Differential Mode Conducted EMI Prediction in Three Phase SiC Inverters

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Abstract. With the application of Silicon carbide (SiC) power MOSFETs in three phase inverters, higher level electromagnetic interference (EMI) produced. Thus, it is more urgent to predict EMI level in the design phase. This paper describes a frequency-domain approach to predict differential mode(DM) EMI of three phase SiC inverters. Both noise path and DM noise source are modeled. The calculated result is compared with experimental result and they agree well with each other. It is indicated that the proposed method with accurate noise and path modeling is an effective method for DM prediction of the system.

1. Introduction

Recently, rapid development of silicon carbide (SiC) based power devices has attracted a lot of attentions. For power converters using SiC components, the switching frequency becomes higher, the switching times become shorter, and the size of converters becomes smaller. However, electromagnetic interference (EMI) problems become more serious and need to be addressed. In this paper, the analyzed system is a 2-level 3-phase inverter system as shown in figure 1, which is widely used in motor drive systems.
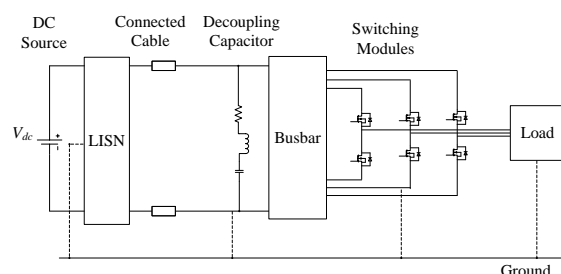

Figure 1. The analyzed inverter system

It is commonly recognized that EMI must be addressed as early as possible during the design phase, and not after[1]. A frequency-domain approach reduces computational effort, and is recommended as the preferred approach [2],[3]. Earlier attempts have used trapezoidal waveforms and simplified lumped

propagation circuits to approximate the equivalent circuit model [4]-[6]. But they can only predict the differential mode(DM) EMI spectra in low frequency region because none of parasitic elements are included. A unified DM EMI prediction model of three phase inverter circuit, which takes parasitic elements into account was proposed in[7]. But the total dc side DM noise is calculated by superposition of DM noises caused by each phase, which ignores phase angle differences of each phase, may lead to inaccuracy in low switching frequency range. In this paper, a new DM noise calculation method is proposed by adding a dc oscillation part in parallel with the conventional dc pulse train part as the current source. Both the noise path and noise current are modeled, and the calculated DM noise is compared with the experimental result.

2. Noise path modelling

2.1 Passive components

2.1.1 Load Inductors with Cables. Three 150uH inductors are connected in 'Y' type as the load. The DM impedance can be measured using impedance meter Agilent E4990A. By multiple sections linear RLC circuits fitting methods [8], DM equivalent circuit model is developed as shown in figure 2. Comparison of measured and modeled DM impedances of the load is shown in figure 3, and they match well with each other.
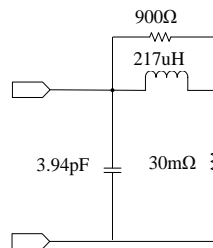
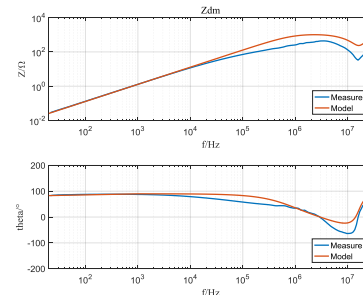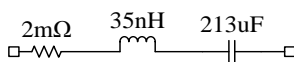

Figure 2. DM impedance equivalent circuit

Figure 3. Comparison of measured and simulated DM impedances of the load

2.1.2 DC bus Capacitor. The impedance of the dc bus capacitor (Photograph shown in figure 4(a)) can be measured by impedance analyzer Agilent E4990A. By the equivalent circuit in figure 4(b), the capacitor impedance can be calculated. The comparison of measured and simulated capacitor impedances is shown in figure 5, and they match well with each other.

(a)

(b)

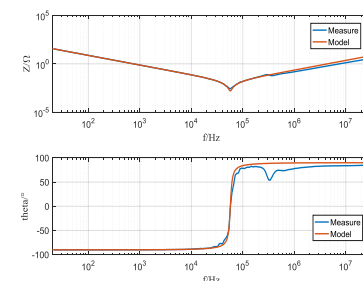**Figure 4.** Capacitor (a) photograph, (b)equivalent circuit.

Figure 5. Comparison of measured and simulated capacitor impedances.

2.1.3 LISN. The line impedance stabilization network (LISN) is a circuit which measures noise terminal voltage. Schwarzbeck NNLK 8130 is used and its equivalent circuit can be obtained from its manual.

2.2 Parasitic components

2.2.1 MOSFET parasitic elements. The switching module used is the commercially available 1.2 kV/300A SiC power MOSFET module named CAS300M12BM2 from Cree (now Wolfspeed). Its stray inductance and output capacitance can be obtained from the datasheet. The on-state resistance can be read from experimental waveforms.

2.2.2 DC bus parasitic inductances. By method of moments (MoM)-based software Ansys Q3D, the inductances of DC bus can be extracted. The coupling effect between each phase has been taken into account and table 1 shows the extracted parasitic inductances.

Table 1. Extracted parasitic inductances by Q3D

Inductance（nH）						
	1N	2N	3N	1P	2P	3P
1N	41.21	22.425	16.379	27.128	17.003	11.956
2N	22.425	29.616	22.389	16.817	20.48	16.779
3N	16.379	22.389	41.119	11.958	16.973	27.039
1P	27.128	16.817	11.958	29.985	14.494	9.7584
2P	17.003	20.48	16.973	14.494	20.892	14.454
3P	11.956	16.779	27.039	9.7584	14.454	29.861

The above table can be simplified based on the concept of loop inductance given in (1).

$$L_{mm} = L_{m+m+} + L_{m-m-} - 2L_{m+m-}$$
$$L_{mn} = L_{m+n+} + L_{m-n-} - L_{m-n+} - L_{m+n-}$$

(1)

Where, $m = A, B, C, n = A, B, C$. Then a reduced parasitic inductance matrix that is more convenient for DM noise analysis can be derived as shown in table 2.

Table 2. Reduced parasitic inductance matrix

Equivalent Inductance(nH)			
	A	B	C
A	16.94	3.1	2.22
B	3.1	9.55	3.09
C	2.22	3.09	16.9

3. Differential mode noise source modelling

For SiC MOSFET converters, oscillations on current sources caused by parasitic elements create a considerable impact on EMI spectra [9] and it is non-ignorable. In this paper, the current source seen by dc link is decoupled into the pulse train source and the parasitic oscillation source.

The pulse train current source $I_{dc}(t)$ can be expressed by the phase currents I_a(t), I_b(t), I_c(t) and the switching functions S_a, S_b, S_c as,

$$I_{dc}(t) = S_a I_a(t) + S_b I_b(t) + S_c I_c(t)$$

(2)

$$S_k = \begin{cases} 1 & \text{when top switch is on} \\ 0 & \text{when bottom switch is on} \end{cases} \quad k = a,b,c$$

By emulating the PWM process in one fundamental cycle, and synthesize the time domain current pulse waveform, the waveform of $I_{dc}(t)$ can be obtained. Then the spectrum of $I_{dc}(t)$ can be calculated using fast Fourier transform (FFT) approach.

Since the DM oscillation current is not affected by the fundamental current direction [7], a unified high frequency inverter phase leg linear model can be drawn, as shown in figure 6. By superposition theorem, all three phase legs of the inverter are unified together to one equivalent circuit [10], as shown in figure 7.
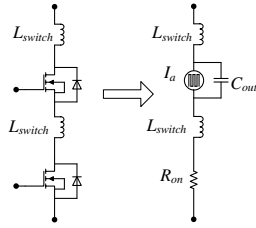

Figure 6. Unified high frequency linear model of inverter phase leg.
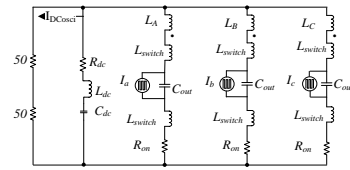
Figure 7. DM oscillation noise source induced model.

In figure 7, C_{out} is the device output capacitance, L_{switch} is the device lead inductance and R_{on} is the device on resistance. The propagation transfer functions $Z_a(f)$, $Z_b(f)$, $Z_c(f)$, from $I_a(f)$, $I_b(f)$, $I_c(f)$ to $I_{DCosci}(f)$ respectively are analyzed in Simplorer and plotted in MATLAB, as shown in figure 8. The amplitude and resonant frequency of the first resonant points are identical for the transfer function of three phases. To simplify further analysis, it is assumed that $Z_a(f)=Z_b(f)=Z_c(f)$.

The current source can be approximated as a pulse series waveform with equal rise time and finite fall time τ_r. During switching transition, the step of the current source can be given as,

$$I_{ks}(t) = \frac{I_k \cdot t}{\tau_r} - \frac{I_k \cdot (t - \tau_r)}{\tau_r}, k = a,b,c \tag{3}$$

Where I_k is the value of current source, and its peak value is I_m. To get Fourier series result, the switching period T_s must be considered. For SVPWM, there are 6 switching transitions in one switching period. Then the peak spectrum of parasitic ringing current source I_{DCosci} can be derived as,

$$I_{DCosci}(f) = \frac{I_m(1 - e^{-s\tau_r})}{\sqrt{2}\tau_r s^2 T_s / 6} \cdot Z_a \tag{4}$$

Where $\sqrt{2}$ accounts for the rms value of spectrum analyzer.

Thus, the differential mode noise source, which is combined of pulse train source and parasitic oscillation source has been modeled.

4. Validation through calculation and experiment

Based on the path and source models derived in sections 2 and 3, the DM EMI spectrum can be calculated using the equivalent circuit shown in figure 9, which is a modification of the model proposed by Nava[2].
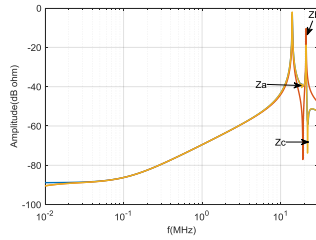
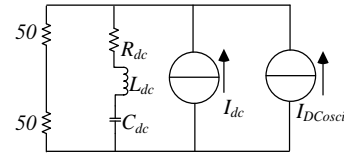**Figure 8.** Propagation path transfer functions of DM oscillation noise.

Figure 9. DM noise equivalent circuit.

The current noise between LISN and the inverter can be calculated as shown in (5).

$$I_{dm} = (I_{dc}(f) + I_{DCosci}(f))?Z(f) \tag{5}$$

Where, the propagation path transfer function can be written as,

$$Z(f) = \frac{R_{dc} + sL_{dc} + \dfrac{1}{sC_{dc}}}{R_{dc} + sL_{dc} + \dfrac{1}{sC_{dc}} + 100} \tag{6}$$

The experimental setup is shown in figure 10. The system adopts SVPWM strategy. The switching frequency is 7.5kHz, the dc bus voltage is 100V, and the output current peak value is 85A. A broadband current probe named BCP-620, whose bandwidth is 10kHz-500MHz, is used to measure DM noise current between the LISNs and the inverter. The measurement configuration is shown in figure 11, where one of the power lines go through the current probe directly, while the other go through the current probe after winding. The output of the current probe is connected to oscilloscope. After getting the noise waveforms, by FFT approach, the measured DM noise spectrum can be obtained.
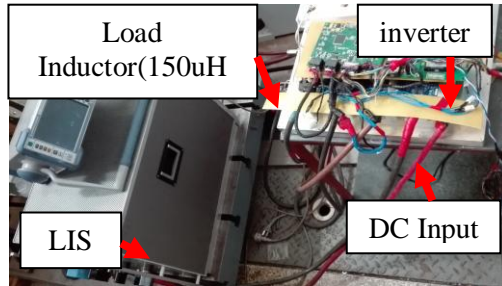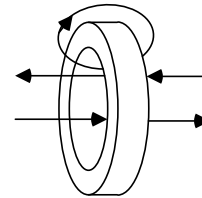

Figure 10. Experimental setup.

Figure 11. Measurement configuration for DM noises.

Comparison of DM noise spectra between frequency-domain calculated result and the measured result is shown in figure 12. The frequency-domain calculated result is shown in figure 12 (a), and the experimental result is shown in figure 12 (b).The black waveform in figure 12 (b) is the background noise spectrum of the oscilloscope.
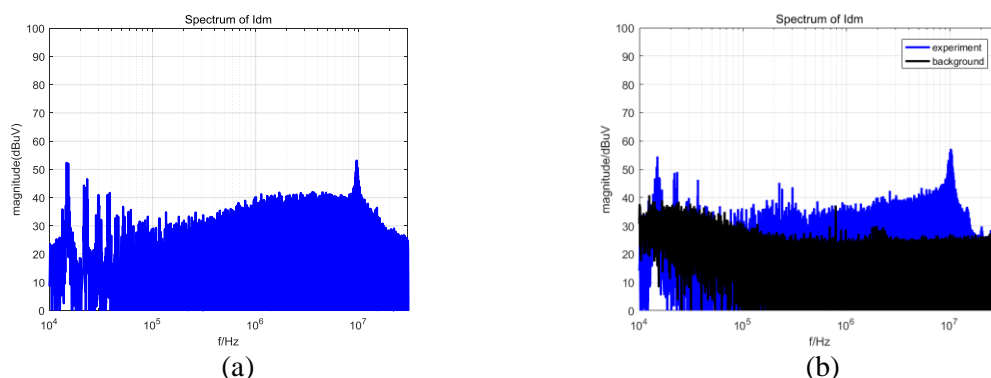
Figure 12. Comparison of DM noise spectra (a) frequency-domain calculated result(b)experimental result

In frequency range from 10kHz to 10MHz, the calculated result by the proposed method matches well with the experimental result. While the difference in 10MHz to 30MHz frequency range may be caused by approximations and ignorance of parameters in the model. However, the comparison of simulation and measurement results indicates good conformance with errors less than 10dB in most of the frequency ranges. Generally, the experimental results show that the proposed EMI calculation method can predict the DM noise of the three-phase inverter system well.

5. Conclusion
A new DM noise frequency prediction method for three-phase SiC MOSFET inverter system has been presented in this paper. Both noise path and noise source are modeled. The impedance analyzer is used to measure and model impedances of passive components. For the busbar, whose impedance is hard to measure, MoM-based tool Q3D is used for parasitic parameters extraction. The noise current source is decoupled into pulse train part and oscillation part, and the oscillation source induced equivalent circuit is identified. Calculation and experimental comparison result shows that the proposed frequency-domain calculation result matches well with the experimental result in both switching frequency range and resonant frequency range.

Acknowledgments

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