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An Improved Analytical Model for Crosstalk of SiC MOSFET in a Bridge-Arm Configuration

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Abstract: SiC MOSFETs have an excellent characteristic of high switching speed, which can improve the efficiency and power density of converters significantly. However, the fast switching processes of SiC MOSFETs cause serious crosstalk problems in bridge-arm configurations, which restricts the devices' performances. This paper presents a detailed and accurate improved crosstalk analytical model, which takes into account the nonlinear capacitances, the parasitic inductances, the reverse recovery characteristics of the anti-parallel diodes, and the nonlinear voltage switching and damping oscillation process. The novelty of the proposed model lies in the fact that under the condition of comprehensively considering all these non-ideal factors of the bridge-arm, the effects of multi-parasitic elements and multi-variables coupling to the crosstalk are hierarchically divided. The parasitic elements and their correlations are described in detail and the direct and indirect variables' impacts are clearly traced. Thus, according to the different variables switching stages, the influence processes of these parasitic elements and variables can be integrated and a complete equivalent analytical model of the crosstalk process can be derived. The simulation and experiment platforms are established and a series of experimental verifications and comparisons prove that the model can replicate experimental measurements of crosstalk with good accuracy and detail.

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1. Introduction

With the advantages of low switching loss, high block-voltage, high switching speed, and high thermal conductivity, SiC MOSFETs increasingly replace traditional silicon-based MOSFETs in high frequency, high efficiency, and high power density applications [1–4]. However, many kinds of converter topologies are designed based on the bridge-arm structure, the rapid change current and voltage seriously impact the parasitic elements and coupling affect the upper and lower devices, which interfere with the driver loop states and cause wrongly triggered. That is commonly referred to as crosstalk problems [5–7]. Especially in the case of SiC MOSFETs are employed in bridge-arm with high frequency, the high current and voltage switching rates significantly increase the amplitude of crosstalk [8]. Furthermore, the gate-source voltage of SiC MOSFETs have lower turn-on threshold voltage and negative breakdown voltage compared with silicon-based MOSFETs [9], and the gate-source parasitic capacitance is relatively smaller [10], these characteristics make the driver loop of SiC MOSFETs more susceptible to the coupling effects. These two factors contradict each other that restrict the SiC MOSFETs' performances in the bridge-arm, increase the switching loss, and even may cause unintentional turn-on to affect the stability and reliability. Moreover, the crosstalk problem not only exists in common hard-switching conditions but also occurs in the soft-switching bridge-arm [11]. Thus, in order to take full advantage of SiC MOSFETs and avoid the obstacles, crosstalk is a key factor that must be evaluated.

Crosstalk is a phenomenon that involves many parasitic elements of the power loop and driver loop of the bridge-arm and is impacted by different variables during the switching processes. So far, the crosstalk phenomenon has been studied in many works. The most convenient and widely used model is to calculate the peak value of crosstalk, as described in [12–17]. It is ascribed to a function of the drain-source voltage switching rate, the input capacitance of the device, and the resistance of the driver loop, as expressed as follows:

$$V_{gs} = R_g C_{gd} \frac{dV_{ds}}{dt} \left(1 - e^{-\frac{t}{R_g(C_{gd} + C_{gs})}} \right) \quad (1)$$

Furthermore, in many practical engineering applications, the drain-source voltage switching rates are commonly considered as infinite to obtain the extremum of crosstalk [15–17]. The theoretical extremum of crosstalk can be derived from Equation (1) as $V_{gs_max} = V_{DD} C_{gd} / (C_{gd} + C_{gs})$, where V_{DD} is the drain-source voltage when the SiC MOSFET fully turn-off. In fact, these models have undergone a series of simplifications, only the impacts of driving resistances and constant capacitances of the devices are considered, while the parasitic inductances of the bridge-arm are all ignored, which leads the accuracy cannot be fully guaranteed. More detailed models take into account the influence of other parameters of the bridge-arm, the models in Ref. [18] and Ref. [19] consider the influences of parasitic inductances, which indicate that the switching current of the bridge-arm generates an induced voltage on the common source parasitic inductance, it is superimposed on the driver loop and affects the crosstalk. Literature [20] established a detailed crosstalk equivalent analytical model that includes all the parasitic inductances of the power loop and the driver loop. The time-domain response of the crosstalk can be obtained, which is summarized as a fourth-order differential equation. However, the model still relies on the given drain-source voltage switching rate as the input source, it is not been further evaluated in detail. Moreover, since the existence of the parasitic inductances in the whole bridge-arm, the reverse recovery current of the anti-parallel diode during the turn-on process also impacts the crosstalk voltage [21,22], and in Ref. [23], the effects of the reverse recovery current on crosstalk are analyzed in detail. More and more complete considerations of the parasitic elements improve the accuracy of the crosstalk analysis but greatly increases the complexity of the model establishment. In addition, the current studies mainly focus on the coupling relationship between the crosstalk and the parasitic elements, while the analysis of the variable responses as the influence sources, caused by the bridge-arm switching processes, to the crosstalk is relatively simple.

Another critical aspect is the precise description of parasitic elements and the correlative variables, it is ultimately related to the accuracy of the crosstalk response. It is well known that the interelectrode capacitances of SiC MOSFETs have nonlinear characteristics, which should be accurately described during the voltage switching processes [24]. Although the effect of nonlinear capacitances on the switching processes of SiC MOSFETs has been thoroughly studied [21,25–28], its impacts have been not enough detailed analysis for the crosstalk, which include the effects on the switching speed of the active-controlled device and the influence on the characteristics of crosstalk response circuits. The variable responses of the bridge-arm are commonly the main influence sources of the crosstalk. Elbanhawy et al. proposed a method that the voltage switching rate is regarded as a constant value [15–17], it could cause calculation errors for crosstalk from the source. In fact, the voltage switching is a nonlinear change process [20], Ref. [20] proposes a method of using experimental measurement switching rates to facilitate the calculation accuracy, but it cannot play a role in predicting the crosstalk fundamentally and explain how the nonlinear voltage switching rate influences the response. During the whole switching process of the bridge-arm, the main variables are switched by stages (voltage switching stage, current switching stage, and the reverse recovery process of the anti-parallel diode). In order to get the entire crosstalk response, the coupling effects of these variables and the coupling parasitic elements are different, which should be discussed separately and precisely [21,24]. In addition, some variable responses have overlapping situations, such

as the reverse recovery current drop process coincides with the voltage rising stage in the turn-on process [8,29], which increases the complexity of the analysis. Moreover, due to the inertial effect of the parasitic elements, there are oscillation responses of the voltage and current after the full switching process of the bridge-arm [30–33], which are also coupled to the driver loop and impact the crosstalk. To sum up, although these special characteristics and response processes greatly cumber the model design, they ultimately related to the accuracy of the crosstalk response. However, there is no complete model that can describe the crosstalk process correlates with these parasitic elements and the key variable responses of the bridge-arm particularly and comprehensively.

This paper presents a detailed and accurate analytical model of crosstalk for SiC MOSFETs applied in the bridge-arm configuration. In view of the complex coupling characteristics of multi-parasitic elements and multi-variables in the crosstalk response process, the influencing factors are summarized first. Crucially, the nonlinear characteristics of the interelectrode capacitances, the reverse recovery characteristics of the anti-parallel diodes, the parasitic inductances of the power loop, and the nonlinear voltage switching and damping oscillation process are analyzed in detail to ensure accuracy from the source. According to the switching processes of the bridge-arm, the correlation of the variables and parasitic elements with crosstalk in each stage are distinguished and the influence mechanisms behind these parameters are traced clearly. On this basis, a detailed analytical model of crosstalk is established, the source impact variables and the critical parasitic elements are all considered and combined, their direct or indirect coupling effects to the crosstalk in different switching stages are clarified. The equivalent circuits of the switching stages are established and calculation procedures are derived hierarchically. In the most critical voltage switching stages, the multitude of impact factors can be linked to the response of the drain current, it plays a key role in the impacts of crosstalk from the power loop to the driver loop, and then the crosstalk can be further evaluated with this unique direct impact source. A double pulse test platform composed of 600 V/36 A SiC MOSFETs is set up and a series of experimental results are provided to verify the correctness and accuracy of the proposed analytical model. Further, the simulation calculations and experimental results are compared with the calculation results in [17,24], and the theoretical extremums, the results show that the proposed model can describe crosstalk response in more detail and accuracy.

2. Analysis of Crosstalk Response Processes and Related Factors

A double-pulse test based on a bridge-arm configuration is a simple and effective way to analyze the crosstalk process. In this paper, a double-pulse test circuit is established to explain the crosstalk phenomenon. The equivalent circuit model is shown in Figure 1. The DC input bus voltage V_{DD} is modeled as a constant voltage source, and the output current I_{LOAD} is set as a constant current source. The upper SiC MOSFET, Q_H , is the active-controlled device, which works in the normal turn-on/off states, the lower SiC MOSFET Q_L is set at a constant turn-off state, which is used to observe the V_{gsL} crosstalk response. During the Q_H cut-off period, the load current I_{LOAD} flows through the antiparallel freewheeling diode of Q_L , D_{bL} .

Figure 1. Equivalent model of double pulse test circuit and the parasitic elements of SiC MOSFET and bridge arm.

The parasitic elements and of SiC MOSFETs devices and the bridge-arm circuits should be considered as comprehensively as possible to ensure accuracy. In practice, there are usually parasitic inductances distributed in the power loop, which are summarized as L_{LOOP} . The parasitic elements of the SiC MOSFETs need to consider the internal of the device package and the external circuit traces, as shown in Figure 1. Taking Q_L as an example, the internal parasitic elements include interelectrode capacitances C_{gsL} , C_{gdL} , C_{dsL} , where they have nonlinear characteristics, parasitic inductances of gate, drain, and source, L_{gL_in} , L_{sL_in} , and L_{dL_in} , internal gate driver resistance R_{gL_in} , and the antiparallel freewheeling diode D_{BL} , where it has reverse recovery characteristic. The external parasitic elements are the stray inductances of the electrode pins and the circuit traces connected with the gate, source, and drain, which are represented by L_{gH_ex} , L_{dH_ex} , and L_{sH_ex} , respectively. Generally, there is an external driver resistance R_{gL_ex} in the driver loop to regulate the switching speed. In order to facilitate the analysis, denoting $L_{gH} = L_{gH_ex} + L_{gH_in}$, $L_{dH} = L_{dH_ex} + L_{dH_in}$, $L_{sH} = L_{sH_ex} + L_{sH_in}$ as the total gate, drain, and source inductances of Q_L , and $R_{gL} = R_{gL_ex} + R_{gL_in}$ as the total gate resistance of Q_L . Q_H has the same structure.

At present, most discrete SiC MOSFET commercial products have two package structures, non-Kelvin package and Kelvin package. For non-Kelvin package device (such as TO-247-3), the parasitic inductance of the source should be emphasized, it is contained in the power loop and the driver loop, which makes the driver loop more vulnerable to the coupling effect of the current variation of the power loop, as is shown in Figure 2a. While the source of the Kelvin package device (such as TO-247-4) is separated for the driver loop and the power loop, which could get rid of this drawback, as shown in Figure 2b. However, the internal common source parasitic inductance cannot be completely ignored, the coupling impacts from the power loop still exist [24]. Moreover, non-Kelvin package SiC MOSFETs are widely adopted in engineering applications. Thus, the crosstalk with non-Kelvin package SiC MOSFETs is mainly discussed in this paper. The impacts of common source parasitic inductance are greater than that of the Kelvin package devices, and the analysis processes are more complicated due to the superposition influences of the power loop. This analysis method can also be simplified and transplanted to similar Kelvin packaged discrete devices and modules.

Figure 2. Structures of discrete SiC MOSFET with non-Kelvin package and Kelvin package. (a) non-Kelvin package (TO-247-3). (b) Kelvin package (TO-247-4).

In the bridge-arm configuration, the parasitic elements and their characteristics not only directly determine the response characteristics of the crosstalk voltage, but also affects the voltage and current switching process of the bridge-arm, which indirectly determine the impact sources to the crosstalk. Therefore, all these parasitic elements and their influence on the whole switching process should be integrally considered. At present, the operation characteristics and switching processes modeling of various types of power devices have been thoroughly studied in many researches. For a non-linear variation switching process of the device, the piecewise linear model is the most convenient and widely used analytical model [34–36]. Based on the piecewise linear analytical model for the switching devices, the analysis can be further related to the influence on the switching processes and crosstalk responses of bridge-arm configuration. From the perspective of crosstalk voltage, the key parameters and variables associated with crosstalk in each stage are further extracted. Thus, a thorough analysis of the influences of parasitic elements and switching variables on crosstalk and switching variables is presented. Figure 3 illustrates a typical double-pulse test turn-on and turn-off qualitative switching waveforms. Among them, Stage 2 and Stage 7 are the current switching stages of the turn-on and turn-off process, respectively. Stage 3 and Stage 6 are the voltage switching stages of the turn-on and turn-off process, respectively. The fluctuation of V_{gsL} is displayed in red in Figure 3.

Figure 3. Qualitative analytical switching sequence of turn-on and turn-off stages.

The current and voltage switching in the power loop of the bridge-arm cause two key effects that affect the crosstalk. The first one is the rapid current switching generates an induced voltage on the common source parasitic inductance, L_{sL} , which is superimposed in the driver loop. The second one is the voltage switching causes a displacement current of the gate-drain capacitance, C_{gdL} , which forms a circuit through the driver loop. Both of them are directly related factors to the responses of V_{gsL} .

More specifically, in the turn-on process of Q_H , the supply current of I_{LOAD} transfers from D_{BL} to the channel of Q_H first, which corresponds to Stage 2 in Figure 3. The freewheeling current flowing through D_{BL} decreases rapidly. Due to the reverse recovery characteristics, after the freewheeling current drops to zero, the minority carriers of D_{BL} need to be elicited first, which causes a positive reverse recovery current increase. Thereafter, D_{BL} turns to the reverse-biased state and the blocking voltage starts to build up.

During the whole process before the formation of the depletion layer of D_{BL} , there is no reverse-blocking capability and V_{dsL} does not increase. Thus, only the induced voltage of L_{sL} is coupled to the driver loop of Q_L in this stage. The equivalent circuit of this stage is shown in Figure 4a.

Figure 4. Equivalent circuit of Q_L for crosstalk analysis during turn-on process. (a) current decrease stage. (b) voltage rise stage.

The equations of the driver loop of this stage can be expressed as

$$i_{gL} = C_{gsL} \frac{dV_{gsL}}{dt} \quad (2)$$

$$R_{gL}i_{gL} + L_{gL} \frac{di_{gL}}{dt} + V_{gsL} = -L_{sL} \frac{di_{dL}}{dt} \quad (3)$$

During Stage 3, the drain-source voltage of Q_L is built up (the blocking voltage of D_{BL} increases). The gate-drain capacitance C_{gdL} is charged, the charging current flows through C_{gsL} and the driver loop of Q_L , which causes a positive crosstalk voltage of V_{gsL} . In addition, it must be noted that the power loop maintains the reverse recovery current at the end of the previous stage, which approaches its peak value and keeps the rising rate. It can be considered as the initial conditions during the increase process of V_{dsL} . With the increase of V_{dsL} , the current charging the parasitic capacitances of Q_L (C_{gdL} , C_{dsL} , and C_{gsL}), i_{dL} , decreases gradually. The equivalent circuit of this stage is shown in Figure 4b. The induced voltage on L_{sL} becomes negative, which also causes a positive impact on V_{gsL} . The circuit equations are established as

$$V_{gsL} = V_{gdL} + V_{dsL} \quad (4)$$

$$i_{dL} = -C_{gdL} \frac{dV_{gdL}}{dt} + C_{dsL} \frac{dV_{dsL}}{dt} \quad (5)$$

$$i_{gL} = C_{gdL} \frac{dV_{gdL}}{dt} + C_{gsL} \frac{dV_{gsL}}{dt} \quad (6)$$

$$V_{DSL} = L_{dL} \frac{di_{dL}}{dt} - V_{gdL} - L_{gL} \frac{di_{gL}}{dt} - R_{gL}i_{gL} \quad (7)$$

$$V_{DSL} = (L_{dL} + L_{sL}) \frac{di_{dL}}{dt} + V_{dsL} + L_{sL} \frac{di_{gL}}{dt} \quad (8)$$

where V_{DSL} is the voltage between the phase node and ground.

When V_{dsH} reaches the on-state voltage, Q_H is completely cut-off. V_{dsL} rises to V_{DD} and keeps the rising rate at t_4 , then it starts oscillating due to the parasitic inductances of the power loop and the parasitic capacitances of Q_L . The oscillations of voltage and current are coupled to the driver loop through C_{gdL} and L_{sL} , which further cause fluctuations on crosstalk, as shown waveforms after t_3 in Figure 3. The equivalent circuit is the same as Figure 4b.

In the turn-off process of Q_H , its channel resistance starts to increase first, which corresponds to Stage 6 in Figure 3. V_{dsH} starts to rise and V_{dsL} decreases, D_{BL} withstands a negative voltage and the current does not yet switch in this stage. However, the charging and discharging of the parasitic capacitances lead part of the bridge-arm current transferred to maintain the constant of the load current. The initial states of the parasitic capacitances discharge current of Q_L are zero, and the equivalent circuit is shown in Figure 5a. Same as Stage 3 in the turn-on process, the voltage and current variations all affect the crosstalk response, but the variables' directions are opposite. V_{gsL} has a negative fluctuation. When V_{dsL} drops to the forward turn-on voltage of D_{BL} , the voltage switching is completed and the current begins to divert from the channel of Q_H to D_{BL} . A negative induced voltage on L_{sL} is superimposed on the driver loop of Q_L , which causes a positive fluctuation impact on V_{gsL} . as shown the Stage 7 in Figure 3. The equivalent circuit is shown in Figure 5b. the expressions of these two stages are the same as the turn-on stages, Stage 3 and Stage 2, respectively.

Figure 5. Equivalent circuit of Q_L for crosstalk analysis during turn-off process. (a) voltage drop stage. (b) current rise stage.

Based on the analyses above, it can be seen that the response of V_{gsL} in different stages corresponds to different variables of the bridge-arm and correlate with many parasitic elements, which makes the modeling cumbersome. Especially in the voltage switching processes, the coupling of the two influence paths associates five independent state variables, the crosstalk response of V_{gsL} is simplified to a fourth-order differential equation, it is difficult to derive the time domain solutions. Moreover, compared with Equation (4) in [24], the order of V_{gsL} response is significantly increased, which is due to the influence of parasitic inductances are taken into account.

Further, there are several special parameter characteristics of SiC MOSFETs in the above-mentioned switching processes also affect the accuracy of crosstalk that must be carefully considered. SiC MOSFETs have smaller parasitic capacitances than silicon-based devices, and their values are nonlinear versus the drain-source voltage. For the SiC

MOSFETs with type CMD080120D, the nonlinear curves are shown in Figure 6. It can be found that especially the nonlinear variation of reverse capacitance C_{rss} (C_{gd}), it decreases nearly a hundredfold with the increase of V_{dsL} . This variation has an obvious influence on the characteristic of the C_{gd} displacement current loop, which directly associates with the crosstalk response according to the Equations (5)–(8). Literature [24] also clearly verified the impact of the nonlinear characteristics of C_{gdL} on the accuracy of the crosstalk response.

Figure 6. Parasitic capacitances in relationship with the drain-source voltage.

Moreover, the nonlinear variation of the parasitic capacitances of Q_H also affects the voltage switching rate. According to the literature [20,24,37], the switching rate of V_{dsH} is determined by the driving current i_{gH} and the C_{gdH} . Due to the restriction of the voltage dynamic equilibrium of the whole bridge-arm, the response of V_{dsL} presents a nonlinear variation process. As one of the input sources of the crosstalk in Stage 3 and Stage 6, the response precision of V_{dsL} must be ensured. Thus, this indirect effect of nonlinear capacitances also needs to be considered.

For the response of V_{dsL} , another critical factor is the impact of parasitic inductances. Literature [24] indicates that the switching process of V_{dsL} is nonlinear and simplified with a piecewise linear method, but it did not take into account the influence of parasitic inductances of the power loop. It considers that the response of V_{dsL} is the reversed response of V_{dsH} , and the crosstalk amplitude of V_{gsL} reaches the maximum when V_{dsL} reaches the bus voltage V_{DD} in the turn-on process. These settings will bring errors from the source. In practice, due to the parasitic inductances of the power loop, on the one hand, the parasitic inductances participate in the voltage switching process, which shared parts of voltage with the transient variation of V_{dsL} , the sum of V_{dsH} , V_{dsL} and the induced voltages of parasitic inductances is V_{DD} . On the other hand, V_{dsL} will continue to oscillate after V_{dsL} reaches V_{DD} in the turn-on process, which will impact the driver loop and cause fluctuation of V_{gsL} . The time of V_{gsL} to reach the maximum value, T_m , will also be delayed, as shown in Figure 7.

Figure 7. The formation of crosstalk in voltage switching stage of turn-on process.

To sum up, the nonlinear characteristics of parasitic capacitances, the nonlinear variation of the drain-source voltage of Q_H , the parasitic inductance of the power loop, and the voltage oscillations are all correlated to the crosstalk response, and they are coupled with each other multiply that increases the obstacles to the modeling. Therefore, a detailed and accurate model is necessary for the design guide of bridge-arm configuration to help evaluate crosstalk, as well as it could explain the influence mechanism of relevant parameters.

3. The Analytical Model for Crosstalk

According to the analyses in Section 2, the crosstalk phenomenon lasts almost throughout the whole switching process of the bridge-arm and different parasitic elements are coupled in different switching stages. It is necessary to consider the parasitic elements specifically and their correlative effects on the related variables in each stage should be studied deeply. Establishing accurate models of them is an important step for the crosstalk analysis. So, the modeling of special parameters and main variables are provided first, based on them, a crosstalk model can be derived in detail.

3.1. Model Establishments of Special Parameters and Main Variables

3.1.1. Nonlinear Capacitances

The device manufacturer provides the curves of the input capacitance C_{iss} , the output capacitance C_{oss} , and the reverse capacitance C_{rss} in the datasheet in general. According to the method provided in Refs. [26,27], the values of them can be piecewise fitted with high precision, which can be described by the equation

$$C(V_{ds}) = C_{j0} \left(1 + \frac{V_{ds}}{V_j} \right)^{-m} \quad (9)$$

where C_{j0} is the capacitance value when $V_{ds} = 0$, V_j is the built-in voltage and can be treated as an undetermined coefficient, m is the capacitance gradient factor. They can be adjusted according to the range of V_{ds} . Figure 8 shows the fitted capacitances simulation curves compared with the datasheet curves.

Figure 8. Comparison between the data from datasheet and the simulation.

Interelectrode capacitances are usually used in actual model establishment and calculations, which can be extracted according to Equation (10). Their comparison with the datasheet curves is shown in Figure 9. Since C_{gs} and C_{ds} are much larger than C_{gd} , the interelectrode capacitances curves are almost the same with the parasitic capacitance curves from the datasheet.

$$\begin{cases} C_{iss} = C_{gs} + C_{gd} \\ C_{oss} = C_{ds} + C_{gd} \\ C_{rss} = C_{gd} \end{cases} \quad (10)$$

Figure 9. Comparison between the capacitances from datasheet and the interelectrode capacitances.

3.1.2. Current Switching Rate and Reverse Recovery Current of D_{BL}

Essentially, the current switching rate is determined by the active-controlled device. When Q_H works in the saturation region, Q_H is modeled as a dependent current source controlled by V_{gsH} . The expression of the channel current is given by

$$i_{ch} = g_f(V_{gsH} - V_{th}) \quad (11)$$

where i_{ch} is the channel current of Q_H , g_f is the transconductance coefficient and it remains at the maximum. The rising rate of V_{gsH} is entirely dependent on the driver loop characteristics of Q_H . It worth noting that the induced voltage on L_{sH} is superimposed in the driver loop, which provides negative feedback to the driver loop [21,37]. This factor restricts the rising rate of V_{gsH} and makes the current switching rate reach a maximum. The circuit equations of the Q_H driver loop can be expressed as

$$i_{gH} = C_{gsH} \frac{dV_{gsH}}{dt} \quad (12)$$

$$R_{gH}i_{gH} + (L_{gH} + L_{sH}) \frac{di_{gH}}{dt} + V_{idsH} + V_{gsH} = V_{GATE} \quad (13)$$

where V_{idsH} is induced voltage on L_{sH} caused by the changing of i_{dH}

$$V_{idsH} = L_{sH} \frac{di_{dH}}{dt} \quad (14)$$

When the current switching rate reaches the maximum, the induced voltage of the parasitic inductances of the power loop remains constant, and the interelectrode capacitances do not have charge current, thus, the current of the power loop can be considered as $i_{dH} = i_{ch}$.

In Stage 2, when the channel current rises to I_{LOAD} , the freewheeling current of D_{BL} drops to zero and starts the reverse recovery process, i_{dL} increases in a positive direction and its rate remains the same as the drop rate before, it is because the equilibrium state remains unchanged (the characteristics of the Q_H driver loop and the transconductance). In order to obtain the rising duration of the reverse recovery current, the diode model in reference [38] is employed, which separates the reverse recovery process and charging process of junction capacitance into two stages. It is more suitable to describe the processes of current rising and voltage build-up respectively, as is shown in Figure 10. In general, the reverse recovery charge given in the datasheet is the charge for the entire reverse recovery process, Q_{rr} ; thus, the charge of the reverse recovery process, Q_{rs} , needs to derive according to the provided test conditions. The equations are established as follows:

$$t_{rr} = t_{rs} + t_{rf} \quad (15)$$

$$Q_{rr} = Q_{rs} + Q_{rf} \quad (16)$$

$$i_{rr_peak} = t_{rs} \left. \frac{di_{dL}}{dt} \right|_{i_{dL}=I_{LOAD}} \quad (17)$$

$$Q_{rr} = \frac{1}{2} i_{rr_peak} t_{rr} \quad (18)$$

$$Q_{rs} = \frac{1}{2} i_{rr_peak} t_{rs} \quad (19)$$

Figure 10. Modeling of the reverse recovery process of the antiparallel freewheeling diode.

The test conditions are listed in Table 1, the relationship between Q_{rs} and Q_{rr} can be derived as Equation (20).

$$Q_{rs} = 2 \left(\frac{Q_{rr}}{t_{rr}} \right)^2 / \frac{di_{dL}}{dt} \Big|_{i_{dL}=I_{LOAD}} \quad (20)$$

Table 1. Antiparallel Freewheeling Diode Characteristics and Reverse Recovery Test Parameters.

Parameter	Value	Test Condition
V_{DD} (V)	3.3	$V_{gs} = -5$ V, $I_{LOAD} = 20$ A, $V_{DD} = 800$ V $di_d/dt = 2400$ A/us
t_{rr} (ns)	32	
Q_{rr} (nC)	192	

Therefore, in the analysis of the actual current switching of the bridge-arm, according to Q_{rs} and the rising rate of i_{dL} , the reverse recovery current value at the time of D_{BL} turning to the reverse-biased can be obtained, which is the initial condition of the voltage switching response in the next stage.

3.1.3. Voltage Switching Rate

The impact of the voltage switching is the most obvious source factor on the crosstalk. In the bridge-arm configuration, the voltage switching is controlled by the active-controlled device. When the drain-source voltage of the active-controlled device changes, the drain-source voltage of the passive device changes accordingly to keep the bus supply voltage constant.

For the active-controlled SiC MOSFET, Q_H , the current discharges C_{gd} in this stage, meanwhile, V_{gs} is clamped to the V_{miller} and keeps constant, which has been well known as miller platform voltage. the gate driver works like a current source. There is not a significant charging current through C_{gsH} . i_g can be express as

$$i_{gH} = C_{gdH} \frac{dV_{gdH}}{dt} = C_{gdH} \frac{d(V_{miller} - V_{dsH})}{dt} = -C_{gdH} \frac{dV_{dsH}}{dt} \quad (21)$$

V_{gsH} keeps at V_{miller} , which has a clamping effect on the driver loop, the maximum current that the driver loop can reach is limited. Therefore, i_{gH} remains constant, and the

induced voltage on the parasitic inductances of the driver loop can be ignored. The driver loop equation of Q_H can be expressed as

$$V_{GATE} = R_{gH}i_{gH} + V_{miller} + V_{idsH} \quad (22)$$

where V_{idsH} is induced voltage on L_{sH} caused by i_{dH} . Combining (14), (21), and (22), the switching rate of V_{dsH} can be obtained as

$$\frac{dV_{dsH}}{dt} = -\frac{V_{GATE} - V_{miller}}{R_{gH}C_{gdH}} + \frac{L_{sH}}{R_{gH}C_{gdH}} \frac{di_{dH}}{dt} \quad (23)$$

It can be seen from (23) that the nonlinear variation of V_{dsH} is affected by the nonlinear characteristics of C_{gdH} and the rate of i_{dH} . In addition, the common source parasitic inductance L_{sH} provides negative feedback to the switching rate of V_{dsH} . In the turn-on process, since the reverse recovery current decrease, a negative induced voltage is generated on L_{sH} , which will increase the drop rate of V_{dsH} . Conversely, it suppresses the rising rate of V_{dsH} in the turn-off process.

Due to the parasitic inductances exist in the power loop, the response of V_{dsL} cannot be directly described as a reversed response of V_{dsH} . In this paper, V_{dsL} is equivalent to a response of the whole parasitic capacitances of Q_L charged by the applied voltage $V_{DD} - V_{dsH}$, where all the parasitic inductances of the power loop and the driver loop branch are taken into consideration. The equivalent circuits are shown in Figure 11. In this way, an accurate nonlinear input is given from the source and the influences of parasitic elements are thoroughly considered. These settings can ensure the accuracy of the response of V_{dsL} .

Figure 11. Equivalent circuits for V_{dsL} analysis circuit.

Furthermore, for the turn-on voltage switching stage (Stage 3), the rising of V_{dsL} is accompanied by the reverse recovery process of D_{BL} (the blocking voltage building up), there are the initial value and rate of the drain current, $i_{dL}|_{t=t_2}$ and $di_{dL}/dt|_{t=t_2}$, at the

end time of Stage 2, which can be obtained from Section 2. For the whole structs of Q_L , the reverse recovery model of the independently packaged freewheeling diode applied in Refs. [21,38] is inadequate to describe the drain-source voltage building process. It is because the freewheeling diode is inversely paralleled with the SiC MOSFET; hence, all the parasitic capacitances need to be considered. The independent barrier capacitance charge model will bring about dynamic accumulation errors for the response of V_{dsL} . To sum up, combining with the bridge-arm structure, all the factors of the reverse recovery current and the parasitic elements should be considered as a whole in this stage.

The capacitances, C_{gsL} , C_{gdL} , and C_{dsL} are in delta connection and the driver loop is paralleled with C_{gsL} and L_{sL} . The delta connection of capacitances can be transformed into the star connection, C_{gL} , C_{dL} , and C_{sL} [39], as shown in Figure 12a. The expressions of capacitances, C_{gL} , C_{dL} , and C_{sL} are given as follows

$$\begin{cases} C_{gL} = C_{gsL} + C_{gdL} + C_{gsL}C_{gdL}/C_{dsL} \\ C_{sL} = C_{gsL} + C_{dsL} + C_{gsL}C_{dsL}/C_{gdL} \\ C_{dL} = C_{gdL} + C_{dsL} + C_{gdL}C_{dsL}/C_{gsL} \end{cases} \quad (24)$$

Figure 12. Simplified equivalent circuit for bridge-arm in Stage 3. (a) Star connection equivalent circuit; (b) Second-order equivalent circuit.

Considered from the perspective of the power loop, the paralleled driver loop and the C_{sL} - L_{sL} branch can be further regarded as the equivalent impedance to evaluate its impact. The impedance of the driver loop can be expressed as

$$X_{gL} = R_{gL} + j\omega_{n_P}L_{gL} + \frac{1}{j\omega_{n_P}C_{gL}} \quad (25)$$

The impedance of the C_{sL} - L_{sL} branch is expressed as

$$X_{sL} = j\omega_{n_P}L_{sL} + \frac{1}{j\omega_{n_P}C_{sL}} \quad (26)$$

where ω_{n_P} is the resonance frequency of the power loop and it can be approximated by

$$\omega_{n_P} \approx \frac{1}{\sqrt{(L_{dH} + L_{sH} + L_{dL} + L_{sL})C_{eqL}}} \quad (27)$$

$$C_{eqL} = \frac{C_{dL} C_{sL}}{C_{dL} + C_{sL}} \quad (28)$$

The total equivalent impedance of the lower SiC MOSFET the power loop is

$$X_L = \frac{X_{gL} \cdot X_{sL}}{X_{gL} + X_{sL}} \quad (29)$$

The equivalent damping resistance is determined by the ratio of current flowing through R_{gL} and the total current of the power loop, which can be expressed by Equation (30).

$$\begin{aligned} R_{gL_eq} &= R_{gL} \left(\frac{|X_L|}{|X_{gL}|} \right)^2 \\ &= R_{gL} \frac{\left(\omega_{n_P} L_{sL} - \frac{1}{\omega_{n_P} C_{sL}} \right)^2}{R_{gL}^2 + \left(\omega_{n_P} L_{gL} - \frac{1}{\omega_{n_P} C_{gL}} + \omega_{n_P} L_{sL} - \frac{1}{\omega_{n_P} C_{sL}} \right)^2} \end{aligned} \quad (30)$$

Thus, the power loop is equivalent to a second-order circuit, as shown in Figure 12b. The total equivalent resistance can be express by

$$R_{eqL} = R_{gL_eq} + R_{dsH_on} + R_{LOOP} \quad (31)$$

The power loop equations can be modeled as

$$V_{dsL} + R_{eqL} i_{dL} + (L_{dH} + L_{sH} + L_{dL} + L_{LOOP}) \frac{di_{dL}}{dt} = V_{DD} - V_{dsH} \quad (32)$$

$$i_{dL} = C_{eqL} \frac{dV_{dsL}}{dt} \quad (33)$$

It is worth noting that the nonlinear changes of C_g , C_d , and C_s also need to consider. They can be derived according to the curves of interelectrode capacitances, which is shown in Figure 13.

Figure 13. Comparison between the equivalent capacitances and the interelectrode capacitances.

When V_{dsH} drops to the on-state voltage V_{dsH_on} (Stage 4), although Q_H enters the ohmic region, V_{dsL} of Q_L will continue to oscillate after it rises to V_{DD} due to the RLC characteristic of the power loop. The ideal voltage division given in [24] cannot be used

here, since it does not take into account the power loop oscillation. In Stage 4, the supply voltage is $V_{DD} - V_{dsH_on}$ and the equivalent circuit is shown in Figure 14. The structure of the equivalent circuit is the same as in the previous stage and the expressions of C_{eqL} and R_{eqL} can also be described as (28) and (30). The power loop equation is transformed from (32) to (34).

$$V_{dsL} + R_{eq}i_{dL} + (L_{dH} + L_{sH} + L_{dL} + L_{LOOP})\frac{di_{dL}}{dt} = V_{DD} - V_{dsH_on} \quad (34)$$

Figure 14. Simplified equivalent circuit for bridge-arm in Stage 4. (a) Star connection equivalent circuit; (b) Second-order equivalent circuit.

In the turn-off process, the circuit topologies transformations for the analysis of the voltage and current switching stages are the same as those in the turn-on stages. The voltage and current switching rates are still controlled by Q_H . The only difference is that the initial value and rate of the drain current are zero at the beginning of Stage 6.

3.2. Analytical Model of Crosstalk

During the current switching stages (Stage 2 and Stage 7), V_{dsL} does not change, only the induced voltage on the common source parasitic inductance, L_{sL} , affects the driver loop, combining the equations of the driver loop (2), (3) with the equations of the channel currents (11)–(14), the response of V_{gsL} is correlated with V_{gsH} ultimately:

$$L_{gL}C_{gsL}\frac{d^2V_{gsL}}{dt^2} + R_{gL}C_{gsL}\frac{dV_{gsL}}{dt} + V_{gsL} = -L_{sL}g_f\frac{dV_{gsH}}{dt} \quad (35)$$

$$(L_{gH} + L_{sH})C_{gsH}\frac{d^2V_{gsH}}{dt^2} + (R_{gH}C_{gsH} + L_{sH}g_f)\frac{dV_{gsH}}{dt} + V_{gsH} = V_{GATE} \quad (36)$$

During the voltage switching stages (Stage 3 and Stage 6), it can be generalized that the initial state of the drain current, the nonlinear switching rate of the drain-source voltage, the parasitic inductances of the power loop and common source, and the nonlinear characteristics of the interelectrode capacitances affect the response of V_{dsL} . Through the analysis in section A. All these factors can be further directly or indirectly linked with the response of i_{dL} . i_{dL} becomes a key vinculum to reflect the impacts of the voltage switching. The correlations are summarized in Figure 15.

Figure 15. Relationships of i_{dL} with the parameters of the bridge-arm.

Combining (31)–(33) with (23), i_{dL} can be expressed as

$$C_{eqL}(L_{dH} + L_{sH} + L_{dL} + L_{LOOP}) \frac{d^2 i_{dL}}{dt^2} + C_{eqL} \left(R_{eqL} + \frac{L_{sH}}{R_{gH} C_{gdH}} \right) \frac{di_{dL}}{dt} + i_{dL} = \frac{(V_{GATE} - V_{miller})}{R_{gH} C_{gdH}} C_{eqL} \quad (37)$$

After V_{dsL} rises to V_{DD} in the turn-on process, i_{dL} is expressed as

$$C_{eqL}(L_{dH} + L_{sH} + L_{dL} + L_{LOOP}) \frac{d^2 i_{dL}}{dt^2} + C_{eqL} R_{eqL} \frac{di_{dL}}{dt} + i_{dL} = 0 \quad (38)$$

Next, i_{dL} is used as an input source to analyze its influence on the crosstalk of the driver loop. The interelectrode capacitances can be converted into two branches with the delta-star transformation, which helps simplify the analysis. The current flowing through C_{dL} is i_{dL} . From the perspective of the driver loop, the crosstalk response is converted to the voltage difference between C_{sL} and C_{gL} on the two branches. The equivalent circuit is shown in Figure 16 and the equations can be established as

$$V_{gsL} = V_{gL} + V_{sL} \quad (39)$$

$$i_{dL} = i_{sL} - i_{gL} \quad (40)$$

$$i_{gL} = C_{gL} \frac{dV_{gL}}{dt} \quad (41)$$

$$i_{sL} = C_{sL} \frac{dV_{sL}}{dt} \quad (42)$$

$$-V_{gL} - R_{gL} i_{gL} - L_{gL} \frac{di_{gL}}{dt} = V_{sL} + L_{sL} \frac{di_{sL}}{dt} \quad (43)$$

Figure 16. Equivalent circuit for the driver loop of Q_L .

Therefore, the relationship of i_{gL} and i_{dL} can be obtained as follow

$$(L_{gL} + L_{sL})C_{eqL_d}\frac{d^2i_{gL}}{dt^2} + R_{gL}C_{eqL_d}\frac{di_{gL}}{dt} + i_{gL} = -L_{sL}C_{eqL_d}\frac{d^2i_{dL}}{dt^2} + \frac{C_{eqL_d}}{C_{sL}}i_{dL} \quad (44)$$

V_{gsL} can be expressed by

$$V_{gsL} = -R_{gL}i_{gL} - (L_{gL} + L_{sL})\frac{di_{gL}}{dt} - L_{sL}\frac{di_{dL}}{dt} \quad (45)$$

where $C_{eqL_d} = C_{gL}C_{sL}/(C_{gL} + C_{sL})$. Finally, the response of V_{gsL} can be deduced by correlating the differential Equations (37), (44), and (45).

So far, the crosstalk responses of the turn-on and turn-off processes are fully expressed in the time domain. The Equations (35), (44), and (45) can be discretized and brought into MATLAB, where the directly correlated variable responses are expressed by Equations (36)–(38). Based on the software, the whole simulation waveforms of V_{gsL} can be presented minutely and accurately. The detailed simulation and experimental verification are described in the next section.

4. Simulation and Experimental Verification

To verify the analysis and modeling method in the previous section, a typical double pulse test circuit platform is built, Figure 17 shows the photograph of the test platform. Two 1200 V/36 A SiC MOSFETs C2M0080120D, from Wolfspeed corporation, are connected in series as the upper and lower switching devices of the bridge-arm, a 12 μ H ferrite-core inductance is connected in parallel with the lower SiC MOSFET as the load inductance. The test circuit is supported by 600 V DC supply voltage, multiple groups of 10 μ F capacitances are connected in parallel with the DC input terminal to ensure the voltage stability. The initial parameters of the circuit model are listed in Table 2. Some fixed parameter values were obtained through the datasheet, including Q_{rr} , g_f , R_{g_in} , and R_{ds_on} . The values of parasitic inductances were implemented by an ANSYS Q3D Extractor finite-element analysis (FEA) simulation, which includes internal pins of the devices and the printed

circuit board (PCB) layout [40]. The external pins of the devices and the connecting leads of the power loop were measured by an Agilent 4395A impedance analyzer.

Figure 17. Photograph of the double pulse test platform.

Table 2. Fixed Parameters Setup.

Component	Parameter	Value	Parameter	Value
Power circuit setting	V_{DD}	600 V	R_{LOOP}	230 m Ω
	I_{DD}	20 A	L_{LOOP}	32 nH
Fixed parameters of upper and lower SiC MOSFET (C2M0080120D)	V_{th}	3.3 V	Q_{rr}	192 nC
	V_{GATE}	+20 V / −5 V	V_{GATE}	−5 V
	g_f	8.1 S	V_{SD}	3.3 V
	R_{dsH_on}	40 m Ω	R_{dsH_on}	40 m Ω
	R_{gH_ex}	10 Ω	R_{gL_ex}	10 Ω
	R_{gH_in}	4.6 Ω	R_{gL_in}	4.6 Ω
	L_{gH_in}	1.1 nH	L_{gL_in}	1.7 nH
	L_{gH_ex}	3.0 nH	L_{gL_ex}	2.8 nH
	L_{sH_in}	1.3 nH	L_{sL_in}	1.3 nH
	L_{sH_ex}	3.7 nH	L_{sL_ex}	3.4 nH
	L_{dH_in}	1.2 nH	L_{dL_in}	1.1 nH
	L_{dH_ex}	3.3 nH	L_{dL_ex}	3.6 nH

The circuit setting and the parameter values in the simulation calculations are based on this real circuit experiment platform within a reasonable range. The nonlinear capacitances are piecewise fitted according to the datasheet, the piecewise parameters are listed in Table 3. This piecewise fitting model can be invoked as independent computing units based on the values of V_{dsH} and V_{dsL} to obtain the precise value in every discrete computing period.

Table 3. Parameters of Capacitance Piecewise Fitting Model.

Capacitance	Parameter		V_{ds} Range (V)				
	C_{j0} (pF)	Fitting Parameter	0–9	9–20	20–35	35–500	500–1200
C_{iss}	1700	V_j m	20 0.8		1×10^{-4} 0.028		1×10^{-6} 0.0125
C_{oss}	1650	V_j m		5 0.9		0.3 0.39	1×10^{-7} 0.13
C_{rss}	400	V_j m	350 45	300 43	0.4 0.73	5×10^{-4} 0.29	3×10^{-4} 0.29

In order to distinguish the response stages and accurately divide the related factors of crosstalk in different stages, the terminal value of the current response in Stage 2 and Stage 7 and the voltage response in Stage 3 and Stage 6 need to be identified as the switching judgment conditions. They are listed in Table 4 and the equations also need to be added to the simulation calculation program.

Table 4. Switching Time and Calculations of Judgement Conditions.

Switching Time	$t1$	$t2$	$t3$	$t6$	$t7$	$t8$
Judgement condition	$V_{gsH} > V_{th}$	$i_{dL} > I_{rr_peak}$	$V_{dsH} < V_{th}$	$V_{gsH} < V_{miller}$	$V_{dsL} < 0$	$i_{dL} < -i_{LOAD}$
Calculate equations	(11), (12), and (13) $V_{idsH} = 0$	(10), (11), (12), and (13)	(22) integrated computation from $t2$	(11), (12), and (13) $V_{idsH} = 0$	(22) integrated computation from $t6$	(10), (11), (12), and (13)
Initial states	–	$V_{gsH} \mid_{t=t1}$ $dV_{gsH}/dt \mid_{t=t1}$	V_{DD} $i_{dL} \mid_{t=t2}$ $d_{idL}/dt \mid_{t=t2}$	–	–	$i_{dL} \mid_{t=t7}$ $d_{idL}/dt \mid_{t=t7}$

According to the piecewise conditions, the crosstalk response can be calculated stage-by-stage with iterative method, by discretizing the equations, it is convenient to obtain the numerical solution of the variable responses. The simulated waveforms of crosstalk, when $V_{DD} = 600$ V and $I_{LOAD} = 20$ A, are shown in Figure 18. In addition, the calculation results according to the model presented in the literature [17,24], and the theoretical extremum of crosstalk [15] are also placed on the same coordinates to horizontal compare the differences.

It can be seen from Figure 18, during the current switching stages in the turn-on process, there is a drop on V_{gsL} , which reflects the impacts of the common source parasitic inductance with the rapid current changes. However, the calculations in Refs. [17,24] do not consider the influence of the parasitic inductances, so that there is a significant difference in the negative fluctuations of the crosstalk. There are trends in our data to suggest that the positive peak value of crosstalk is higher compared with the calculation in Ref. [24]. V_{gsL} will continue to rise and start oscillating after the voltage switching process, which embodies the coupling effect of the oscillation caused by the parasitic inductance and parasitic capacitance of the power loop on the crosstalk. In the turn-off process, the negative peak value of the crosstalk proposed in this paper is higher than the calculated value in Ref. [24], which is also due to the influence of parasitic inductance. More specifically, during the voltage switching stages (Stage 6), the parasitic capacitances of Q_L are discharged with the decrease of V_{dsL} , the displacement current of C_{gdL} induces a negative fluctuation in V_{gsL} , but the discharge current for C_{dsL} starts to increase from zero, which causes a negative induced voltage on L_{sL} , thus, it will suppress the negative fluctuation of V_{gsL} . Further, the simulation results of the crosstalk peak values given in this paper and Ref. [24] are both significantly larger than the calculation results with constant capacitances, these all

indicate that the nonlinear parasitic capacitances have obvious influences on the crosstalk evaluation.

(a)

(b)

Figure 18. Crosstalk waveforms provided by the proposed analytical model, the analytical model in [17] the analytical model in [24], and the extremum calculated by theoretical calculation. (a) turn-on process. (b) turn-off process.

The experimental platform is employed to verify the correctness of the proposed model. In the experimental test, since there is no any crosstalk suppression method, to prevent spurious trigger, the negative driver voltage level of Q_L is set at -5 V. Adjusting the negative voltage bias does not have a significant effect on the responses of crosstalk voltage [24]. The double pulse trigger signal position is adjusted to set the load current at 20 A. The first turn-off and the second turn-on waveforms are extracted in this paper due to the drain current value almost maintain the same at the switching time, as is shown in

Figure 19. Figure 20 compares the experiment results and the simulation results of crosstalk waveforms in the turn-on and turn-off process, the experimental data are uniformly added 5 V to facilitate the comparisons. The red lines represent the experiment waveforms and the blue lines represent the simulation waveforms.

Figure 19. Double plus test waveforms.

Figure 20. Comparison of simulation and experimental waveforms of crosstalk in (a) turn-on process, (b) turn-off process.

It is apparent that the proposed analytical model is more detailed and matches better with the experimental results (in terms of the voltage drop in Stage 2, the oscillations after the voltage switching, and the peak values response) than models in Refs. [17,24]. There

are deviations between the simulation and actual results of the oscillation frequencies, it is because the measurement deviation of parasitic inductances and equivalent resistance of the circuit, and the model of the nonlinear capacitances is difficult to maintain a high consistency with the data in the datasheet. These deviations are inevitable but the variation trends can still be correctly inferred.

According to the previous analysis, the variables that directly impact the crosstalk are the current and voltage switching rates and they are closely related to the driver capability of the active-controlled device. In the simulations and experiments, adjusting the driver resistant R_{gH} to change the switching rate of the bridge-arm, this method can adjust both voltage and current switching rates according to (11)–(14) and (23). The simulations and experiments waveforms of crosstalk are shown in Figure 21, and the trend comparisons of the relative crosstalk peak values are provided in Figure 22.

Figure 21. Comparison of simulation and experimental waveforms of crosstalk with different R_{gH} . (a,b): simulation. (c,d): experiment. Left: turn-on. Right: turn-off.

Figure 22. Comparison of simulation and experimental variation trend of relative crosstalk peak value with different R_{gH} . (a) turn-on. (b) turn-off.

It can be seen from Figures 21 and 22 that the experimental waveforms and the variation trends of crosstalk are in good agreement with the simulation results, which verifies the correctness and the accuracy of the proposed model with the situations where the switching rates are dynamically adjusted. However, the deviations of the peak values between the simulation and experiment results also need to be noticed. Especially in the turn-on process, the measured amplitudes of V_{gsL} are smaller than the calculated values. One of the main reasons is that the voltage measurement probes can only be connected between the external terminal nodes of the gate and source, thus the internal voltages of

parasitic elements are included in the measurement values. During the turn-on process, the voltages on the internal gate resistance, the gate parasitic inductances, and the internal source parasitic inductances are all opposite to the gate-source voltage of C_{gsL} , which leads to the measured peak values of V_{gsL} are smaller than the calculated values. In addition, the reverse recovery current decreases rapidly from the beginning of the voltage switching stage, which produces a relatively larger negative induced voltage on the internal common source parasitic inductance. These two factors cause more obvious deviations of the crosstalk voltage in the turn-on process.

These measurement deviations are inevitable but the variation trend can still be correctly inferred, and importantly, the directly calculated analytical values of internal V_{gsL} are essentially needed to evaluate the real crosstalk voltage response rather than the voltage between the external nodes. The measurement values can be considered as indirect reflections of the crosstalk. The precise values and errors of experimental measurements and simulation calculations for the relative amplitudes of the crosstalk are listed in Table 5.

Table 5. Comparison of Experiments and Simulations with Variable R_{gH} .

	R_{gH} (Ω)	10 Ω	14 Ω	18 Ω	22 Ω	26 Ω
Turn-on	Relative V_{gsL} with Simulation (V)	4.282	3.838	3.356	2.898	2.481
	Relative V_{gsL} with Experiment (V)	4.05	3.58	3.10	2.59	2.12
	Difference compared with experiment	5.73%	7.21%	8.26%	11.89%	17.03%
Turn-off	Relative V_{gsL} with Simulation (V)	−2.742	−2.616	−2.443	−2.159	−1.881
	Relative V_{gsL} with Experiment (V)	−2.59	−2.52	−2.38	−2.20	−2.02
	Difference compared with experiment	5.87%	3.81%	2.65%	1.86%	6.88%

According to (32) and the judgment conditions of t_3 and t_6 , the switching rate and duration of the voltage also changed with the different input voltage levels, which indirectly affects the crosstalk responses in the voltage switching process, especially the peak value. To verify the model in a further step, different DC voltages are applied to the double pulse test platform, while other test conditions are the same as the initial setting, the load current is kept at 20 A by adjusting the trigger position of the driver signal. V_{DD} is increased from 200 V to 600 V with 100 V step. The simulation and experimental crosstalk waveforms with different V_{DD} are shown in Figure 23. The precise values and errors of experimental measurements and simulation calculations for the relative amplitudes of the crosstalk are listed in Table 6.

Figure 23. Cont.

Figure 23. Comparison of simulation and experimental waveforms of crosstalk with different V_{DD} . (a,b): simulation. (c,d): experiment. Left: turn-on. Right: turn-off.

Table 6. Comparison of Experiments and Simulations with Variable V_{DD} .

V_{DD} (V)		200	300	400	500	600
Turn-on	Relative V_{gsL} with Simulation (V)	1.860	2.656	3.326	3.870	4.282
	Relative V_{gsL} with Experiment (V)	1.81	2.58	3.22	3.72	4.05
	Difference compared with experiment	2.76%	1.46%	3.29%	4.03%	5.73%
Turn-off	Relative V_{gsL} with Simulation (V)	−0.911	−1.554	−2.166	−2.510	−2.742
	Relative V_{gsL} with Experiment (V)	−1.00	−1.61	−2.11	−2.42	−2.59
	Difference compared with experiment	8.90%	3.48%	2.65%	3.72%	5.87%

As shown in Figure 23, the experimental waveforms of crosstalk agree well with the simulation results with the different supply voltage. Figure 24 shows the comparisons of the relative crosstalk peak values extracted from the experiments with the simulation results and the calculated values according to the algorithm of [24]. It can be seen that the crosstalk peaks simulated by the proposed model are more accurate than the crosstalk peaks values calculated based on the method in [24]. This can be attributed to the errors caused by the parasitic inductances, especially in the turn-off process, the induced voltage on L_{sL} and the impedance of the driver loop have opposite effects on the fluctuation of V_{gsL} , which causes greater deviations.

Figure 24. Comparison of simulation and experimental variation trend of relative crosstalk peak value with different V_{DD} . (a) turn-on. (b) turn-off.

5. Conclusions

In this paper, a detailed and accurate analytical model of crosstalk for SiC MOSFETs bridge-leg configuration is proposed. The source impact variables and the critical parasitic elements to the crosstalk are comprehensively considered. The critical factors of SiC MOSFETs and bridge-arm are taken into account, including the nonlinear characteristics of the interelectrode capacitances, the reverse recovery characteristics of the anti-parallel diodes, the parasitic inductances of the power loop, and the nonlinear voltage switching and damping oscillation process. Their direct or indirect coupling effects to the crosstalk in different switching stages are analyzed. Combining with these factors, the proposed model covers the whole crosstalk response of the bridge-arm switching processes and

the equivalent analysis circuits of each switching stage are established. The correctness and the accuracy of the proposed model are validated by simulation and experiment. Compared with the current crosstalk model, the proposed model can present more details and accurately of crosstalk prediction and is closer to the practical engineering application. The proposed model can be employed to evaluate the crosstalk responses to ensure the stability and reliability of bridge-arms, and can also be used in the design and development of power converters to reduce loss and play full operation performances of SiC MOSFETs. In future work, the influence trends of the parasitic elements and variables on the crosstalk response can be detailed analyzed based on the model. It should be noticed that the temperature characteristics of the model are not concerned in this paper, this may cause deviations between the modeling values and the actual values of parasitic parameters, and in practical projects, the temperature of the switching devices of power converters varies with the operating time. Thus, the improved analytical model for crosstalk with considering temperature characteristics will be studied in future work to improve the accuracy.

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