreased, then \( V_{D} \) and the conduction losses decrease. The limit of this decrease is determined by the \( V_{D}/R_{D} \) ratio.

Equation (4) can be used for any sinusoidal PWM application as long as the operation current, power factor, and modulation index information is available.

2). Dc power supply

The conduction loss expression for the isolated full-bridge dc-dc converter shown in Fig. 2 is as follows:

\[
P_{\text{cond}} = d(I_{D} \cdot V_{D} + I_{D}^2 \cdot R_{D})
\]

where \( d \) is the duty ratio of the diode.

Using the same approach as in the previous subsection, the dominant losses can be found as follows:

\[
I_{D} \cdot R_{D} > ? < I_{D} \cdot V_{D} \]

\[
I_{D} > ? < \frac{V_{D}}{R_{D}}
\]

The significance of (7) can be summarized as follows:

1) If \( I_{D} > \frac{V_{D}}{R_{D}} \), then the resistive losses are higher, keep the doping and \( R_{D} \) constant because decreasing \( R_{D} \) means decreasing \( BV \), which would decrease the voltage blocking capability of the device.

2) If \( I_{D} < \frac{V_{D}}{R_{D}} \), then the \( V_{D} \) losses are higher, decrease the doping so that \( V_{D} \) will be smaller.

For different operation condition, the amount of current passing through each device and the voltage across them are calculated and the results are listed in Table II. According to Table II, \( I_{D} \) varies between 47 A and 119 A for a 5 kW dc-dc converter in the HEV simulation, then applying the above criteria,

- If \( 47A > \frac{V_{D} \cdot V_{f}}{R_{D}} \), then the first criterion applies.
- If \( 119A < \frac{V_{D} \cdot V_{f}}{R_{D}} \), then the second criterion applies.
- If \( 47A < \frac{V_{D} \cdot V_{f}}{R_{D}} \), then it depends on how close the magnitude of the current is to the minimum or maximum values for the majority of the time. For example, if the average load is varying or constant and is in a range between 3.5 and 5 kW, then the current is closer to the upper limit and the second criterion applies. If, on the other hand, the average load is in a range between 2 and 3.5 kW, then the current is closer to the lower limit and the first criterion applies.

This criteria presented here, can be applied to almost any dc-dc converter using SiC devices.

B. Switching Loss Parameters

The diode switching losses occur due to the reverse recovery of the diode, which is caused by the stored charge in the depletion region. Schottky diodes are majority carrier devices, so they do not have stored charge. However, they display a characteristic similar to reverse recovery due to the ringing of the parasitics and the internal pn junction caused by the p-rings. The p-rings are used to reduce the large reverse leakage currents.

For Schottky diodes, the switching losses can be reduced either by reducing the parasitic elements or improving the reverse recovery characteristics of the pn junction formed by the p-rings.

A diode switching loss expression has been derived in [5] using Fig. 6:

\[
P_{\text{sw}} = \frac{V_{D}}{2S} \frac{\partial I_{F}}{\partial t} + \frac{\partial S}{\partial t} \frac{V_{MOSFET}}{2S} + \frac{V_{D}}{R_{D}}
\]

where \( f_{s} \) is the switching frequency, \( V_{D} \) is the reverse blocking voltage, \( I_{F} \) is the forward diode current, \( S \) is the snappiness factor, and \( I_{rr} \) is the reverse recovery time.

<table>
<thead>
<tr>
<th>( P_{out} ) (kW)</th>
<th>( V_{DC} ) (V)</th>
<th>( I_{DIODE} ) (A)</th>
<th>( I_{MOSFET} ) (A)</th>
<th>( I_{DIODE} ) (A)</th>
<th>( I_{MOSFET} ) (A)</th>
</tr>
</thead>
<tbody>
<tr>
<td>2</td>
<td>300</td>
<td>300</td>
<td>6.67</td>
<td>84</td>
<td>47</td>
</tr>
<tr>
<td>2</td>
<td>450</td>
<td>450</td>
<td>4.44</td>
<td>84</td>
<td>47</td>
</tr>
<tr>
<td>5</td>
<td>300</td>
<td>300</td>
<td>16.67</td>
<td>84</td>
<td>119</td>
</tr>
<tr>
<td>5</td>
<td>450</td>
<td>450</td>
<td>11.11</td>
<td>84</td>
<td>119</td>
</tr>
</tbody>
</table>
In this expression all the parameters except $S$ and $t_{vr}$ are circuit dependent. These two parameters can be expressed [9] in other device parameters for a pn diode as follows,

$$S = \frac{2W_d}{b} - 1$$  \hspace{1cm} (9)$$

$$t_{vr} = \frac{2bW_d}{D_n}$$  \hspace{1cm} (10)$$

where $D_n$ is the electron diffusion constant ($D_n = \frac{kT}{q\mu_n}$), $W_d$ is the width of the drift region, and $b$ is a distance in the drift region measured from the $p^+n^-$ junction as shown in Fig. 7, $n(0)$ is the carrier density at the $p^+n^-$ junction when the diode is on, and $n^*$ is the average carrier concentration in the $n^-$ region.

Gathering the $S$ and $t_{vr}$ related terms in (8) and inserting (9) and (10), the following is obtained:

$$P_{cond, Q1} = \frac{1}{2} I^2 R_{DS, on} + \frac{1}{3\pi} M \cos \phi \frac{\delta}{\delta t}$$  \hspace{1cm} (12)$$

The only device related parameter in this expression is $R_{DS, on}$, which can be represented by other device parameters as follows

$$R_{DS, on} = R_{on, sp} = \frac{4BV^2}{\varepsilon_s \mu E_s}$$  \hspace{1cm} (13)$$

for a device with 1 cm² area where $R_{on, sp}$ is the specific on resistance of the MOSFET drift region and $\varepsilon_s$, $E_s$, and $\mu$ are material related constants.

Equation (13) is a rough estimate of a MOSFET resistance which also contains other resistive components like the channel resistance and the contact resistance. The drift resistance cannot be changed much; however, the channel and contact resistances can be lowered with more research.
B. Switching Loss Parameters

The energy loss equation of a MOSFET has been shown in [5] as follows,

\[ E_{sw} = E_{on} + E_{off} = \frac{1}{3} \mu \varepsilon \mu V \left( \frac{g_m}{w} \frac{1}{g_m V_{th}^2} \left( \frac{1}{2} K_1 (V_{th} - V_a) + 1 \right) \right) \]

where \( K_1 = \frac{g_m (V_{th} - V_a)}{J} \) and \( K_2 = \frac{g_m (V_a - V_{th})}{J} \).

\( g_m \) is the transconductance, \( J \) is the current density, \( V_{th} \) is the threshold voltage, \( V_a \) is the gate voltage, \( g_m \) is the drain voltage, \( g_m \) is the oxide capacitance, \( A_{ox} \) is the oxide dielectric constant, \( l \) is the channel length, \( \mu \) is the mobility, \( C_{ox} \) is the oxide capacitance, \( V_D \) is the drain voltage, \( t_{ox} \) is the oxide thickness, and \( A_{ox} \) is the oxide area.

If (14) is rearranged, (15) is obtained.

\[ E_{sw} = \frac{1}{3} \mu \varepsilon \mu V \left( \frac{g_m}{w} \frac{1}{g_m V_{th}^2} \left( \frac{1}{2} K_1 (V_{th} - V_a) + 1 \right) \right) \]

\[ + \frac{J}{g_m (V_{th} - V_a) - J} + \frac{J}{g_m (V_a - V_{th}) + J} \]

The most important parameter contributing to the MOSFET switching energy loss is the transconductance \( g_m \). This parameter can be represented as follows [10],

\[ g_m = \mu \frac{W}{l} C_{ox} V_D = \mu \frac{W}{l} \frac{A_{ox}}{t_{ox}} \frac{\Delta V}{\Delta V} \]

where \( \mu \) is the mobility, \( W \) is the channel width, \( l \) is the channel length, \( C_{ox} \) is the oxide capacitance, \( V_D \) is the drain voltage, \( \varepsilon_{ox} \) is the oxide dielectric constant, \( t_{ox} \) is the oxide thickness, and \( A_{ox} \) is the oxide area.

In (16), \( \mu \) and \( \varepsilon_{ox} \) are material dependent; therefore, for a specific application, four device parameters affect the transconductance, \( W \), \( A_{ox} \), \( l \), and \( t_{ox} \). The first two of these parameters are directly proportional to \( g_m \) and the others are indirectly proportional to it. From (16), the following statements can be derived:

- Decreasing \( t_{ox} \) increases \( g_m \) but \( t_{ox} \) has to be of a minimum thickness to be able to support the rated gate voltage; therefore, it cannot be changed much.
- Decreasing \( l \) increases \( g_m \), but the value of \( l \) is limited by the device processing technology.
- Increasing \( A_{ox} \) increases \( g_m \), but \( A_{ox} \) depends on the device area; it cannot be arbitrarily increases without some difficulty.
- Increasing \( W \) increases \( g_m \). To increase \( W \), the device area has to be increased proportionally.

As a summary, to decrease the MOSFET switching losses, \( g_m \) needs to be increased. Increasing the device’s area and consequently increasing \( A_{ox} \) and \( W \) seem to be the best method to do this.

V. CONCLUSIONS

In this paper, losses of the devices in a traction drive are investigated as functions of device parameters. Some modifications to device parameters are suggested to improve the losses in this drive. The next step is for device researchers to consider these suggestions and evaluate the viability of these modifications.

The interaction of device and power electronics researchers will be extremely useful in producing application specific power devices designed for optimum performance. This study is the first step to achieving this goal.

REFERENCES