

Modeling and Suppression Techniques of Common Mode Conducted EMI Noise in Power Electronic Converters

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Modeling and Suppression Techniques of Common Mode
Conducted EMI noise in Power Electronic Converters

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Abstract

Electromagnetic interference (EMI) issues in power electronic converters are associated with the high-speed operation of semiconductor switches. On one hand, the emergence of wide-gap semiconductors (like Silicon Carbide and Gallium Nitride) with high power and high switching frequency made the EMI emission more prominent. While on the other hand, the modern power converter is developing towards miniaturization and lightweight, which imposes a harsh requirement on the volume of EMI filters. For these reasons, an accurate EMI noise model is essential to study the EMI generation mechanism, evaluate the EMI problems, and predict the effectiveness of the noise filter.

This thesis studies the modeling and mitigation techniques of common mode (CM) conducted EMI noise in power converters. The research objects include DC-DC non-isolation converter, DC-DC isolation converter, and DC-AC inverter. Due to the characteristics and application scenarios of different converters, behavioral modeling method and analytical modeling method are investigated. About the noise mitigation research, a noise filter design method based on the behavioral model and an optimized impedance balancing technique based on the analytical model are investigated in this study.

In chapter one, the research background, basic knowledge about EMI measurement and standards, state of art review of noise modeling and suppression technologies in power electronic converters are introduced.

In chapter two, an improved CM noise modeling method based on terminal port theory

is introduced. Compared with previous behavioral modeling methods, the proposed method has a clear inner structure and better accuracy. Specifically, the clear structure of the two-port terminal can help strengthen the connection between the noise model and the physical circuit. In addition, the proposed interpolation algorithm can improve model accuracy by reducing impedance error. This method is utilized in both the boost converter and the single-phase inverter. Using the proposed noise model, not only the EMI noise filter at the input side but also the harmonic filter at the output side are evaluated as well.

In chapter three, a novel analytical modeling method is developed to analyze the CM noise in a flyback converter. The noise generation mechanism is investigated at first and the simplified noise model is proposed accordingly. Compared with the former work, the accuracy of this proposed noise model is improved in two aspects: 1. The ringing voltages generated with turn on and off of the switch are been considered in the noise source model; 2. The leakage inductance of the flyback transformer is taken into account when analyzing the noise current propagation. The former one could improve the accuracy of the noise source, especially the high frequency segments. The latter one firstly discussed the leakage inductance, which was ignored by the previous studies but plays a certain role in the noise current transmission path. The experimental spectrum could successfully demonstrate the feasibility and effectiveness of the proposed noise model.

In chapter four, the CM noise mitigation technique with balanced impedance is analyzed and modified. The analytical noise modeling method, which is proposed in the third chapter, is utilized to analyze the balanced boost converter, which helps to make clear the EMI noise generation mechanism and investigate the noise current propagation. In this chapter, the noise source attenuation degree is evaluated by combining the noise

spectra and the impedance information. Therefore, the effectiveness of the balance technique could be directly evaluated. In addition, two optimized schemes were proposed to deal with the noise peak, which is a major drawback of the previous work. Experimental results show that the noise peak has been attenuated at the conducted frequency band with proposed schemes, and the noise suppression effect is even better than the normal choke filter.

In chapter five, the entire thesis is summarized and some ideas and proposals are presented for future work.

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Chapter 1 Introduction

1.1 Background

With the application of controllable semiconductor devices that have high power and high switching frequency, the power electronics technology has been more and more widely used in industrial productions and daily life. However, with the development of power electronic technology towards high frequency, miniaturization, and multi-function integration, the electromagnetic compatibility (EMC) problem has become more severe and complicated. An electronic system is required to function compatible with other electronic systems and not produce or be susceptible to interference with its surrounding environment [1].

EMC could be classified into electromagnetic interference (EMI) and electromagnetic susceptibility (EMS). EMI means “without causing electromagnetic interference to other devices”, and is often used in pair with the expression “emission”. Conversely, EMC stands for “subject to electromagnetic interference from other devices”, and emphasizes “immunity”. The relationships of EMC, EMI, and EMS could refer to Fig.1.1.

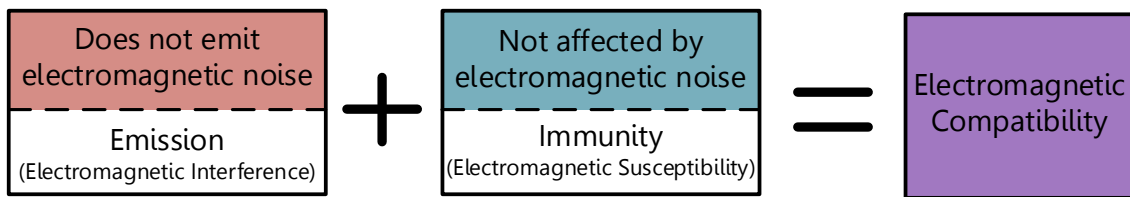


Fig.1.1 The relationship of electromagnetic compatibility (EMC), electromagnetic interference (EMI), and electromagnetic susceptibility (EMS).

According to the electromagnetic energy types, the EMI noise can be categorized as conducted emission and radiated emission. The conducted noise is generally propagated

through wires and PCB wiring, and the radiated noise is emitted through the air. Correspondingly, the EMS can be divided into conducted susceptibility and radiated susceptibility [3]. The propagation route of the EMI noise is shown in Fig.1.2.

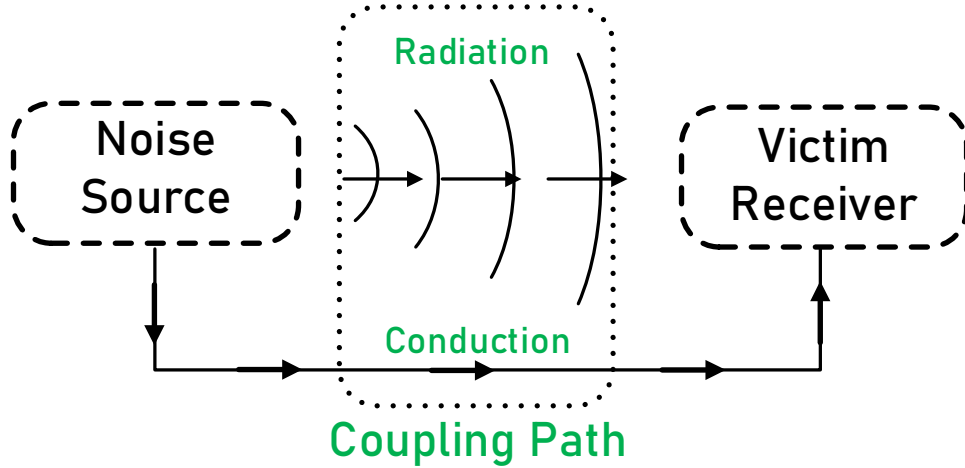


Fig.1.2 Propagation routes of conducted and radiated EMI noise.

1.2 Measurement and Standards of Conducted EMI Noise

For electrical devices, the EMC is not only an important issue related to the performance but also means it has to satisfy the standards and regulations in legal. Nowadays, there are many international and regional organizations, and relevant standards and specifications have been published. Like the Federal Communications Commission (FCC) in the United States, the Japanese Industrial Standards (JIS) in Japan, the Guobiao standards (GB) in China, the International Special Committee on Radio Interference (CISPR), which is a part of the International Electrotechnical Committee (IEC) in European Union. Fig 1.3 briefly illustrates the EMC-related standards [4].

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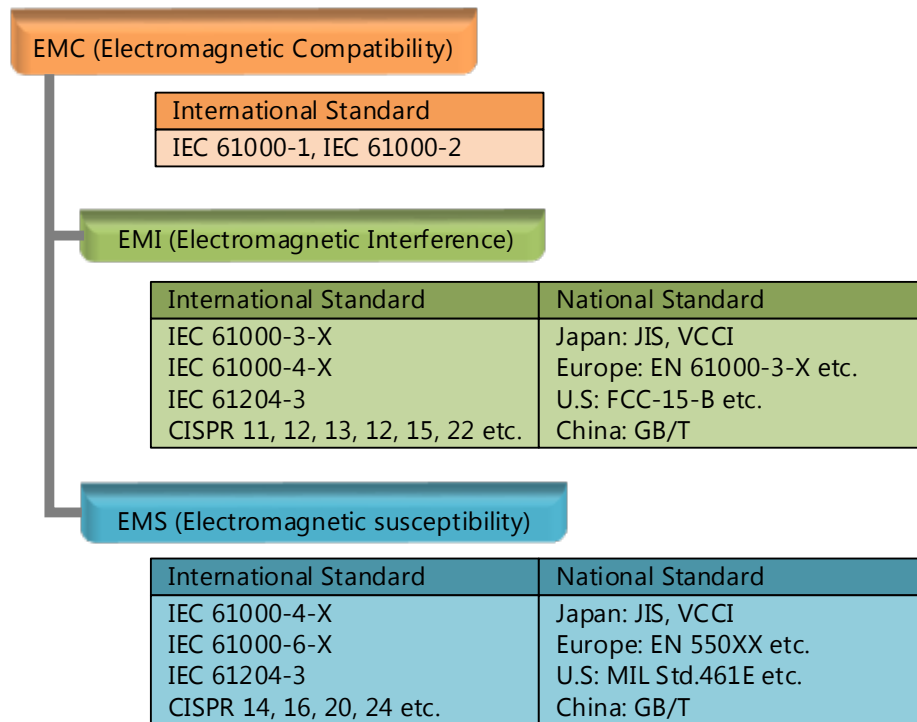


Fig.1.3 International and regional EMC standards.

CISPR and FCC give detailed standards for conducted EMI noise measurement. Fig.1.4 shows a typical test setup for conducted noise. As shown in the figure, a standard Line Impedance Stabilizer Network (LISN) is placed between the equipment under test (EUT) and the power supply.

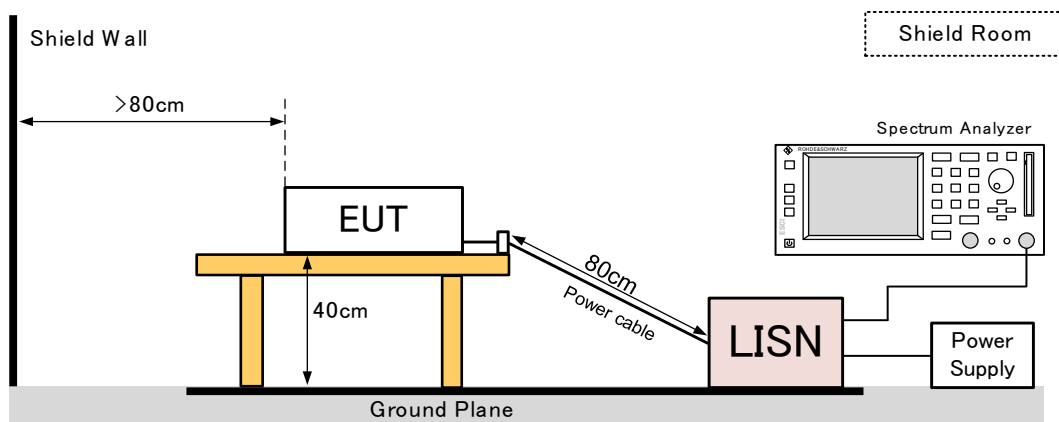


Fig.1.4 Standard test setup for conducted emission measurements.

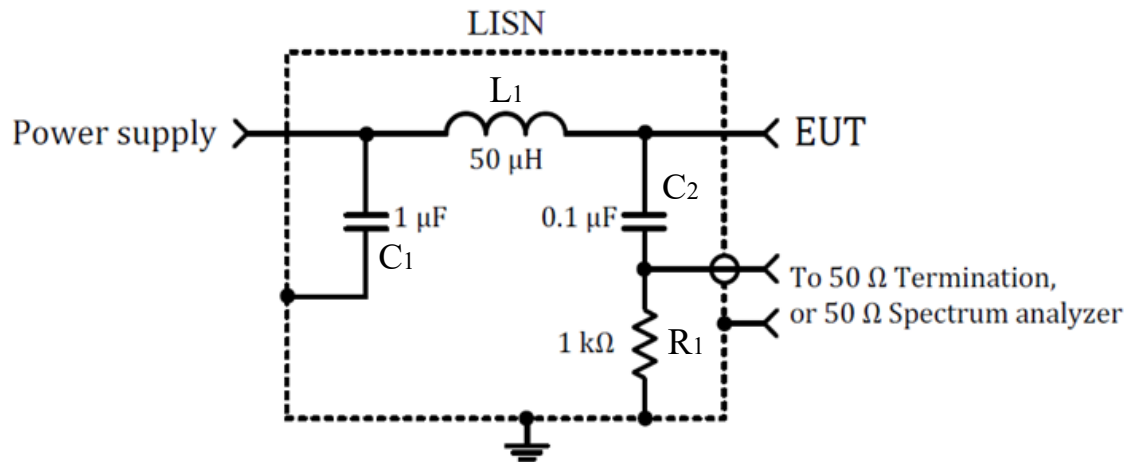


Fig.1.5 Schematics of the Line impedance stabilization network (LISN) for the measurement of conducted noise.

The schematics of LISN is shown in Fig.1.5 [5]. And the role of LISN is mainly reflected in two aspects:

1. Prevent the outside noises from contaminating the measurement. Specifically, L_1 ($50\ \mu\text{H}$) and C_1 ($1\ \mu\text{F}$) in the block could filter the noise coming from the power system within the conducted noise frequency band.
2. Present a constant impedance to the product terminals at the frequency range of conducted noise measurement.

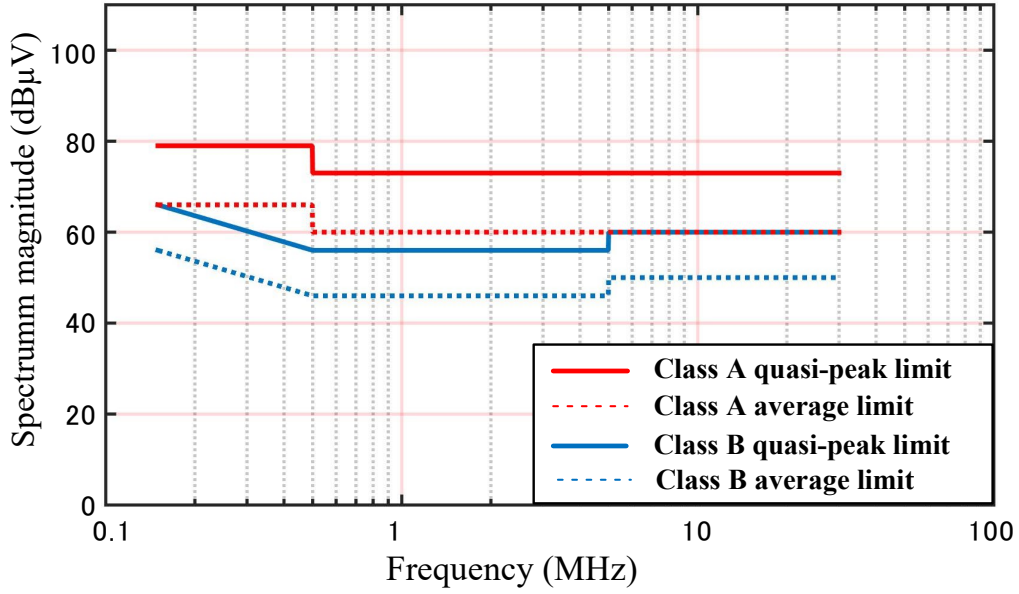


Fig.1.6 The FCC/CISPR conducted emission limits.

Fig.1.6 shows a plot of both the average and the quasi-peak FCC/CISPR conducted noise limits. Note that both quasi-peak and average limits are given. The quasi-peak measurements weigh the signal according to the repetition rate, whereas the average value represents the average amplitude of each signal component.

The commonly used units in EMC measurement are dBμV and dBμA, which are based on the voltage and current, respectively. The decibel (dB) is not an absolute value but represents a logarithmic ratio between two quantities. Take microvolt (μV) and microampere (μA) as references, the dBμV and dBμA could represent as in (1.1) and (1.2).

$$\begin{aligned}
 \text{dB}\mu\text{V} &= 20 \log_{10} \left(\frac{V}{1\mu\text{V}} \right) \\
 &= 20 \log_{10} V - 20 \log_{10} 1\mu\text{V} = 20 \log_{10} V + 120
 \end{aligned} \tag{1.1}$$

$$\begin{aligned}
 \text{dB}\mu\text{A} &= 20 \log_{10} \left(\frac{A}{1\mu\text{A}} \right) \\
 &= 20 \log_{10} A - 20 \log_{10} 1\mu\text{A} = 20 \log_{10} A + 120
 \end{aligned} \tag{1.2}$$

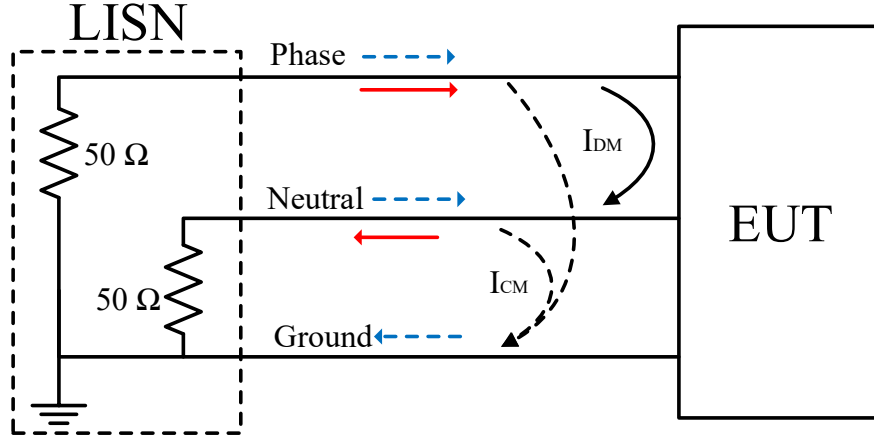


Fig.1.7 Propagation route of CM current and DM current.

In terms of the propagation route, the conducted noise is decomposed into common mode (CM) noise and differential mode (DM) noise. Correspondingly, there are common mode voltage, differential mode voltage, common mode current and differential mode current. As is shown in Fig.1.7, DM currents flow between neutral and phase conducting lines whereas common mode currents flow through the phase and neutral conducting lines, and take ground wire as the return path. Therefore, the current of P phase (I_P) and N phase (I_N) could be represented as:

$$I_P = I_{cm} + I_{dm} \quad (1.3)$$

$$I_N = I_{cm} - I_{dm} \quad (1.4)$$

where I_{cm} and I_{dm} denote the CM current and DM current respectively, depicted by the dashed and solid lines in Fig. 1.7. It can also denote the CM current and DM current expressions. That gives:

$$I_{cm} = \frac{1}{2}(I_P + I_N) \quad (1.5)$$

$$I_{dm} = \frac{1}{2}(I_P - I_N) \quad (1.6)$$

1.3 Lecture Review of Noise Modeling and Suppression Methods

With the development of the modern electronic system, the probability of generating and receiving EMI noise has greatly increased. Therefore, EMI and EMS have become important issues for electronic engineers and researchers.

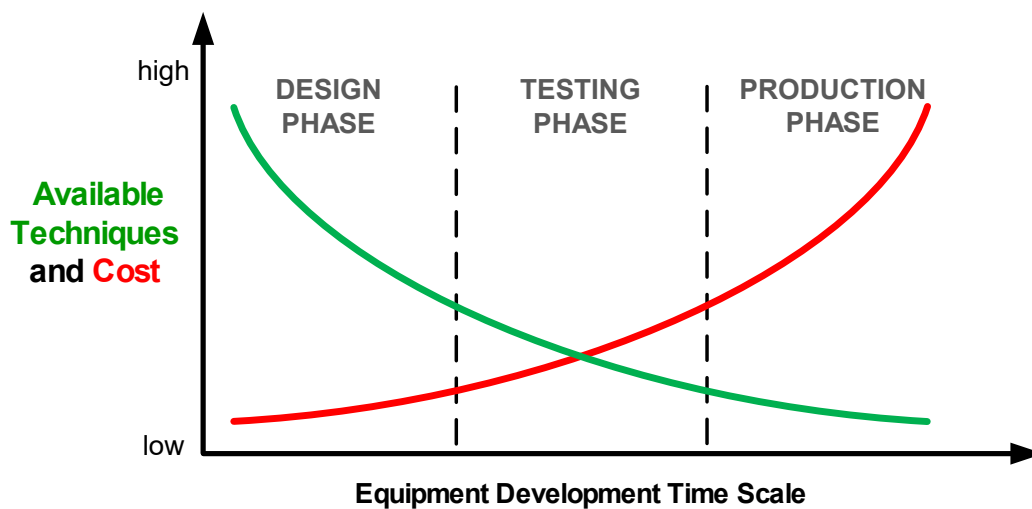


Fig.1.8 The relationships between the development stage and the available noise reduction techniques & cost.

The relationships between the development stage and the available noise reduction techniques & cost are shown in Fig.1.8. The later we realize the EMI problem, the fewer solutions we can adopt, and the higher the overall cost. As a useful tool to investigate the conducted noise, a precise EMI noise model could help to analyze spectrum result and adopt appropriate strategies in the developing stage [7]. The following is the lecture review of the noise modeling and reduction techniques.

1.3.1 Noise Modeling Methods

The conducted EMI noise modeling methodologies of the power converter can be

divided into two categories: time-domain modeling and frequency-domain modeling. The classification of the conducted EMI modeling techniques is shown in Fig.1.9.

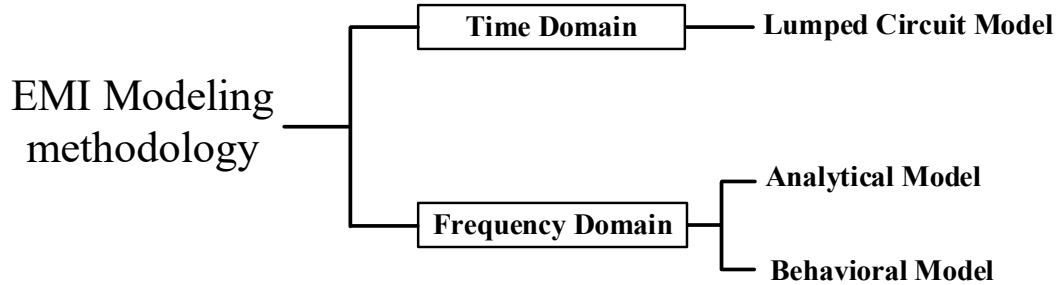


Fig.1.9 Classification of the conducted EMI modeling methods.

The early-stage time domain modeling method is based on the physical model of the semiconductor device and utilized the lumped circuit to represent the stray capacitances [8]-[15]. A physical-based model consisted of a set of nonlinear, high-order, and highly coupled multivariable differential equations [9]. Also, this model needs an amount of the device parameters from the datasheet and device test, not to mention the requirement of relevant knowledge about the semiconductor material properties and operating mechanism of the semiconductor switch.

The time domain modeling technique may have enough accuracy with sufficient details, which are difficult to extract. Therefore, the following researcher optimized the modeling process and utilized the simulation software to build a physical model, like Saber and PSPICE [11]-[14]. This initiative has extended the application scope of the time domain modeling approach. However, it takes time to simulate the circuit operation for at least one fundamental cycle. Besides, the detailed parameters extraction is difficult in practice and the time-domain simulation process may encounter convergence risk.

As is shown in Fig.1.9, there are two major branches in the frequency domain modeling

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method: behavioral modeling method and analytical modeling method. The behavioral modeling method was firstly designed to represent the operation characters of the phase leg and was then extended to multiple kinds of converters [16]-[31].

Instead of establishing the CM noise model and the DM noise model respectively, the behavioral model could predict the CM and DM noise in one integrated model. Both the two-terminal (one port) model and the three-terminal (two ports) model have been extracted based on Thevenin or Norton theorems. The equivalent EMI noise emission model in [19] is shown in Fig.1.10.

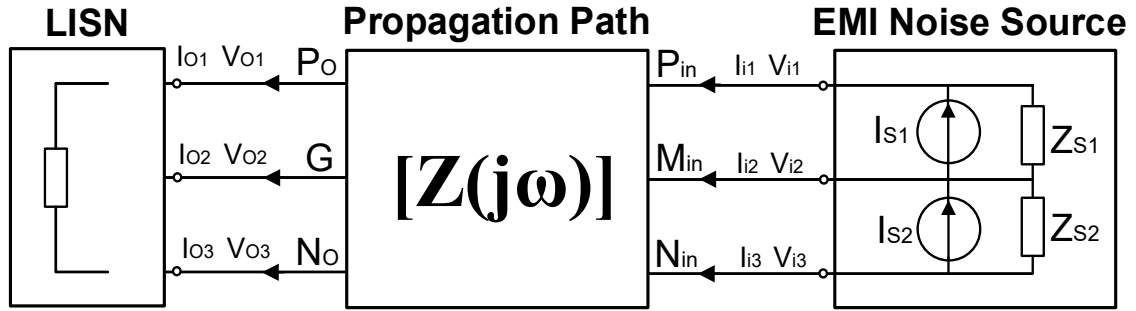


Fig.1.10 Equivalent noise model in [19].

The experimental results show that this method would work well facing simple topologies, like non-isolated converters and half-bridge circuits. Nevertheless, without the analysis of noise propagation, the impedance identification has always been troubled by the lack of physical meaning. This problem becomes especially significant when modeling complex circuits. The impedance error would continue to affect the precision of the prediction results. Even repeating measurements under different series and shunt conditions may not help to reduce the error [36].

On the other hand, the analytical modeling method is based on the analysis of noise generation mechanisms for the individual specific subject. This method emphasizes

Chapter 1

theoretical analysis and processing data with mathematical tools, like Fourier transmission, double-Fourier transformation, wavelet transform, and so on [42]-[53]. The noise source in the former analytical modeling method was simplified with ideal geometric shapes, and use lumped parameters to represent the circuit impedance [42]-[45]. However, over-simplification damaged the accuracy of the noise model and more details should be considered. [46] has applied the analytical modeling method to a boost power factor correction (PFC) converter, and predicted the attenuation effect of the noise filter successfully. However, the full-frequency (150 kHz-30 MHz) range prediction accuracy needs further improved.

More details about the analytical modeling method and the frequency domain modeling method would be analyzed in the following chapters.

1.3.2 Noise Suppression Techniques

The noise suppression techniques could also be divided into two categories, as is shown in Fig.1.11. The first one is increasing the impedance of the noise current propagation or building the bypass circuit. Like shielding and the noise filter, including passive filter, active filter, and hybrid filter. Also covering changing the topologies, like the impedance balancing technique, the balanced topology, and so on. The second aspect is mitigating the noise source, namely the semiconductor switch. For example, the specific Pulse Width Modulation (PWM) generation method, like the widely used random PWM and multitudinous modulation methods [62].

Introduction

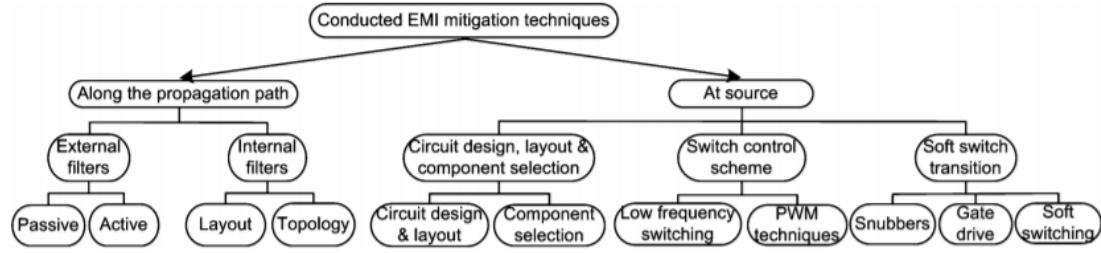


Fig.1.11 Conducted EMI noise mitigation techniques.

In principle, the noise filter cannot prevent the occurrence of interference. The main object of the EMI filter is to minimize the interference effect on the other components of the power converter [63]. Generally speaking, the noise filter is added to the signal or power line, to increase the propagation impedance and limit the noise current. The EMI filters can be broadly classified into three kinds: passive filter, active filter, and hybrid filter [64].

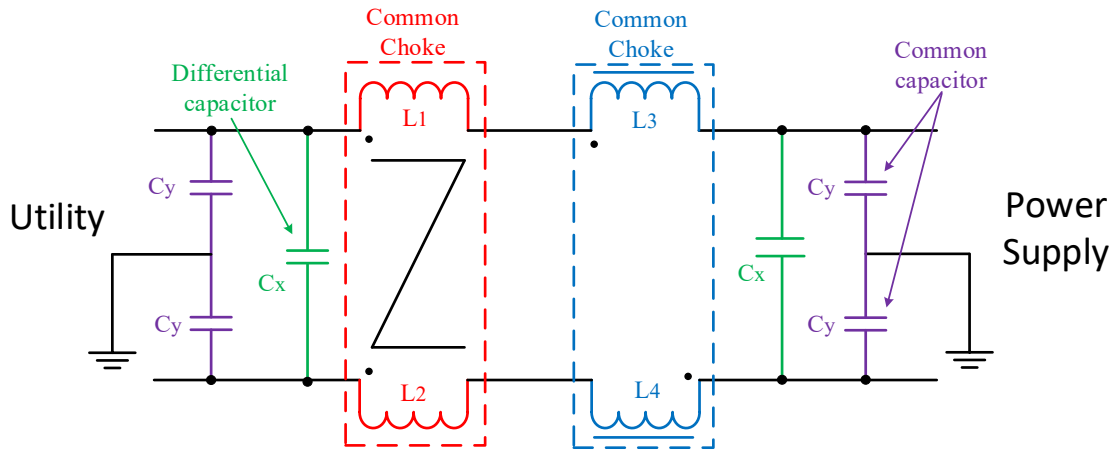


Fig.1.12 Schematic circuit diagram of a typical passive EMI filter system.

The schematic circuit diagram of a typical passive EMI filter system is shown in Fig.1.12. Here the DM filter is consists of a differential capacitor and a differential choke filter (normal inductor). The CM noise filter is constituted by a series CM choke in the phase lines and the parallel capacitances between the phase line and the ground. Many

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scholars have discussed the topology and the parameters design of the passive filter circuit [65]-[68].

Compare with the passive filter, the active filter saves the bulky and more effective in some scenarios [69]-[71]. However, the dependency of the current/voltage sensor and the sensitivity for the feedback system may increase the instability of the whole system. Besides, the hybrid active filter was researched in many publications [72]-[75], which has combined the concepts of both passive filter and active filter.

As the EMI noise is generated with the behaviors of the switch, it is natural to think about how to attenuate the strength of the noise source. The random PWM technique is one of the commonly used methods [76]-[79]. This technique could change the switching frequency continuously and spread the noise peak on the spectrum.

In recent years, researchers have also studied other novel noise mitigation technologies. For example, the impedance balancing method [80]-[97]. In this technique, the switch is surrounded by the impedance bridges while the impedance ratios of each two bridge arms remain the same. Therefore, the noise source could be greatly attenuated and the explanation can stem from the Wheatstone bridge theory. This method has merits like no need external noise filter and easy to implement. Meanwhile, there also exist some challenges and need further research [97].

1.4 Research Objectives and Dissertation Outline

The main motivation of this research is to investigate and mitigate the CM conducted EMI noise in power electronic converters. Research objects include the Non-isolation DC-DC converter, DC-AC converter, and isolated DC-DC converters. In addition,

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multiple noise modeling methods, including behavioral modeling method and analytical modeling method, meanwhile the noise mitigation technique with balanced impedance are studied in this work.

In conclusion, different converters have their characteristics and different scopes of application. Facing specific objects, it is necessary to design EMI noise schemes in combination with the circuit characteristics and the definite requirements.

This thesis is organized into five chapters. The background knowledge, EMI measurement and standards, state of art review of noise modeling and reduction techniques were provided in the first chapter.

In chapter two, a CM noise modeling method based on terminal port theory is introduced. This method is utilized in both the boost converter and the single-phase inverter. With the proposed noise model, not only the EMI noise filter at the input side but also the harmonic filter at the output side are evaluated at the same time. Compared with the former work, this study clarified the internal structure and improved the accuracy of the CM impedance with the proposed interpolation algorithm.

In chapter three, a novel analytical modeling method is utilized to analyze the CM noise in a flyback converter. The noise generation mechanism is investigated at first and the simplified noise model is proposed. Compared with the former work, the accuracy of this proposed noise model is improved in two aspects: firstly, the ringing voltage generated with turn on and off of the switch is been considered in the noise source model; on the other hand, the leakage inductance of the flyback transformer is taken into account when analyzing the noise current propagation. The former could improve the accuracy of the noise source, especially in the high frequency parts. Besides, this study firstly

Chapter 1

discusses the leakage inductance of the flyback transformer, which was ignored by the previous studies but plays a certain role in the noise current transmission path.

In chapter four, the CM noise mitigation technique with balanced impedance is analyzed and optimized. The analytical noise modeling method proposed in the third chapter is utilized to analyze the balanced boost converter, which could help to investigate the noise current propagation in the balanced boost converter. In this research, the noise source attenuation degree is evaluated by combining the noise spectra and impedance information. In addition, two optimized schemes were proposed to deal with the noise peak, which is a drawback of the previous work. The balanced boost with proposed schemes is even better than the additional ordinary passive noise filters.

In Chapter five, the entire thesis is summarized, also some ideas and proposals about future work are presented.

Chapter 2 CM Noise Modeling Method with Two-port Terminal Theory

2.1 Introduction to Behavioral Modeling Method

As was introduced in chapter one, the noise modeling method could be divided into two kinds: time domain and frequency domain. Moreover, the behavioral modeling method is the main branch of the frequency domain modeling method and has become an important issue in recent years.

The early-stage noise model is usually based on the behavior of the device, but lack practicability [11]. To solve this problem, Liu firstly proposed the modular-terminal-behavioral (MTB) equivalent noise source modeling approach in [16]. Here, a three-terminal noise model, which consists of equivalent noise current sources paralleled with impedances, has been extracted from a phase leg module. The proposed structure contains a five-terminal passive network and was designed for analyzing the CM and DM noise simultaneously. Firstly, the equivalent three-terminal Norton model extracted from a half-bridge phase-leg is shown in Fig.2.1, where I_{S1} and I_{S2} are two separate current sources.

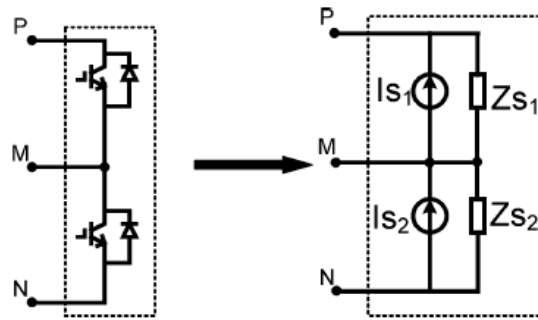


Fig.2.1 Equivalent Norton circuit for a half-bridge module.

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However, this approach is complex to obtain and the relationships between I_{S1} and I_{S2} are not clear. This structure was selected just to satisfy the port number of the impedance matrix: P, M, and N. Moreover, the order of the impedance matrix would become four or more when considering the ground. Therefore, it is hard to determine and measure the impedance parameters. In the following work [17]-[19], Liu has continuously refined the modeling process, applied this modeling approach to other converter topologies and discussed the relationships between the switching conditions and the EMI noise.

Baisden and Bishnoi have made improvements on the basis of the MTB modeling method, and furtherly proposed the General Terminal Modeling (GTM) approach and Unterminated Behavioral Modes (UBM) [20]-[25]. These improvements are mainly reflected in the noise model accuracy and practicality. The accuracy of the model has been increased to 30 MHz and the physical meaning of the impedance matrix has also been investigated. The UBM is also utilized to model the transmission line and could provide convenience for passive filter selection. However, the impedance measurement still needs calibrating processes under some shunt and parallel situations.

The following scholars from this laboratory have continued the research, adjusted the behavioral noise model to adapt to the trend of the WBD devices and more complex circuit topologies [26]-[28]. Especially, the proposed approach has been verified when the GaN-based three-phase inverter operating at 500 kHz. Nevertheless, the impedance value still needs reprocessing with the lengthy optimization algorithm.

At the same time, scholars were developing the behavioral modeling approach to investigate the system EMI noise modeling [29]-[31]. As a whole, the principles adopted in these papers are almost similar. The objective circuits were taken as one “black box”,

CM Noise Modeling Method with Two-port Terminal Theory

and each article has its focal point. [29] mainly investigated voltage source in the “Comprehensive Approach” and considered the influences of transmission lines and harness. [30] verified the network structure with three different PCB layouts. Nevertheless, prediction accuracy needs further improvement. The keynote of [31] was to illustrate the “Masking Impedance” role of an EMC input filter. Especially, the sensitivity of the impedance network was also studied.

As was analyzed, the identification and measurement of the noise source and impedance parameters are significant to the noise model, which largely determine the accuracy of the model [32]-[35]. Therefore, the CM and DM noise source impedances were studied separately by aggregate analyzing the results of the signal generator and spectrum analyzer in [32]. The calculated value matches well with the direct measurement results from the impedance analyzer. However, this approach still needs to fulfill some certain conditions, like the noise source impedance should be much larger or smaller than the value of the inserted component impedance. [34], [35] have also given the modified schemes to solve the conducted noise of the inverter circuit. The design process of the EMI filter design has proceeded but the high frequency noise needs further attenuation.

According to the lecture review above, the main ideas and development context of behavioral modeling methods have been summarized. There are two major merits of the behavioral modeling method. One is fitting for the complex circuit system and the other one is convenient for the noise filter design.

However, the existing literature still has some inadequacies and there is still space for improvement. Firstly, the link between the converter behavior and its physical parameters is vague and needs more research. Also, the relationships between multiple noise sources

Chapter 2

are not so clear. Furtherly, the identification and measurement of the “Black Box” are complex and lack practicability. Therefore, a study focused on these two points has been done and the details would be given in the following.

2.2 Terminal Port Theory Modeling in Boost Converter

To further improve the accuracy of the noise model, this study gives a CM noise modeling method based on terminal port theory. In this modeling process, the inner structure is discussed and the identification of equivalent noise source and impedance is studied [36]-[37]. This modeling method is not only verified by the DC-DC converter but also applied in a single-phase inverter. On the other side, this noise model is utilized to predict the effectiveness of the EMI filter at the input side and output side harmonic filter. Comparison between the measured and predicted results show that the proposed CM noise modeling method can provide accurate prediction even in complex conditions.

2.2.1 CM Noise Analysis of the Boost Converter

Fig.2.2 shows the configuration of the boost converter connecting with LISN. The LISN is placed between the DC power supply and the boost circuit, which can evaluate the conducted noise level by cooperating with a spectrum analyzer. C_{IN} and C_{OUT} are the input side and output side capacitance respectively, and L is the power inductor. The C_{CM} is the capacitor connecting with the ground and the drain of the MOSFET switch, which simulates the stray capacitors between the heatsink and the frame ground. These detailed parameters would be given in the following.

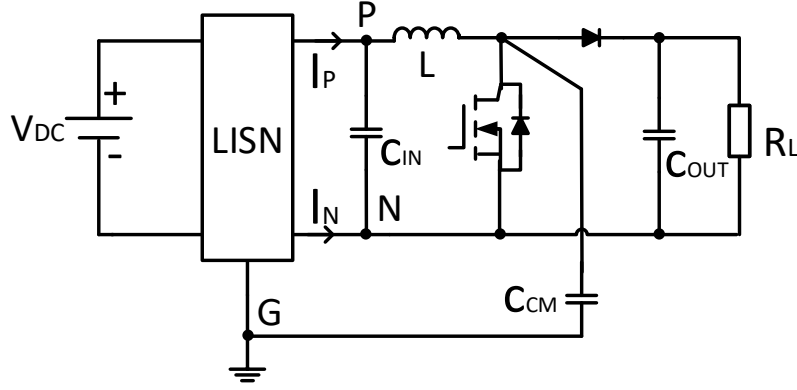


Fig. 2.2 Configuration of the boost converter and LISN.

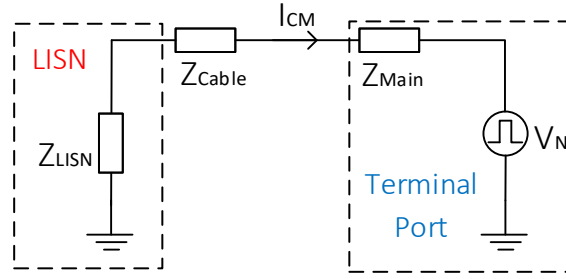


Fig 2.3 CM noise model of boost converter

The current flowing through the phase line and the neutral line are I_P and I_N , respectively. The value of I_P and I_N can be measured simultaneously by the current probes. And then the CM current can be calculated according to equations (1.5)-(1.6). As the existence of the electrolytic capacitor at input side, the noise measured by LISN is dominated by the CM noise.

As the modeling target circuit can be taken as a black box, the boost converter can be taken as a two-port terminal in Fig.2.3. Here the CM impedance of boost converter Z_{Main} and the equivalent noise source V_N make up this terminal port box. Z_{LISN} is The impedance of LISN. To further improve the accuracy of the model, the impedance of connection cable (Z_{Cable}) between EUT and LISN has also been taken into consideration.

2.2.2 Impedance Extraction in Noise Model

Based on the proposed CM model, the first step is to confirm the CM impedance. Scholars have done lots of innovations regarding the impedance measurements. Impedance matrix was used in [25], which also has referred to shunt-condition and series-condition to minimize the errors. [37] has discussed the influences caused by the switch's state. "ON", "OFF" and the transition state were investigated under the off-line state. Furthermore, [38] segmented the whole frequency band and repeated the measurement process several times.

Generally speaking, the common feature in CM impedance measurement is to adopt the equivalent circuit given by the impedance analyzer. However, deviation always appears regardless of the kind of impedance models. Therefore, a novel interpolation algorithm is proposed in this study to eliminate the errors, which could be breezily realized by software.

According to the CM impedance measuring principle, the positive and negative poles have been short connected and the CM impedance of the boost main circuit is largely dominated by the grounding capacitor. Meanwhile, the inductance of the PCB board would affect the corner frequency. The normal operation with impedance analyzer Agilent 4294A is shown in Fig.2.4.

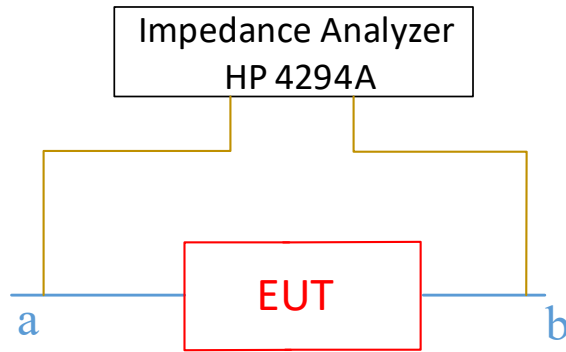


Fig.2.4 Impedance value confirm by impedance analyzer.

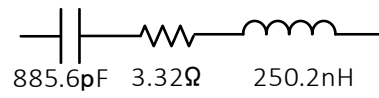


Fig.2.5 CM equivalent circuit of Z_{main} given by impedance analyzer.

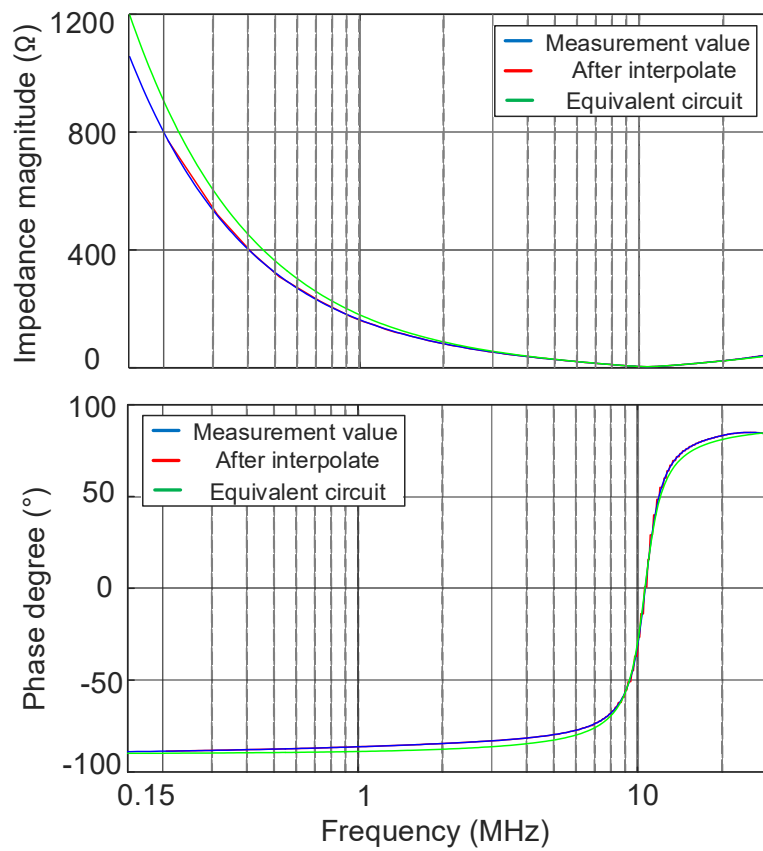


Fig.2.6 CM impedance magnitude and phase of Z_{main} .

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After selecting a proper noise model structure, the value of each segment in the impedance model could be given by the impedance analyzer. Fig. 2.5 shows the CM equivalent circuit of PCB board impedance Z_{main} obtained by impedance analyzer.

The comparisons of measured results, the value of the equivalent model, and the value after the interpolation algorithm are shown in Fig.2.6. The blue curve is the original data measured by an impedance analyzer. The red one is constituted by the data after interpolated the original impedance information. And the green one is the impedance curve of the equivalent circuit. The corner frequency turning from capacitive to inductive appears around 10 MHz, and the two curves are matching almost in all the frequency ranges which means the equivalent impedance model has been selected properly. But compared with the equivalent noise model, the interpolated data matches better.

Similarly, Fig. 2.7 and Fig. 2.8 have given the equivalent impedance model and frequency domain characteristics of Z_{Cable} , respectively. Still, errors of magnitude and phase angle existed even this impedance model has also been selected properly.

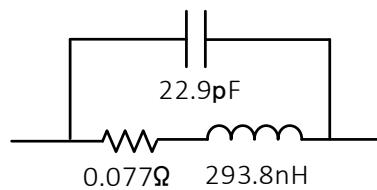


Fig.2.7 CM equivalent circuit of connection cable by impedance analyzer.

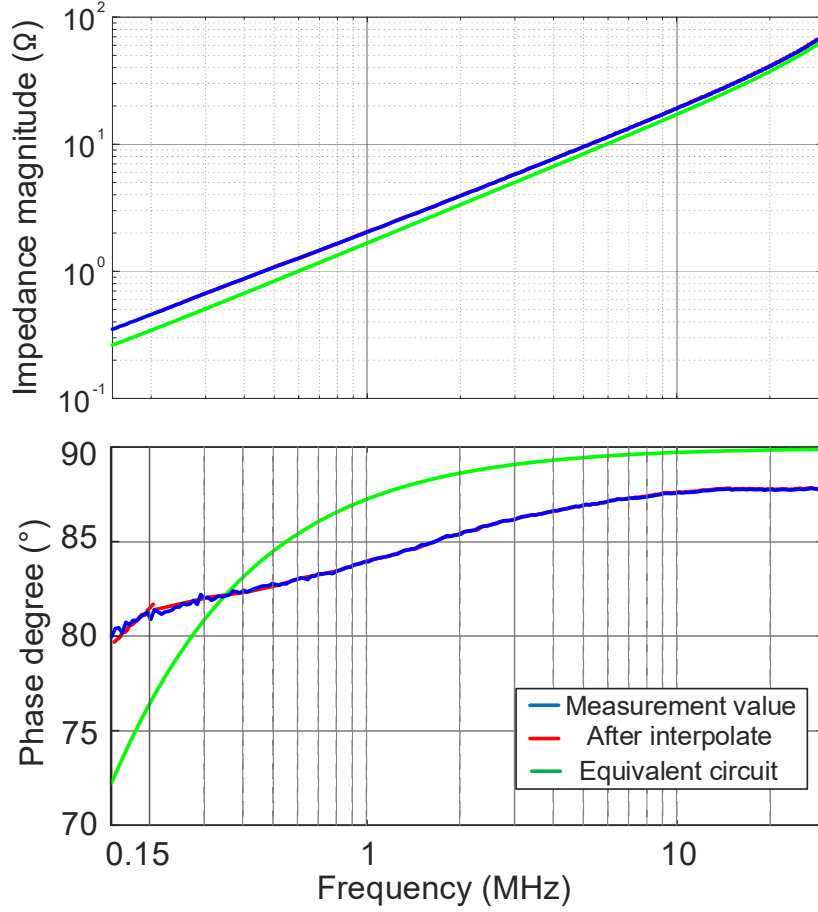


Fig.2.8 CM impedance magnitude and phase comparison of Z_{cable} .

Deviations caused by equivalent circuits could be seen clearly after comparison, and this deviation could not be eliminated. What's more, this error would deteriorate the whole noise model selection. The inappropriate RLC model may cause a great influence on the whole system modeling. Therefore, the interpolation algorithm based on Matlab has been utilized to tracking the active impedance curve and improve the accuracy. Meanwhile, this algorithm also enlarges the point's number and makes preparations for further noise source modeling.

2.2.3 CM Noise Source Extraction and Validation

Then is the data collection of noise current, which is another indispensable element to build the noise model. The experiment is essential for parameters determination. Fig.2.9 shows the experimental setup for capturing the voltage waveform and measuring the conducted EMI noise. The connection order is the same with the configuration in Fig2.2. In addition, the spectrum analyzer is connected with LISN to receive and analyze the signal, and the PC is used to display and record the spectrum results.

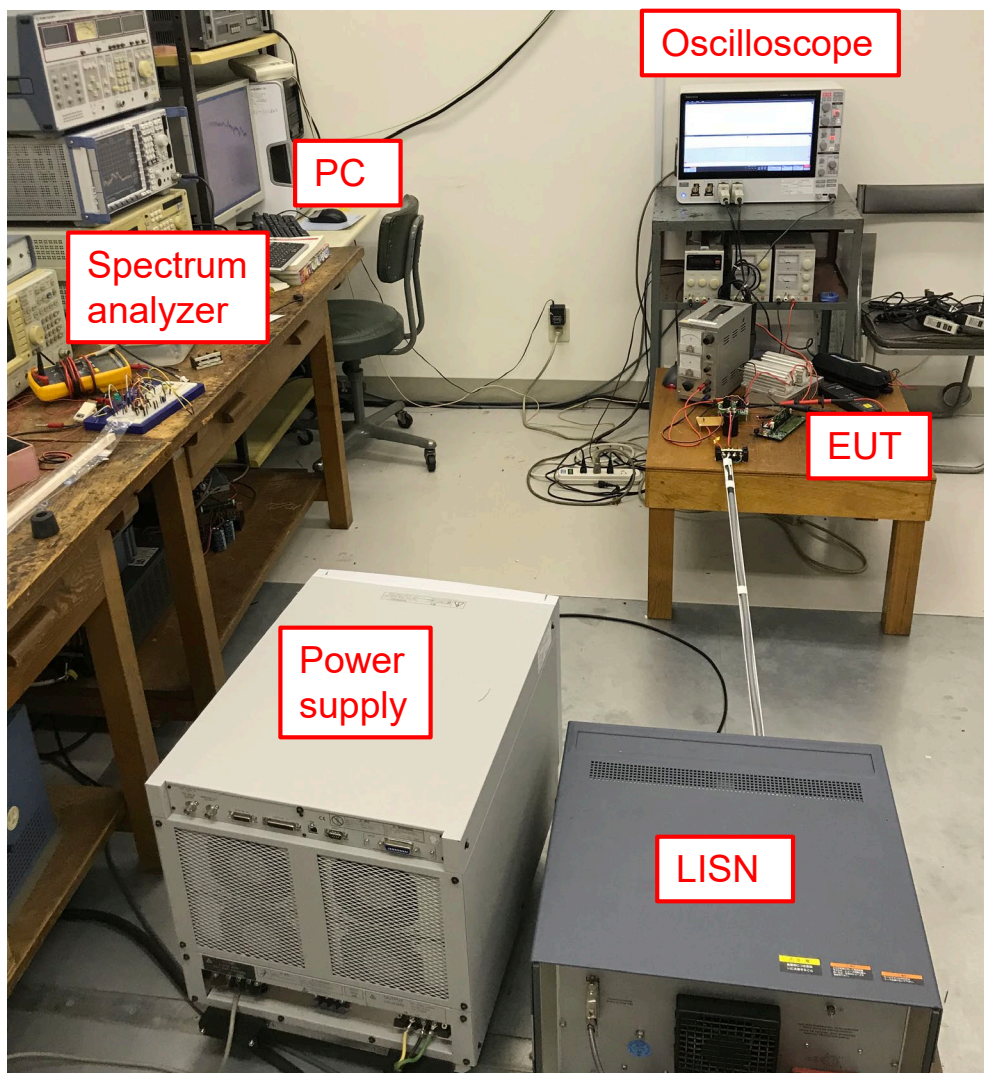


Fig.2.9. The experimental setup of conducted noise measurement.

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The experimental parameters of the boost converter are given in Table 2-1. Parameters of noise filters in the noise model applications are also included. The input voltage is 12 V. Therefore the output voltage should be 60 V due to the duty ratio of the switch. In experimental results, the actual output voltage is about 52 V because of the power loss. After calculation, the circuit efficiency is about to 86%.

TABLE 2-1 EXPERIMENTAL PARAMETERS OF THE BOOST CONVERTER

Parameter	Symbol	Value
Input voltage (V)	V_{IN}	12
Duty ratio		0.8
Switching Frequency (kHz)	f_{sw}	103
Power inductance (μH)	L	420
Load (Ω)	R_L	50
Capacitor to the ground (nF)	C_{CM}	1
Bypass Y-Capacitor (nF)	Z_{CY}	2
Common-Mode choke circuit (mH)	Z_{Choke}	2.36

The experimental waveform of I_{CM} is shown in Fig. 2.10. As the switching frequency is set at 103 kHz, the period is about 9.7 μs . The current peaks appear when the switch turns on and off. And the amplitude of turn off current is bigger than turn on current. This is partly caused by the asymmetric of the propagation route. For this reason, the selected duration should more than one period and the noise is expected to repeat itself. As the target frequency scope is from 150 kHz to 30 MHz. The resolution of the oscilloscope should also be adjusted and the corresponding record length ought to over 60 million

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points, due to the symmetry of the Fourier series.

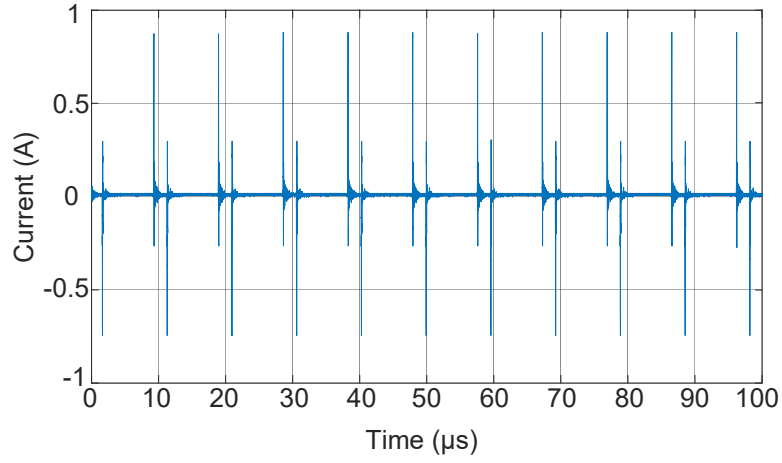


Fig.2.10 CM current waveform in the time domain.

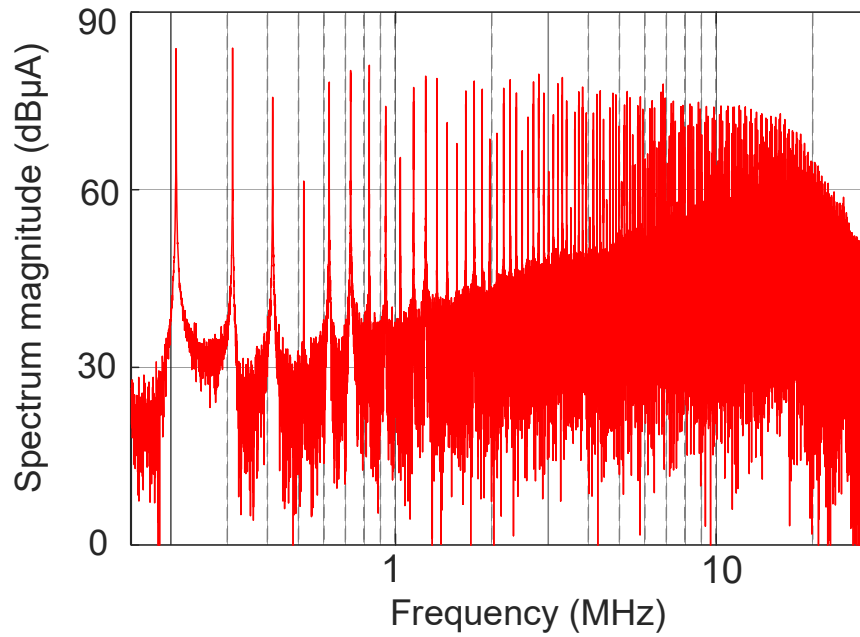


Fig.2.11 Spectrum of the CM current in the frequency domain.

As the noise model is built in frequency domain, the transition of I_{CM} from the time domain to frequency domain is necessary. The Fast Fourier Transform (FFT) calculation process could be realized by the oscilloscope or by the software. The spectrum of I_{CM} is

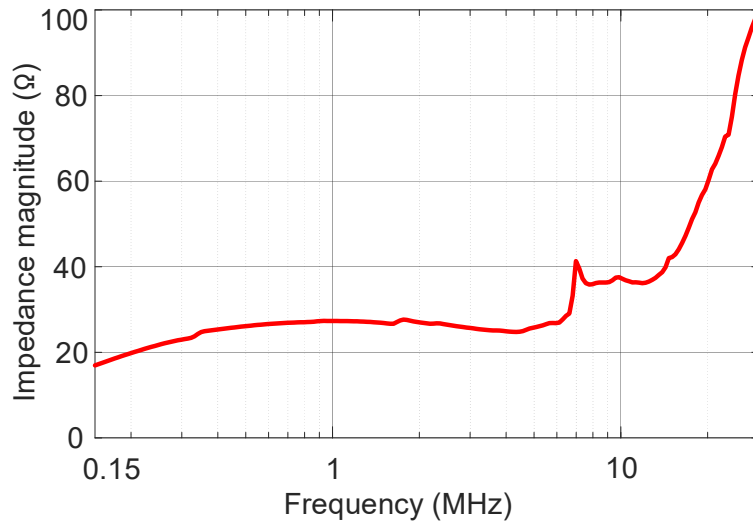
shown in Fig. 2.11.

Until now, the necessary elements are completed and the equivalent noise source V_N could be calculated with the following equation, in which all the variables are the complex number.

$$V_N = I_{CM}(Z_{Main} + Z_{Cable} + Z_{LISN}) \quad (2.1)$$

As the noise source in boost converter is the Mosfet, here we could validate the accuracy of V_N by comparing with the voltage between drain and source of the switch (V_{ds}). The experimental setup is in Fig.2.9.

As the real experimental conditions are not ideal, it is necessary to correct the experimental equipments. Before calculating the CM noise spectrum in frequency domain, this study also measured the CM impedance of LISN in reality. The measured impedance magnitude and phase are shown in Fig.2.12. Compared with calculating with ideal 25 Ω , the noise predictions based on real impedance value are more accurate.



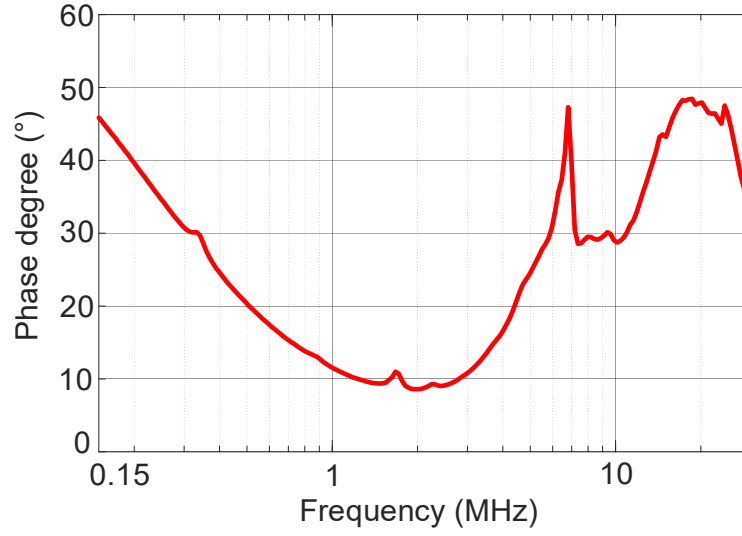


Fig.2.12 Impedance magnitude and phase degree of LISN from impedance analyzer.

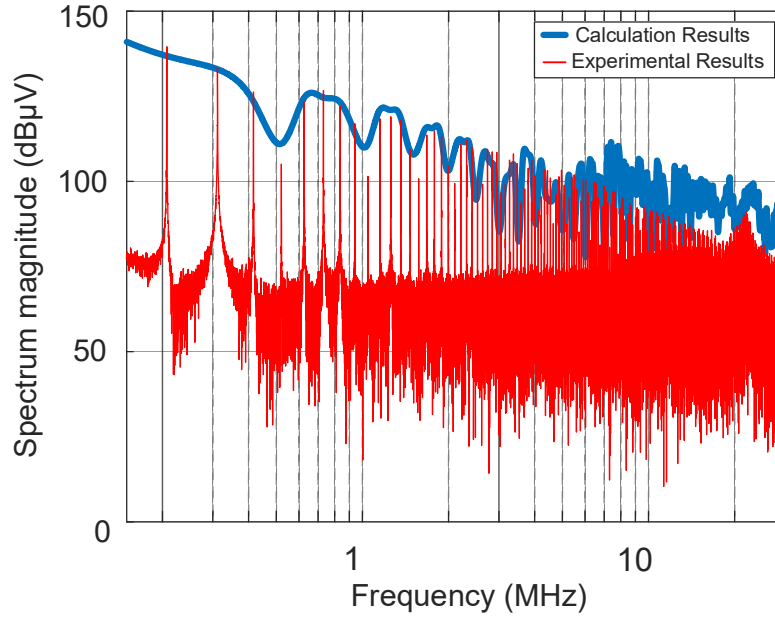


Fig.2.13 Noise source spectrum comparison.

The comparison between the calculation result and the experimental result is shown in Fig. 2.13. The red spectrum is the FFT results of the experimental measured V_{ds} . The peak at about 22 MHz is caused by the ringing voltage. In the figure, the blue envelope is the equivalent noise source V_N , as the calculation results of (2.1). These two curves can match

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well in most frequency range especially at low-frequency range. The deviation at high frequency may be caused by the accuracy limitation of the current probe.

This comparison does not only proves the effectiveness of the terminal port modeling theory, but also tests the accuracy of the measured current and the CM impedance. Therefore, this model is ready for predicting the effectiveness of the passive noise filters. However, one most important prerequisite of this modeling method is that the stability of the equivalent noise source, namely that the V_N would not fluctuate when operating conditions keep the same. This would be proven in the next section.

2.3 Passive Noise Filter Evaluation with the Noise Model

Passive noise filters are widely used in power converters, Fig.2.14 shows a typical noise filter system. Both the CM noise filter and DM noise filter are included. Based on the CM model proposed in the last section, this study adopts two kinds of CM filter to evaluate the accuracy and robustness of the model. This model would give an evaluation of the EMI filter according to the impedance value.

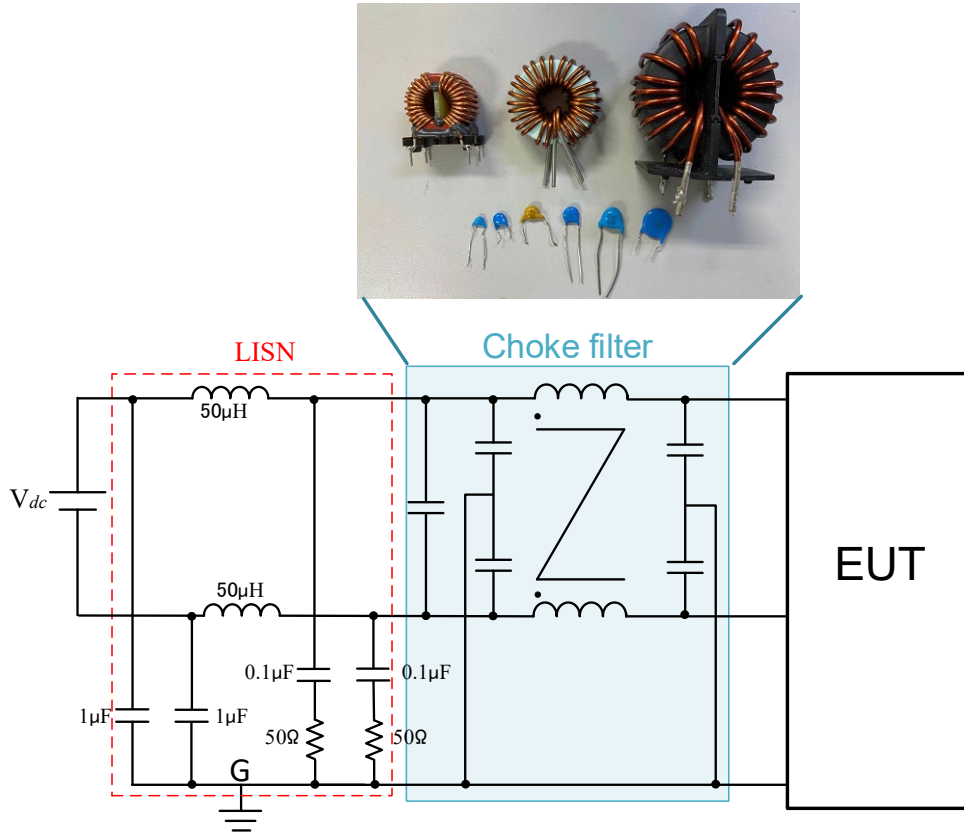
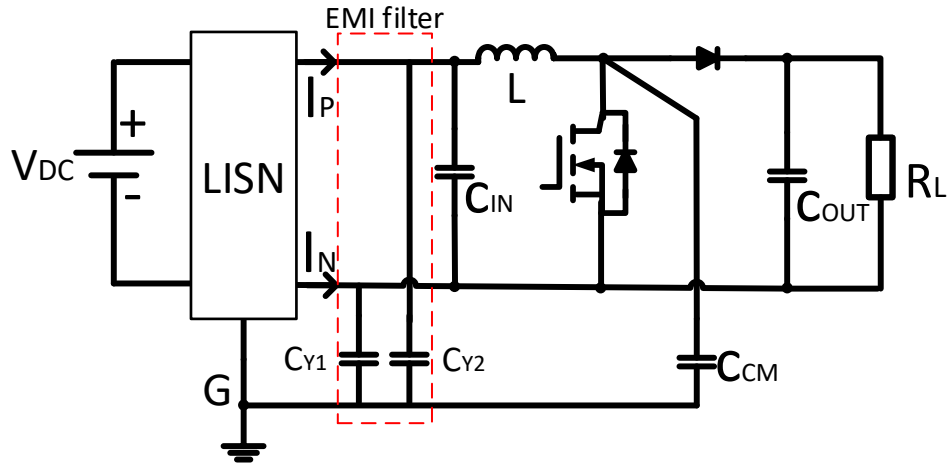


Fig.2.14 Typical CM noise filter.

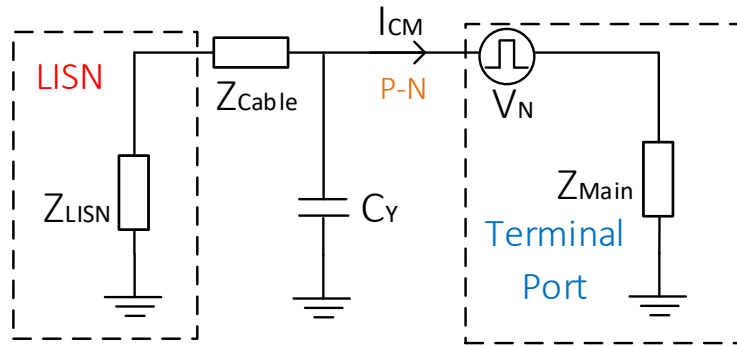
2.3.1 Bypass Y-capacitor EMI Filter

Class-Y capacitor, also known as the "the line bypass capacitor", namely the capacitor placed between line and ground. This capacitor can build a low impedance route between the line and ground. Thereby, the current flow through LISN would be decreased to some extent.

The configuration and equivalent CM model of boost converter after adding Y-capacitor is been shown in Fig.2.15. Here the positive line and negative line are connected to the bypass Y-Capacitor respectively to keep the impedance balance. Thus two bypass capacitors have been connected in parallel in this CM model. The amplitude and phase information of Y-capacitors is shown in Fig.2.16.



(a)



(b)

Fig.2.15 Boost converter after adding Y-capacitor EMI filter: (a). configuration. (b).CM model.

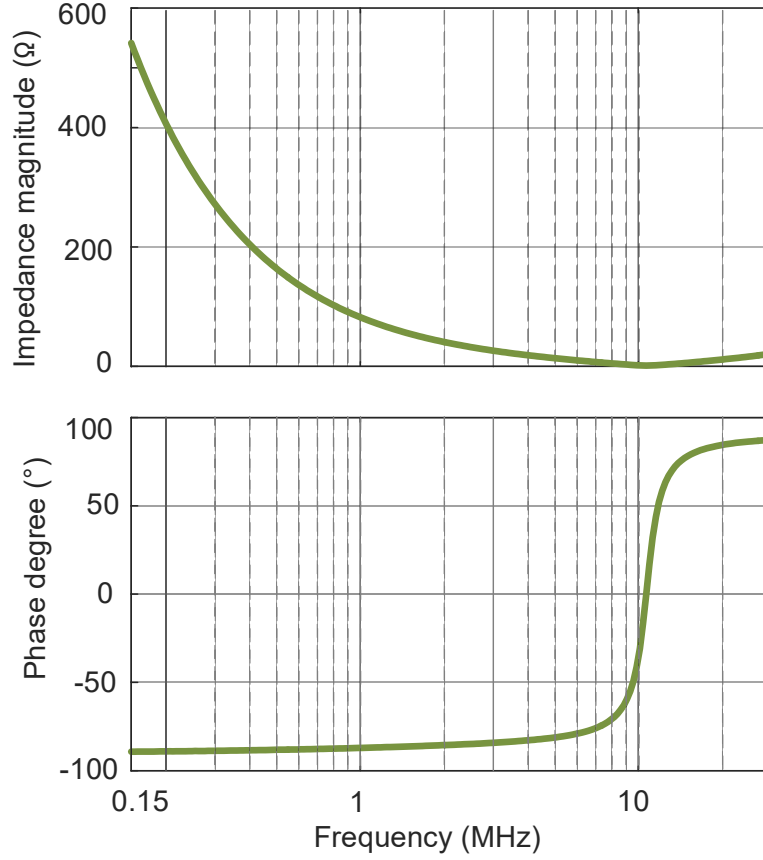


Fig.2.16 CM impedance magnitude and phase of C_Y .

According to the Kirchhoff voltage theorem, meanwhile referring to the equivalent CM model shown in Fig. 2.15, the voltage expression of LISN would be written as follows:

$$V_{CM} = Z_{LISN} \cdot \frac{V_N}{Z_{Main} + \frac{Z_{CY} \cdot (Z_{Cable} + Z_{LISN})}{Z_{CY} + Z_{Cable} + Z_{LISN}}} \cdot \frac{Z_{CY}}{Z_{CY} + Z_{Cable} + Z_{LISN}} \quad (2.2)$$

Note that Z_{LISN} is the impedance of LISN, and Z_{CY} is the impedance of the parallel Y-capacitor. The other coefficients stay the same with the former section. With (2.2), the impact after adding Y-Capacitor could be predicted.

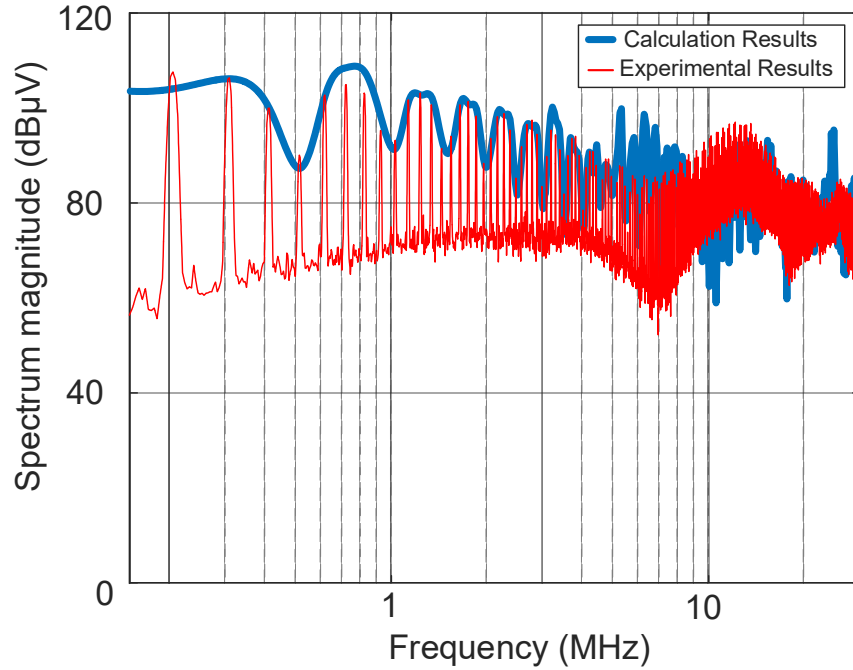


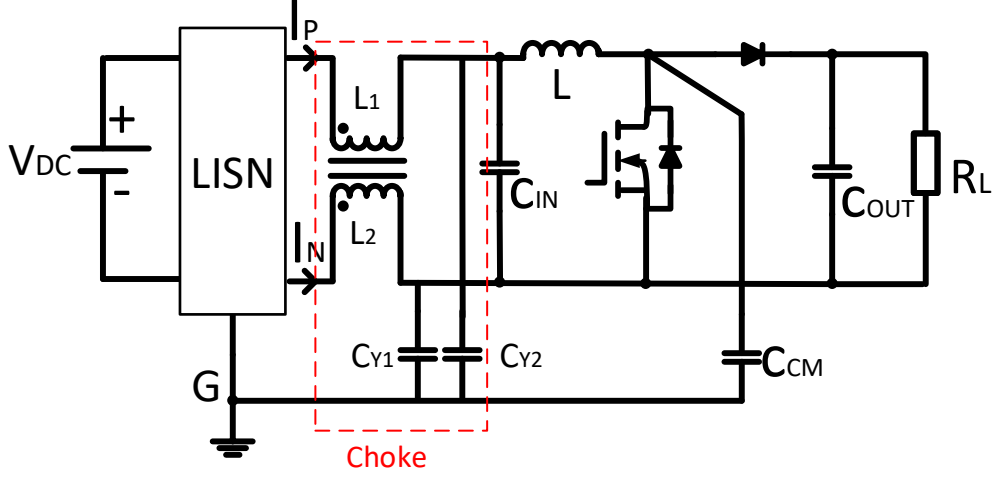
Fig.2.17 Comparison between experimental result and prediction results.

The comparison between experimental results and prediction results have shown in Fig. 2.17. The red spectrum is the experimental results of LISN, while the blue envelope is the prediction results. These two curves could match well in most frequency, and the peaks and valleys appear at the same frequency. These prediction results could represent the characters of the bypass Y-capacitor well.

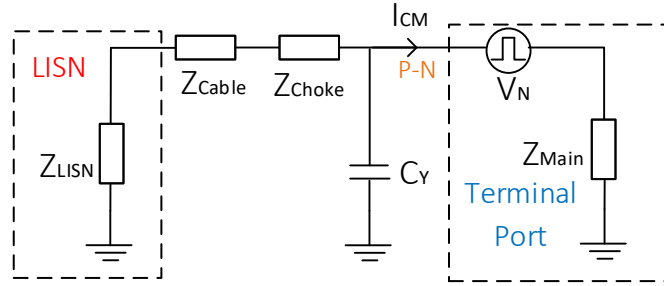
2.3.2 CM Choke Filter Circuit

In order to further suppress the CM current, a CM choke circuit coordinated with Y-capacitor is applied in this study. The choke circuit allows current to flow back in these coils, thus canceling the magnetic field in the core and may suppress CM current effectively. In the past, the effectiveness of a choke could only be verified by experiments. Whereas the model proposed in this paper offers a better way to predict the suppression effect with no need to practice manufacture. The configuration and equivalent CM model

of boost converter after adding the choke and Y-capacitor circuit is shown in Fig.2.18.



(a)



(b)

Fig.2.18 Boost converter adding common-choke EMI filter: (a). configuration. (b).CM noise model.

Similarly, according to the CM equivalent model in Fig.2.18, the voltage of LISN could be expressed in (2.3):

$$V_{CM} = Z_{LISN} \cdot \frac{V_N}{Z_{Main} + \frac{Z_{CY} \cdot (Z_{Cable} + Z_{LISN} + Z_{Choke})}{Z_{CY} + Z_{Cable} + Z_{LISN} + Z_{Choke}}} \cdot \frac{Z_{CY}}{Z_{CY} + Z_{Cable} + Z_{LISN} + Z_{Choke}} \quad (2.3)$$

Where Z_{Choke} is the impedance of the choke circuit. The amplitude and phase information of the choke is shown in Fig. 2.19. With the necessary impedance information,

CM Noise Modeling Method with Two-port Terminal Theory

the evaluation of this CM choke could be done by substituting parameters into (2.3). All the calculation process is done in the frequency domain.

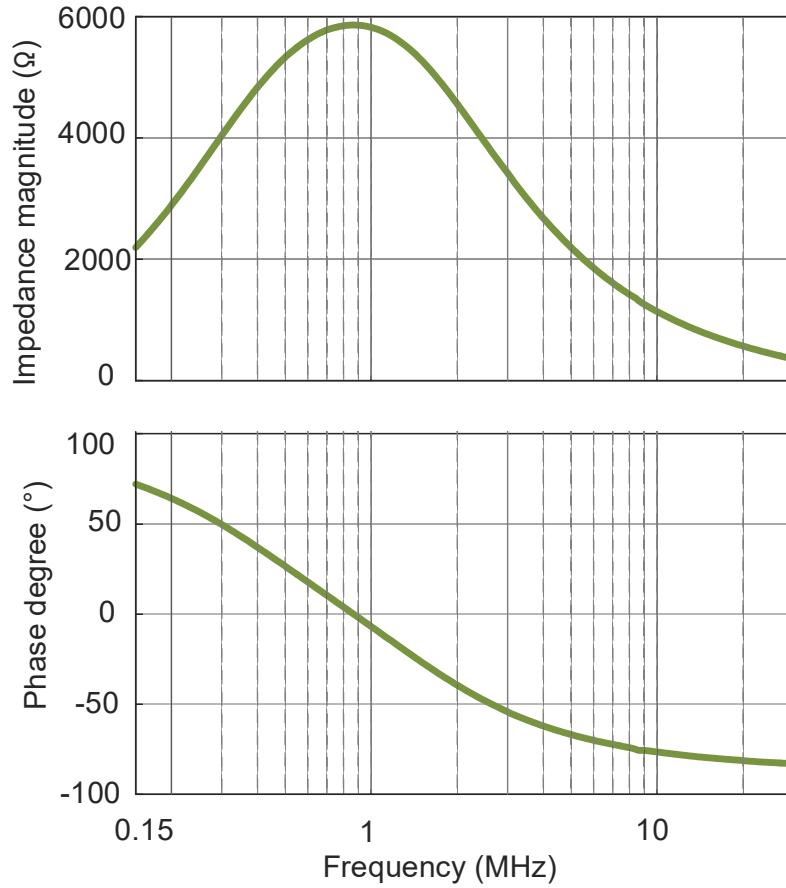


Fig.2.19 CM impedance magnitude and phase of Z_{Choke} .

The comparison between experimental results and prediction results has been shown in Fig.2.20, in which the deviations between the two curves are within a reasonable range. Also, compared to the experimental results in Fig.2.19 and Fig.2.20, the choke circuit can attenuate the CM noise better than the Y-capacitor.

The noise model application in EMI filter evaluation, namely the prediction of the filter can illustrate the effectiveness of this modeling method. What's more, the above two cases would also verify the viewpoints that the noise source would keep constant when

the operation condition stays the same. The passive filter circuit would only affect the propagation route impedance.

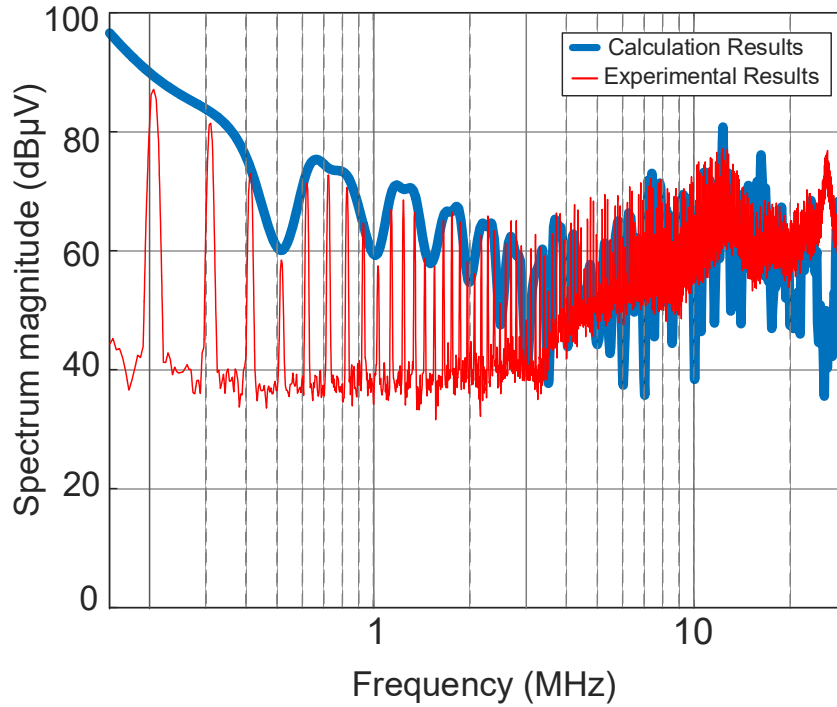


Fig.2.20 Comparison between experimental result and prediction results.

2.4 CM Noise Model of the Single-Phase Inverter

This study also applied this terminal port theory modeling method to a single-phase inverter. The modeling process is similar with the boost converter. The equivalent noise voltage source is extracted following the Thevenin Theorem. Here T-type structure is adopted for extending the available range of this model, especially the prediction after adding filter at input side and output side at the same time. The filter interaction effect is investigated here to optimize the filter design and selection from the view of overall system. At last, comparisons between the measured and predicted results show that this EMI modeling method can provide adequate prediction even the system is complex.

2.4.1 CM Noise Analysis in Single-Phase Inverter

The detailed circuit of a single-phase inverter connecting with LISN is shown in Fig. 2.21. V_{dc} is the DC source. P, N, and G represent phase, neutral, and ground respectively. C_{AG} and C_{BG} are the capacitors connecting with the ground and the middle point of each phase-leg, which represent the stray capacitors between the heatsink and the frame ground. The current flowing through the phase line and the neutral line is I_P and I_N respectively.

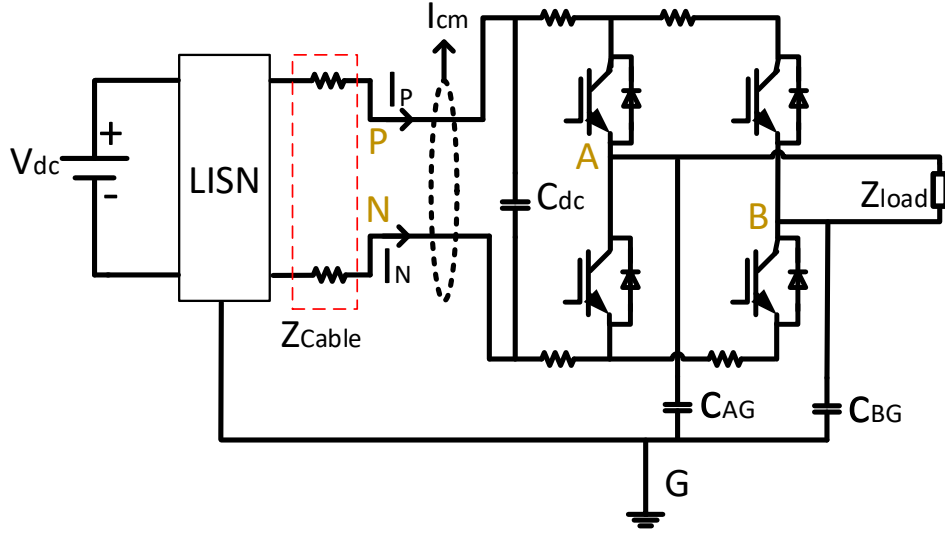


Fig.2.21 Configuration of single-phase inverter connected with LISN.

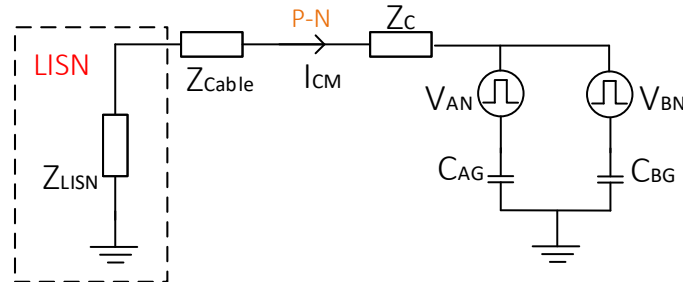


Fig.2.22 CM equivalent circuit according to substitution theorem.

According to the definition of (1.5) – (1.6), the CM current could be represented by I_P and I_N , namely the line current. As is shown in Fig. 2.22, due to the substitution theorem,

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the circuit behavior would stay the same if replace the phase-leg by a high-frequency time-varying voltage source, namely V_{AN} and V_{BN} . Therefore the CM voltage can be determined by the following equation:

$$V'_N = (V_{AN} + V_{BN})/2 \quad (2.4)$$

Here V_{AN} and V_{BN} , namely the voltage between drain and source of the lower switch in each bridge. In Fig.2.22, Z_C is the impedance of inverter circuit, which can be ignored when measuring the V_{AN} and V_{BN} .

Moreover, the CM noise model could be also obtained by the terminal port theory. The structure of the model is shown in Fig. 2.23. The CM impedance of inverter circuit Z_{SW} , ground capacitor C_{SW} and the noise source V_N can be seen connected in series in the terminal port box. Z_{Cable} is the cable impedance between the EUT and LISN, which is the same as in the boost converter model.

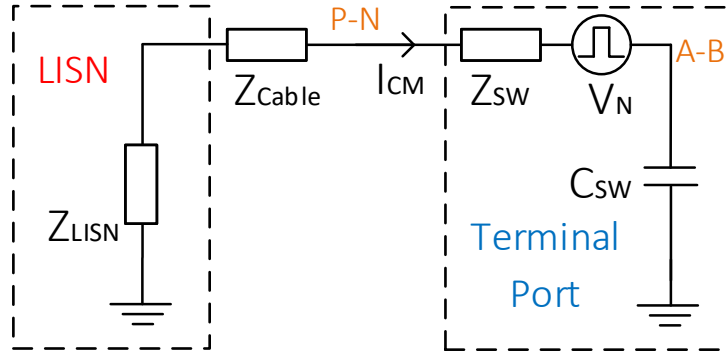


Fig. 2.23 CM model of single-phase inverter.

Different inner structures have been adopted in the former study. Like two noise sources that aim to separate the DM noise and CM noise [16]. Or the two noise sources in series with impedance matrix structure in [23], which can predict the input side and output side insertion loss at one time, but the relationships between two noise sources need further

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research. This study optimized the terminal port structure and adopted one equivalent noise source in series with CM impedance, which is simple but effective.

The other necessary step is the identification of impedance. The impedance definition process is similar to the former boost converter. Generally speaking, the steps of CM modeling with terminal port theory could be summarized as follows:

- 1) Investigate the propagation of noise current, investigate the inner structure of the terminal port
- 2) Measure the off-line CM impedance of each segment with the impedance analyzer. The amplitude and phase information should be recorded and fit for the interpolation algorithm.
- 3) Connect with the power supply and operate the inverter circuit. Then measure the CM current I_{CM} . Here the current period should more than one line period. The appropriate value may also consider the sampling resolution of the oscilloscope.
- 4) The equivalent CM noise source could be calculated with the impedance and current value. Finally, the modeling process is completed.

This study measured the off-line CM impedance of each segment with the impedance analyzer and process the data with the proposed interpolation algorithm.

2.4.2 Noise Source Extraction of the Model

As was mentioned, comparing with the conventional CM noise modeling technique, the biggest difference in this method is the noise source. Noted that V_N in Fig. 2.23 is not the paralleled calculation results of V_{AN} and V_{BN} in Fig. 2.22. V_{AN} and V_{BN} represent the

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changing electric potential to the ground, which can be measured by voltage probe directly. Oppositely, V_N is the calculation value indirectly. Here definition method is given, just to compare with the terminal port method and give the reference.

CM current I_{CM} is required to build the model. After dealing with time-domain current data with Fourier transformation, the noise source in frequency-domain would be calculated by the following formula.

$$V_N = I_{CM}(Z_{SW} + Z_{Csw} + Z_{Cable} + Z_{LISN}) \quad (2.5)$$

CM impedance has been measured by impedance analyzer Agilent 4294A. The main circuit impedance is measured when all switches are shorted (drain and source were short connected). The CM noise sources comparison is shown in Fig. 2.24.

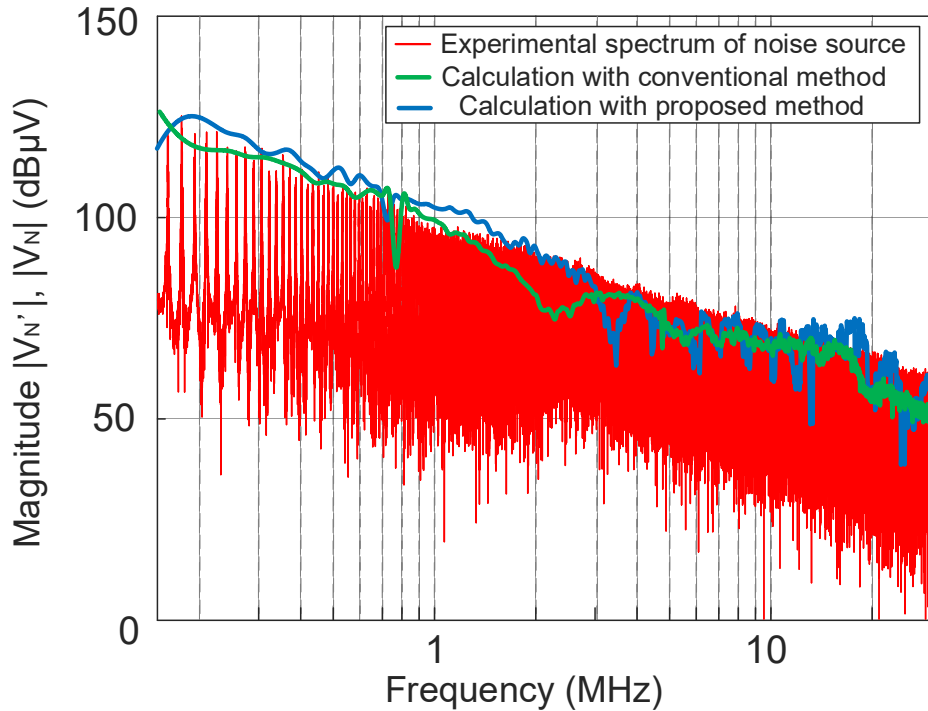


Fig. 2.24 Noise source spectrum comparison of single-phase inverter.

Here the red spectrum is the experimental result V_N' , namely the consequence of

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(2.4). And the blue curve is the equivalent noise source V_N calculated by (2.5), where the data is from the proposed algorithm. The green envelope is from the conventional impedance measurement method. After comparison, the envelope with proposed method can match well in the frequency range. This can not only prove the accuracy of the equivalent noise source but also show the CM current and impedance have been measured exactly.

The key experimental parameters of single-phase inverter and the filters in the following study are given in Table.2-2.

TABLE 2-2 EXPERIMENTAL PARAMETERS OF THE SINGLE PHASE INVERTER CONVERTER

Parameter	Symbol	Value
Input voltage (V)	V_{dc}	50
Input side capacitor (μF)	C_{dc}	330
Load (Ω)	Z_{load}	50
Switching frequency (kHz)	f_{sw}	16
Output voltage frequency (Hz)	f_{Vout}	50
L-type harmonic filter (mH)	L_{f1}, L_{f2}	0.8
Y-capacitor (nF)	C_{Y1}, C_{Y2} C_{Y3}, C_{Y4}	1

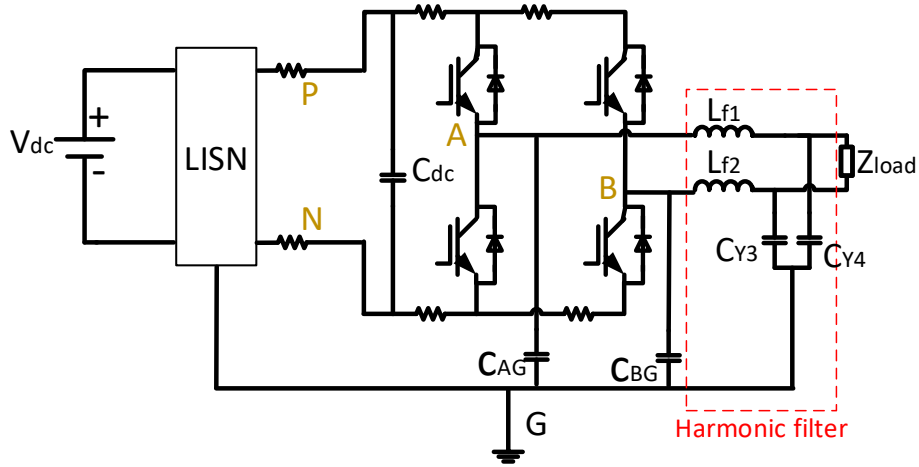
2.5 The Noise Filter and Harmonic Filter Design with the Proposed Model

Based on the CM noise model established in the above section, three different situations have been set to evaluate this model. Different from the traditional terminal port modeling methods which could only predict the impact of input side filter, or the optimized methods which could reflect the impedance changing at the input side and output side, but need two noise sources and complex impedance measurement. The modeling method proposed in this study determined the structure in the terminal port, this optimization expanded the range of application while ensuring accuracy.

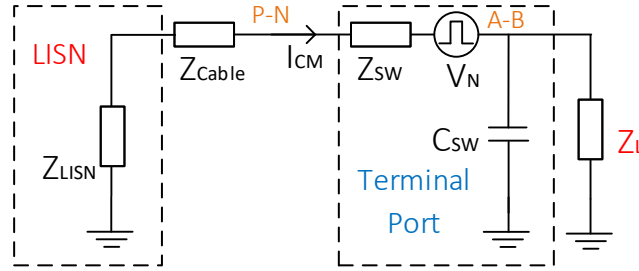
The inverter in different cases has been set working at the same conditions. The input side voltage (V_{dc}) is 100V, the switching frequency is 16 kHz, and control signals have been generated by the F28069 Piccolo control card.

2.5.1 Model Validation after Adding Inductance Harmonic Filter

Harmonic filters at the output side are commonly used in some situations when power quality is required, such as inverter working at grid-connected conditions. This study utilizes the L-type filter and mainly investigates its CM conductive noise impact. The configuration and the equivalent CM noise model are shown in Fig.2.25.



(a)



(b)

Fig.2.25 Single-phase inverter after adding inductance harmonic filter,
(a). Configuration (b). CM equivalent model.

In order to achieve a better filtering effect, two same inductances have been in series with the A and B phases respectively. And the inductances are connecting with the frame ground through a 1nF capacitor, which simulates the stray capacitance.

Fig 2.26 gives the inductance filter CM impedance magnitude and phase information. Because of the ground capacitor, the inductance turns to capacitive at around 600 kHz. After the turning point, the magnitude becomes smaller with the increase of frequency.

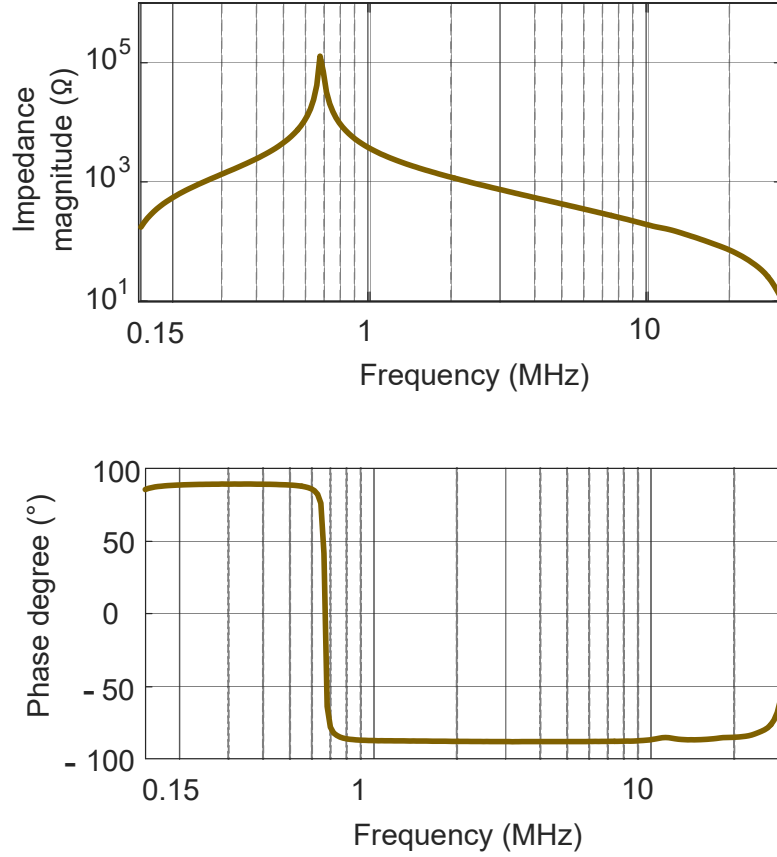


Fig.2.26 CM impedance magnitude and phase of L-type harmonic filter.

According to the proposed noise model, the voltage across the LISN can be expressed in (2.6).

$$V_{CM} = Z_{LISN} \cdot \frac{V_N}{Z_{LISN} + Z_{Cable} + Z_{SW} + \frac{C_{SW} \cdot Z_L}{C_{SW} + Z_L}} \quad (2.6)$$

$$\text{Where: } Z_L = (j\omega L_{f1} + \frac{1}{j\omega C_{Y4}}) // (j\omega L_{f2} + \frac{1}{j\omega C_{Y3}}) \quad (2.7)$$

Here Z_{LISN} is the impedance of LISN, Z_L is the impedance of the L-type filter and V_N is the equivalent noise source extracted in the last section. Then the spectrum could be plotted after transferring V_{LISN} from time-domain to frequency-domain. Set the frequency range from 150 kHz to 30 MHz, and the calculation result is shown as the blue curve in Fig.2.27. Compared with the experimental result represented by the red curve. The two

envelopes can match well in most frequency.

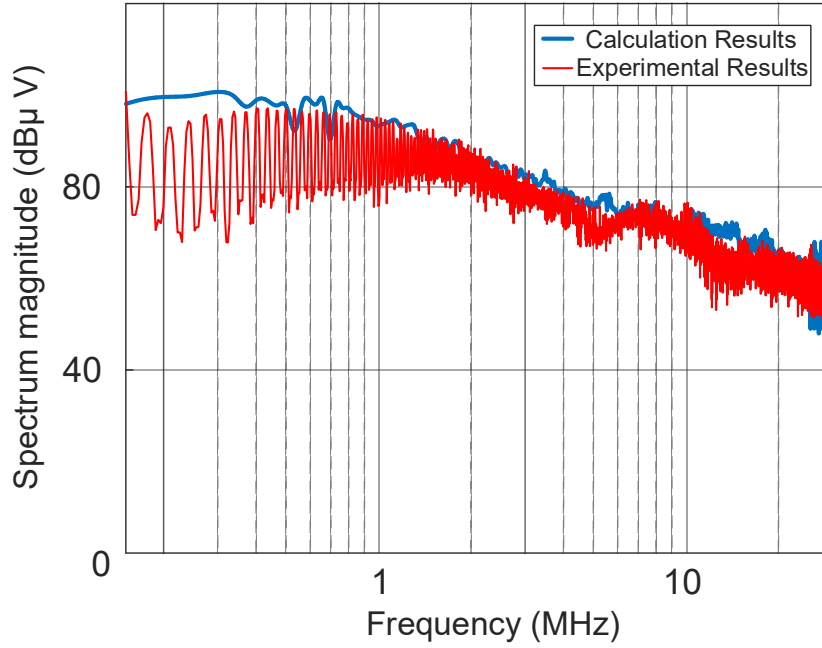
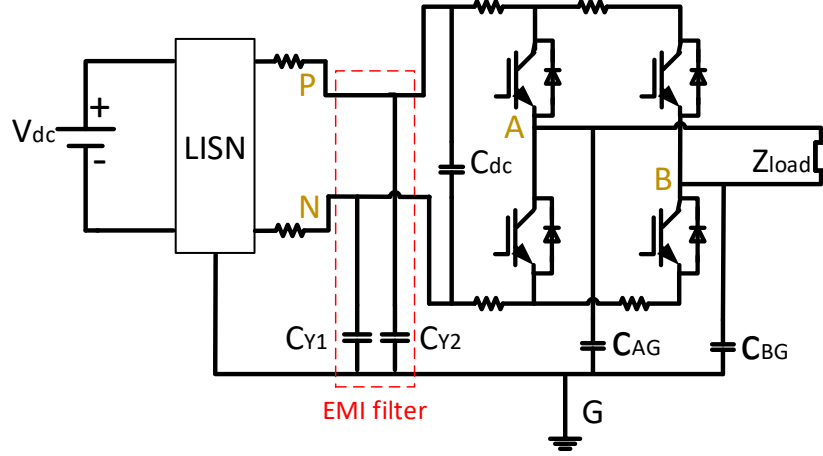


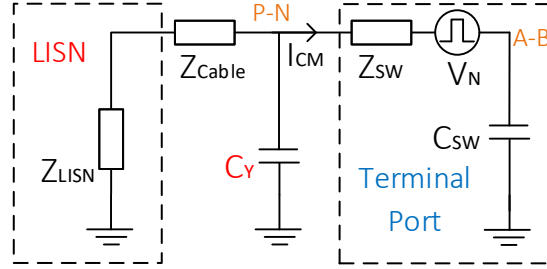
Fig.2.27 Comparison between experimental result and prediction results.

2.5.2 Model Validation after Adding Bypass Y-capacitor EMI Filter

In this section, two Y-capacitors are placed between the power line and the ground (P-G and N-G). The value of Y-capacitor cannot be too large due to the limit of leakage current, generally range from a few hundred Pico farad to a few nano farads. Similar with the boost converter, 1 nF high voltage ceramic capacitor is adopted at here. The CM model after adding EMI filter is shown in Fig.2.28.



(a)



(b)

Fig.2.28 Single-phase inverter after adding bypass Y-capacitor EMI filter:

(a). configuration; (b). CM model.

According to the Kirchhoff voltage theorem, meanwhile comparing the equivalent CM circuit before and after adding the EMI filter, the voltage expression of LISN would be written as follows:

$$V_{CM} = Z_{LISN} \frac{V_N}{Z_{SW} + Z_{C_{SW}} + \frac{Z_{CY}(Z_{Cable} + Z_{LISN})}{Z_{CY} + Z_{Cable} + Z_{LISN}}} \frac{Z_{CY}}{Z_{CY} + Z_{Cable} + Z_{LISN}} \quad (2.8)$$

$$\text{Where:} \quad C_Y = C_{Y1} + C_{Y2} \quad (2.9)$$

Here Z_Y is the impedance of the Y-capacitor. The impact after adding Y-Capacitor could be predicted precisely.

CM Noise Modeling Method with Two-port Terminal Theory

Fig 2.29 shows the CM impedance magnitude and phase versus frequency curve of Y-capacitor. The turning frequency point happens around 10 MHz. The comparison of experimental results and prediction is shown in Fig.2.30. A few dBs attenuations have been achieved compared with Fig.2.27.

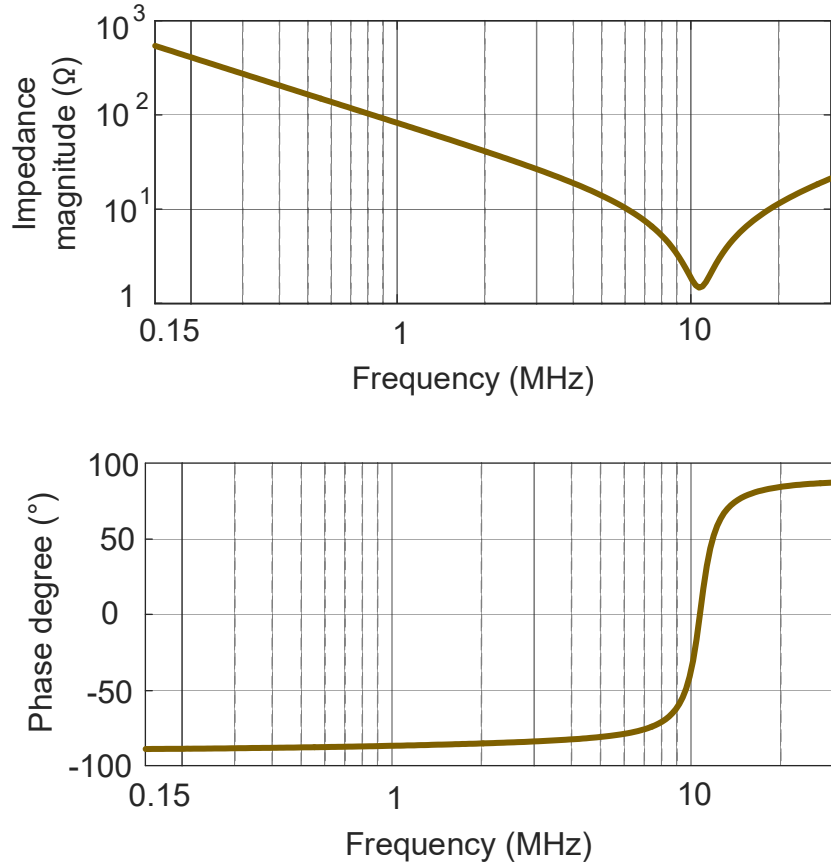


Fig.2.29 CM impedance magnitude and phase of Y-capacitor.

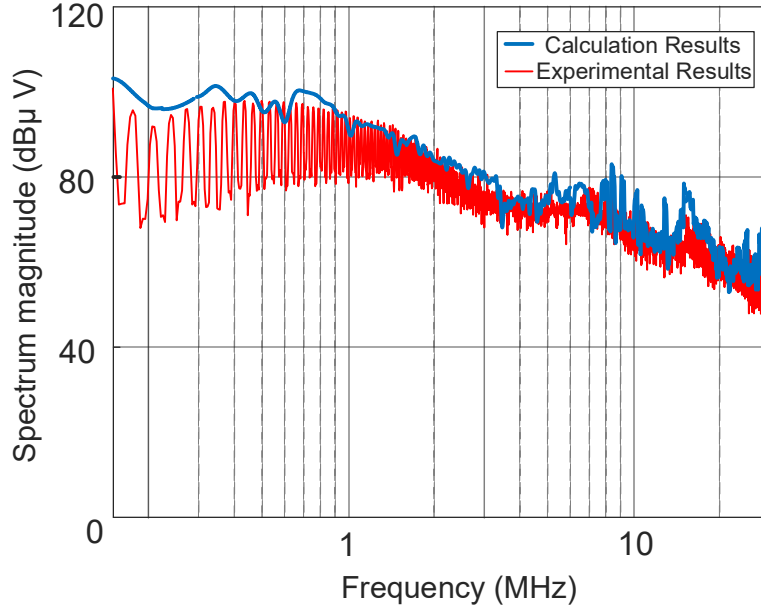


Fig.2.30 Comparison between experimental result and prediction results.

2.5.3 Model Validation after Adding Bypass Y-capacitor EMI Filter and L-type Harmonic Filter

The circuit configuration and CM equivalent model after adding the Y-capacitor EMI filter and L-type harmonic filter are shown in Fig.2.31. Here the advantages of T-type structure in the terminal port could be seen clearly. Not only the input side or output side filter would be evaluated separately, but also the combination system could be predicted easily and precisely.

Similarly, according to the CM equivalent model shown in Fig.2.32, the voltage of LISN could be expressed in (2.10):

$$V_{CM} = Z_{LISN} \frac{V_N}{Z_{SW} + \frac{Z_{CSW} Z_L}{Z_{CSW} + Z_L} + \frac{Z_{CY}(Z_{Cable} + Z_{LISN})}{Z_{CY} + Z_{Cable} + Z_{LISN}}} \frac{Z_{CY}}{Z_{CY} + Z_{Cable} + Z_{LISN}} \quad (2.10)$$

Where: $C_Y = C_{Y1} + C_{Y2}$ (2.11)

$$Z_L = (j\omega L_{f1} + \frac{1}{j\omega C_{Y4}}) // (j\omega L_{f2} + \frac{1}{j\omega C_{Y3}})$$
 (2.12)

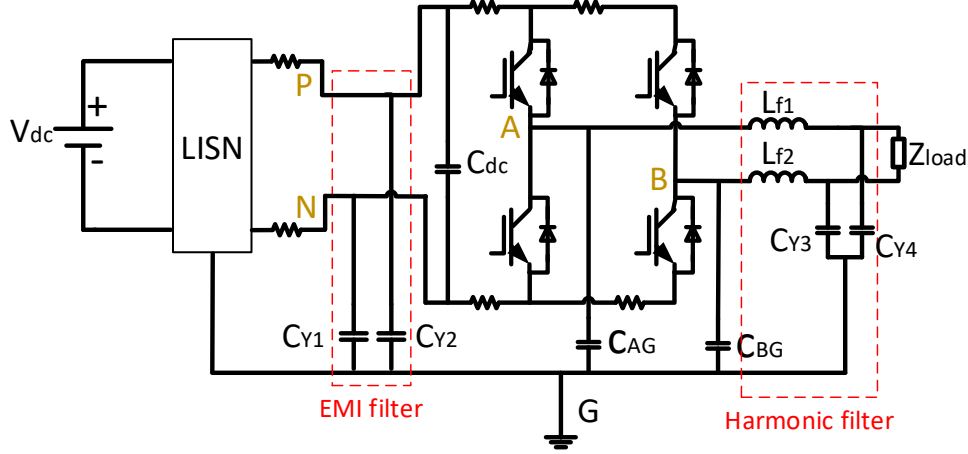


Fig.2.31 Configuration after adding bypass Y-capacitor filter and L-type filter.

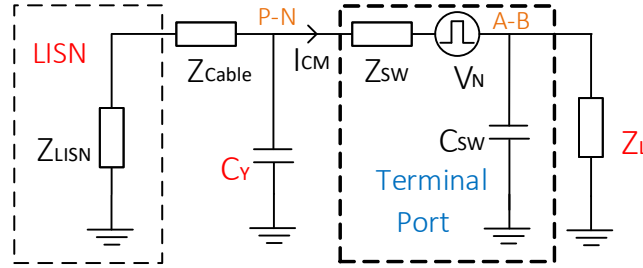


Fig.2.32 CM equivalent Model.

Then the comparison between experimental result and prediction result is shown in Fig.2.33. Compared with the former filter results, high-frequency oscillation occurs because of the insertion of the harmonic filter. However, the oscillation trends of the prediction and experimental results are the same, and the deviations between the two curves are within a reasonable range. The comparison could illustrate this model can predict the interaction impact of EMI filter and harmonic filter effectively.

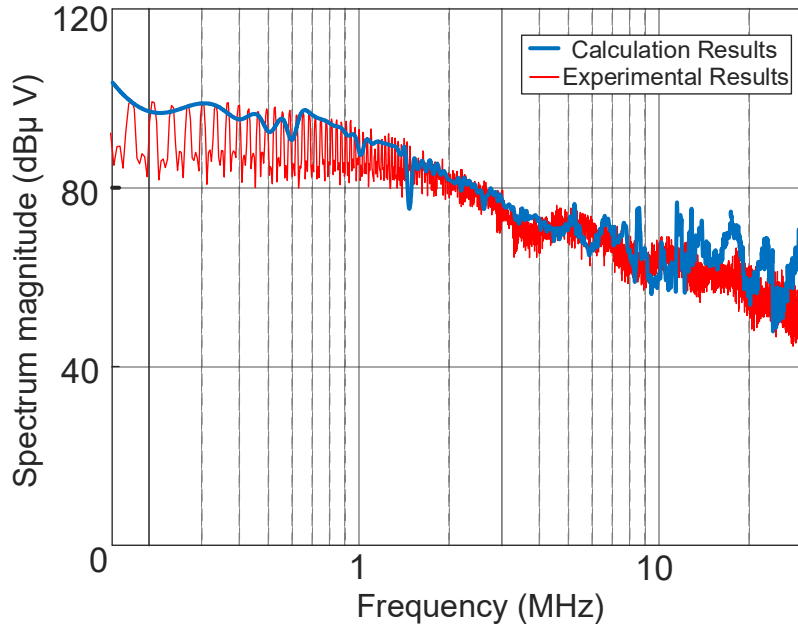


Fig.2.33 Comparison between experimental results and prediction results.

2.6 Conclusions

In this chapter, an improved behavioral modeling method based on the terminal port theory is proposed. This modeling method is utilized in the DC-DC boost converter and DC-AC single-phase inverter to evaluate and optimize the noise filter design. The main features of the proposed modeling method are as follows:

- 1) The internal structure of the terminal port box is investigated. Compared to the former behavioral modeling methods which treated the converter as a “black box”, this proposed structure could help to identify the noise source and CM impedance.
- 2) The accuracy of the impedance is improved with the interpolation algorithm. As the modeling process is processed in the frequency domain, the magnitude and the

phase angle of the impedance are the most significant information. However, the previous behavioral models were depended on the physical model from the impedance analyzer. The proposed interpolation algorithm could help to modify the measurement results for the frequency domain calculation directly.

- 3) Not only the EMI filter at the input side but also the harmonic filter at the output side are evaluated with the proposed model. As the additional harmonic filters may change the route of the noise current, and further affect its performance of the conducted noise. The comparison between experimental results and the prediction results could verify that different kinds of filters could be evaluated with the proposed noise model.

Chapter 3 Optimized CM Noise Analytical Modeling

Method

3.1 Fourier Series and Noise Spectra Analysis

As was mentioned in the first chapter, another major branch of the frequency domain modeling method is the analytical modeling method. Generally speaking, this method needs to investigate the noise propagation and analyze the current or voltage waveform directly.

The analytical modeling method required a certain knowledge of mathematical and power electronics. In the meantime, this analysis process helps engineers to have a deeper understanding of the generation mechanism of EMI noise.

In the study of EMI issues in power converters, the periodic voltage fluctuations are the reasons for the conducted and radiated noise. The high dv/dt and di/dt are caused by the behaviors of the semiconductor devices in the circuit. Therefore, the waveform of the semiconductors is selected as the object for mathematical analysis.

Fourier transformation gives us a useful tool to estimate the time-domain signal in frequency domain. The periodic function could be represented by a sum of trigonometric functions. The voltage and current waveforms in the power converter which satisfied the definitions were taken as continuous-time (CT) signals [38]. The time-domain description of CT can be expressed as $x(t)$, a function of time t . After Fourier transformation, the function $S(f)$, which depicts amplitude versus frequency, could reveal the component sinusoidal frequencies and their amplitudes. The data processing is shown in Fig.3.1.

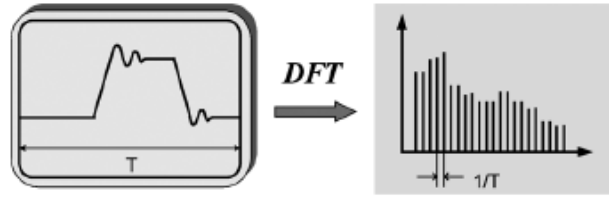


Fig.3.1 Illustration of data processing.

Based on the characteristics of the waveform, the voltage across the switch would be taken as a rectangle or trapezoid pulse train. Therefore in this section, the Fourier transform theorem will be introduced first, and then rectangle and trapezoid will be given as examples to show the Fourier series calculation process.

3.1.1 Introduction to Fourier Transform Theorem

In 1804, French mathematician Jean-Baptiste Joseph Fourier firstly proposed that "any function of a variable, whether continuous or discontinuous, can be expanded in a series of sines of multiples of the variable", but did not give a strict proof. Then in 1829, the French mathematician Dirichlet proposed the Dirichlet's theorem, which laid a theoretical foundation for the Fourier series [39].

According to the theorem, if a periodic signal $x(t)$ could satisfy the Dirichlet conditions, then it could be expressed with the complex exponential form as in (3.1). where the T is period and the ω is the angular frequency of the signal $x(t)$.

$$x(t) = \sum_{n=-\infty}^{\infty} C_n e^{jn\omega t} \quad (3.1)$$

Where $\omega = \frac{2\pi}{T}$, Then the Fourier coefficient C_n would be calculated with (3.2) on the basis of orthogonality of complex exponentials.

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$$C_n = \frac{1}{T} \int_{t_0}^{T+t_0} x(t) e^{-jn\omega t} dt = \frac{1}{T} \int_{\langle T \rangle} x(t) e^{-jn\omega t} dt \quad (3.2)$$

The detailed mathematical analysis and strict proofs would be referred to as the relevant textbooks [39]-[41]. This article focus on the application of the theorem, examples in engineering applications will be introduced below.

3.1.2 Fourier Series Calculation of Rectangular Waveform

A typical periodic rectangular waveform is shown in Fig.3.2, and the signal in one period. The information of the period T and pulse width τ are contained in the periodic square waveform.

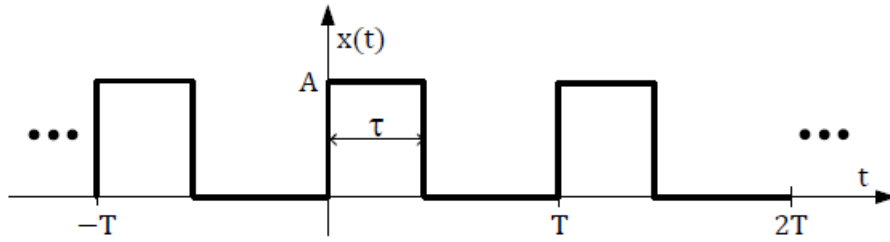


Fig.3.2 Periodic rectangular waveform.

The function in the time domain could be expressed by (3.3).

$$x(t) = \begin{cases} A & 0 < t \leq \tau \\ 0 & \tau < t \leq T \end{cases} \quad (3.3)$$

According to the definition in (3.2), the Fourier coefficient of the rectangular waveform could be expressed by the following equation:

$$C_n = \frac{1}{T} \int_{\langle T \rangle} x(t) e^{-jn\omega t} dt = \int_0^\tau A e^{-jn\omega t} dt = \frac{A}{jn\omega T} (1 - e^{-jn\omega \tau}) \quad (3.4)$$

These coefficients can be expressed in terms of a commonly used function known as the *sinc* function.

$$\begin{aligned}
 C_n &= \frac{A}{jn\omega_0 T_0} [1 - e^{-jn\omega_0 \tau}] = \frac{Ae^{-\frac{jn\omega_0 \tau}{2}}}{jn\omega_0 T_0} \left[e^{\frac{jn\omega_0 \tau}{2}} - e^{-\frac{jn\omega_0 \tau}{2}} \right] \\
 &= \frac{Ae^{-\frac{jn\omega_0 \tau}{2}}}{jn\omega_0 T_0} 2j \sin\left(\frac{n\omega_0 \tau}{2}\right) = \frac{A\tau}{T_0} e^{-\frac{jn\omega_0 \tau}{2}} \frac{\sin\left(\frac{n\omega_0 \tau}{2}\right)}{\frac{n\omega_0 \tau}{2}} = \frac{A\tau}{T_0} e^{-jn\pi \frac{\tau}{T_0}} \frac{\sin\left(n\pi \frac{\tau}{T_0}\right)}{n\pi \frac{\tau}{T_0}} \\
 &= \frac{A\tau}{T_0} e^{-jn\pi \frac{\tau}{T_0}} \text{sinc}\left(n \frac{\tau}{T_0}\right)
 \end{aligned} \tag{3.5}$$

where the *sinc* function is defined by

$$\text{sinc}(x) = \frac{\sin(\pi x)}{\pi x} \tag{3.6}$$

The expression of *sinc* could help us to estimate the breaking point in the frequency domain spectra. The spectrum of $\text{sinc}(x)$ is given in Fig.3.3. As the value of $\sin(x)$ is 0.84 when $x=1$, its corresponding spectrum on the double logarithmic axes, The slope would become -20dB/dec when x becomes bigger than 1 [41].

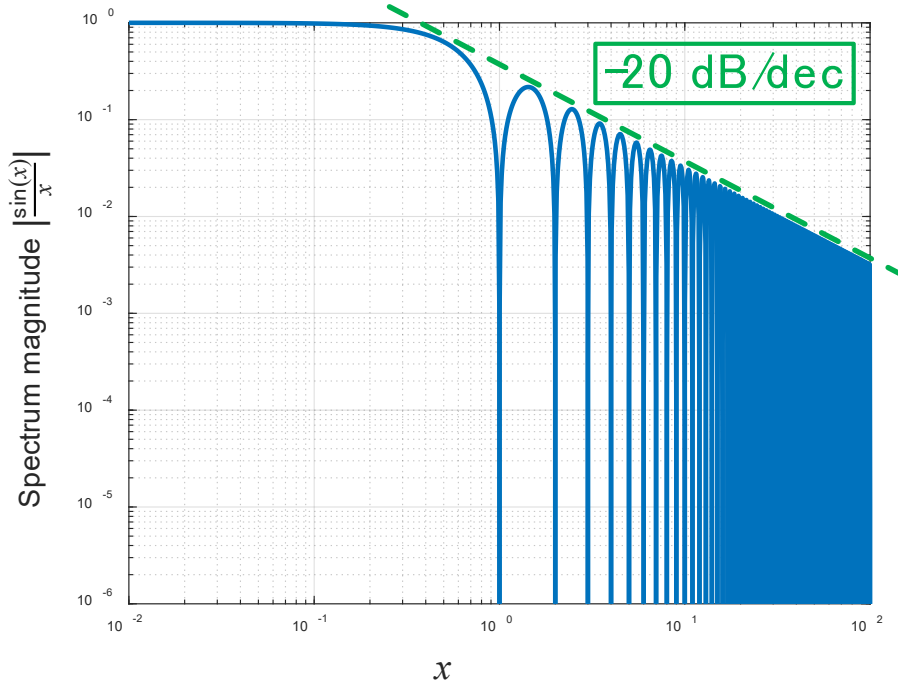


Fig.3.3 The spectrum of $\text{sinc}(x)$.

Accordingly, the spectral envelope of a periodic rectangular waveform is shown in Fig.3.4.

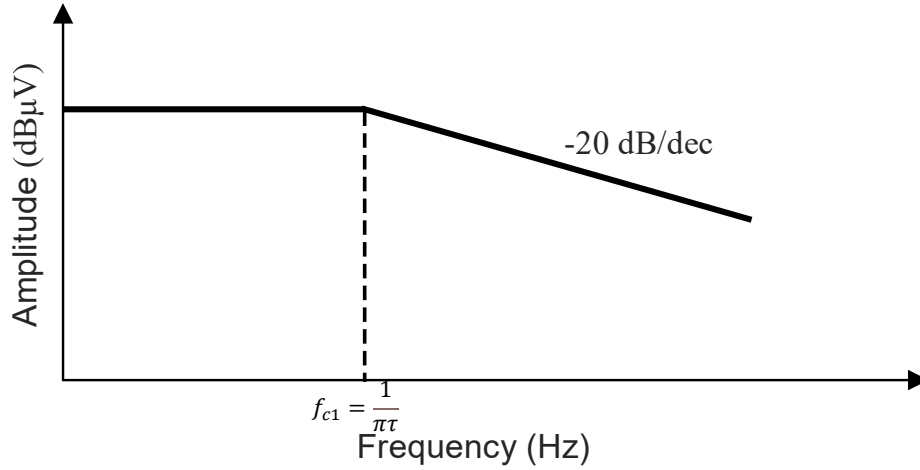


Fig.3.4 The spectral envelope of a periodic rectangular waveform.

3.1.3 Fourier Series Calculation of Trapezoidal Waveform

In a practical circuit, the rise time/fall time of voltage and current waveform cannot be ignored. Therefore, scholars have applied the trapezoid waveform to represent the real switching waveform. Furtherly, the Fourier coefficients and the spectrum analysis would be shown below.

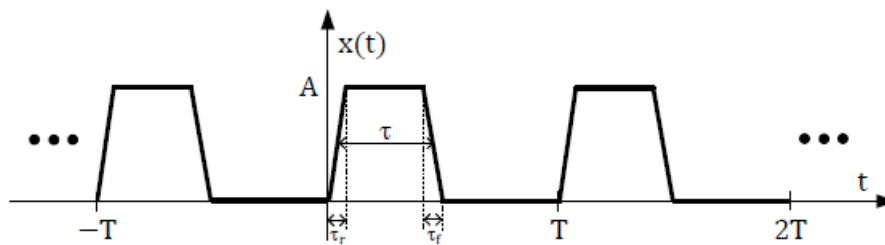


Fig.3.5 Periodic trapezoidal waveform.

The function of the trapezoidal waveform would be expressed as (3.7)

$$x(t) = \begin{cases} \frac{A}{\tau_r} x & 0 < x < \tau_r \\ A & \tau_r < x < \tau + \frac{\tau_r - \tau_f}{2} \\ A - \frac{A}{\tau_f} x & \tau + \frac{\tau_r - \tau_f}{2} < x < \tau + \frac{\tau_r + \tau_f}{2} \end{cases} \quad (3.7)$$

However, it is difficult to calculate the corresponding Fourier coefficients directly according to the expression.

The differentiation properties of the Fourier series could help to simplify the calculation process. After deriving (3.7) twice and then summing up, the Fourier series calculation expression of the trapezoidal wave can be obtained, as shown in (3.8).

$$C_n = \frac{A}{(jn\omega)^2 T} \left[\frac{1}{\tau_r} - \frac{e^{-jn\omega\tau_r}}{\tau_r} - \frac{e^{-jn\omega\left(\tau + \frac{\tau_r - \tau_f}{2}\right)}}{\tau_f} + \frac{e^{-jn\omega\left(\tau + \frac{\tau_r + \tau_f}{2}\right)}}{\tau_f} \right] \quad (3.8)$$

If assuming $\tau_r = \tau_f$, which means the waveform is symmetrical. The Fourier coefficient of a periodic trapezoidal waveform can be expressed by (3.9) with the sinc functions. This substitution could help to estimate the breaking point in the frequency domain performance. The spectral envelope of a symmetric periodic trapezoidal waveform is given in Fig. 3.6.

$$\begin{aligned} C_n &= \frac{A}{(jn\omega)^2 T} \left[\frac{1}{\tau_r} - \frac{e^{-jn\omega\tau_r}}{\tau_r} - \frac{e^{-jn\omega\left(\tau + \frac{\tau_r - \tau_f}{2}\right)}}{\tau_f} + \frac{e^{-jn\omega\left(\tau + \frac{\tau_r + \tau_f}{2}\right)}}{\tau_f} \right] \\ &= \frac{1}{(jn\omega)^2} \cdot \frac{1}{T} \cdot \frac{A}{\tau_r} (1 - e^{-jn\omega\tau})(1 - e^{-jn\omega\tau_r}) = \frac{A\tau}{T} e^{-jn\omega\frac{\tau + \tau_f}{2}} \text{sinc}(n\omega\frac{\tau}{2}) \text{sinc}(n\omega\frac{\tau_r}{2}) \end{aligned} \quad (3.9)$$

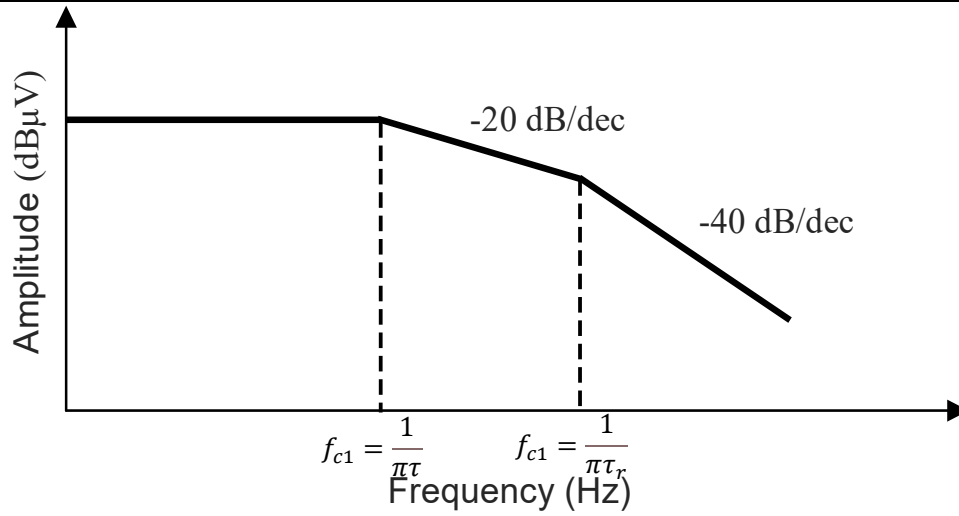


Fig.3.6 The spectral envelope of a symmetric periodic trapezoidal waveform

3.1.4 Lecture Review of Analytical Modeling Method

Since an accurate EMI noise model can help a lot to evaluate the EMI noise results and adopt appropriate attenuation strategies in the development and design stages. Researchers have proposed the analytical model in the frequency domain [42]-[53]. These investigation approaches are based on investigating the generation and propagation route of the EMI noise.

The waveform analysis methods described above are widely used in the early stage EMI noise analysis modeling methods [42]–[45]. Compared with the time domain modeling methods, the analytical modeling approaches speed up the modeling process and offer more accurate predictions. The simplified lumped parameters and ideal geometric shapes are utilized to replace the circuit impedance and the noise source, respectively [41], [45]. The prediction can match well with the experimental spectrum until MHz. However, such approximation has to deal with over-simplification and the accuracy of noise prediction needs further improvement.

Optimized CM Noise Analytical Modeling Method

As the ringing voltage in the real circuit is unavoidable when the switch turns on and turns off, and also may cause certain influences to the conducted noise (150kHz-30MHz). Consequently, it is imprecise to utilize a standard trapezoid shape to represent the real switch voltage waveform. With technological advances, plenty of optimized analytical models have been proposed based on the basic rectangular and trapezoidal waveforms. These schemes have considered the physical characteristics of the semiconductor switch and focused on the nonlinear switching transient process [46]-[52].

Dr. Ma has placed great emphasis on this problem and tried to find a better method to improve model accuracy [46]-[47]. A piece-wise linear approximation method was expanded in [46]. The turn-on and turn-off waveforms have been divided into several stages by using multiple dv/dt and di/dt slopes. However, this method still remains in the experimental verification of the semiconductor device.

Researchers in [48] investigated the dynamic switching process and its frequency dispersion performance. In order to remedy the ringing issue, techniques such as soft switching are applied and the EMI noise is improved accordingly [49]. However, the characteristics of the noise source need further research, which could benefit the noise mitigation design. Then more details of time-domain waveform were included in the noise source model, and the prediction results were improved accordingly [50]-[51]. Nevertheless, it should be noted that the modeling process became more complex at the same time. Additionally, the requirement for the hardware setups, like oscilloscopes and probes were also increased at the same time.

Besides, other mathematical analysis tools, like wavelet transform has been utilized to investigate the voltage waveform [52]. But the proposed continuous wavelet transform

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(CWT) is more complex to understand and implement than the widespread Discrete Fourier Transform.

Overall, one disadvantage of the aforementioned schemes is that data processing requires high-precision probes. In addition, data processing takes a long time to operate even with the help of mathematical software. Zhang in [53] has proposed an optimized scheme that utilized the S-shape waveform to describe the real waveform. The comparison of the spectral envelope between rectangular, trapezoidal, and S-shape waveform is shown in Fig.3.7. In this figure, the τ_{r1} is the period of the ringing voltage. Compared with the square waveform and the trapezoid waveform, the S-shape could reflect the influence of ringing voltage indeed. Nevertheless, this method needs a long processing time to calculate the high order derivative series, and the break frequency is usually beyond the conducted noise frequency band. Consequently, a more general method is proposed in this work to reduce the dependency of the instruments and speed up the data processing.

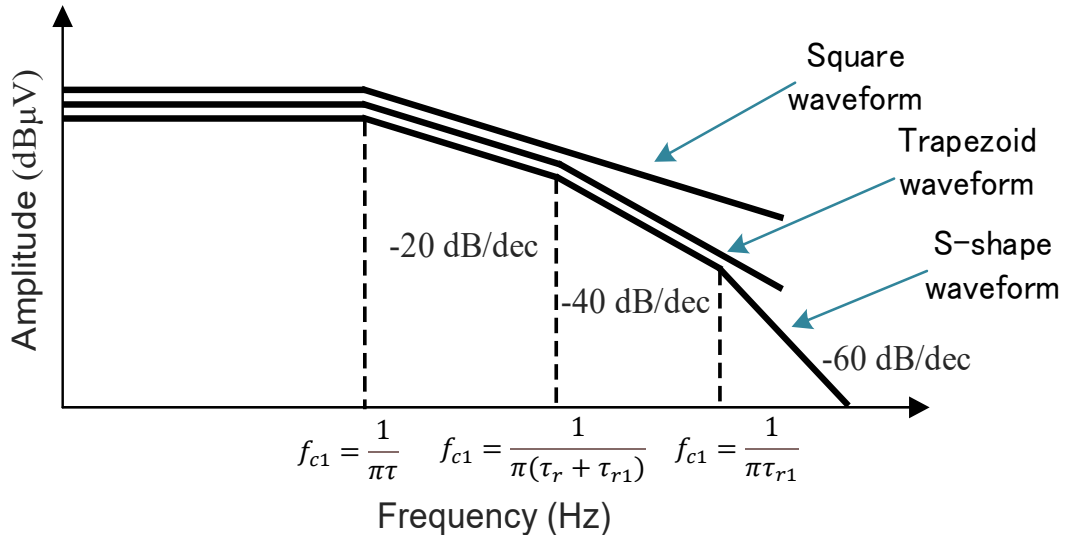


Fig.3.7 Spectra comparison of rectangular, trapezoidal, and S-shape waveforms.

3.2 CM Noise Modeling of the Flyback Converter

Integrated with the AC-DC rectifier circuit, the flyback converters are widely used in power adapters. While the switching frequency of the rectifier is low, most EMI noise is generated by the turn on and off of the switch in flyback converter. In addition, the high-frequency transformer in flyback converter makes the propagation of CM noise current extremely complicated. Therefore, it is necessary to investigate the noise generation mechanism and the propagation route in depth [54]-[61].

At present, most research on the transformer modeling is focusing on the interwinding capacitance and different impedance structures. Researchers in [54] gave the model of interwinding capacitances by comparing the energy conservation and displacement current conservation. The proposed two-capacitor model has a wide range of adaptation and can be extended to multi-winding transformers. Reference [55] was furtherly simplified the transformer model with one capacitor and calculated the leakage inductance of different windings. Nevertheless, neither of the two research included the leakage inductance components in the entirety noise model.

Therefore, researchers have done a lot of related work on the transformer leakage inductance [58]-[61]. [56] discussed the impact of leakage inductance on the CM conducted noise, but the measurement conditions were complex. Furthermore, [57] investigated experimental measurement methods to improve the accuracy of the abstracted parameters in the model. [60] has proposed multiconductor transmission line theory to model the transformer in the converter, the prediction result would match well until 8 MHz. These papers have laid a solid foundation for further research.

In this study, a developed CM noise modeling approach is introduced based on a DC-

DC flyback converter. The accuracy of the noise model has been improved in two aspects: investigating transformer leakage inductance, and considering the dynamic behavior of the switching voltage. The spectra comparison between the calculated and experimental results is provided to validate the feasibility and accuracy of the proposed noise model.

3.2.1 Noise Current Propagation Analysis

The configuration of the flyback converter, connected with the LISN is shown in Fig.3.8. The inner structure of LISN is presented in the red dashed box, which could be considered as a $25\ \Omega$ resistor in the CM noise model within the conducted noise frequency range. In the same frequency band, the input side capacitance C_i could be regarded as a short circuit in the CM noise model. C_p and C_s represent the parasitic capacitances between primary and secondary of the transformer and the ground respectively. Consequently, the other blue dotted line indicates the propagation path of the CM current.

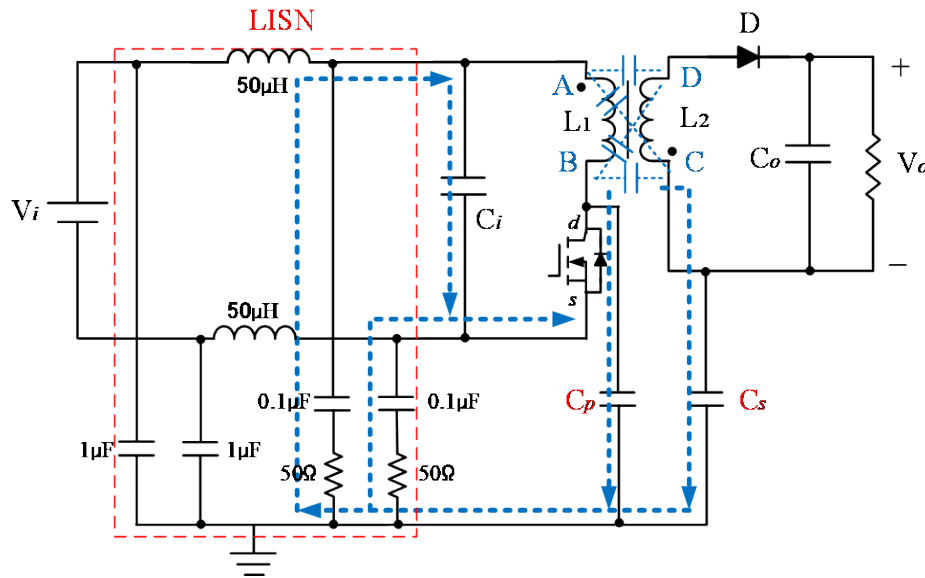


Fig.3.8 Configuration of the flyback converter and LISN setup.

Optimized CM Noise Analytical Modeling Method

Fig.3.9 (a) indicates the CM equivalent circuit of flyback converter after applying the substitution theory. V_N represents the voltage fluctuation between the drain and source of the switch, while I_D indicates the current impulse of the diode. According to the superposition theory, the impact of the noise sources V_N and I_D could be analyzed separately [54]. As shown in Fig.3.9(b), when analyzing the effect of I_D , V_N is replaced by a short circuit. Therefore I_D may not affect the CM noise current flow through LISN. On the contrary, in Fig.3.9(c), I_D is replaced by an open circuit when analyzing the voltage source V_N . Consequently, V_N generates the CM noise, and the noise current flows through the C_P , C_S , interwinding capacitance, and LISN.

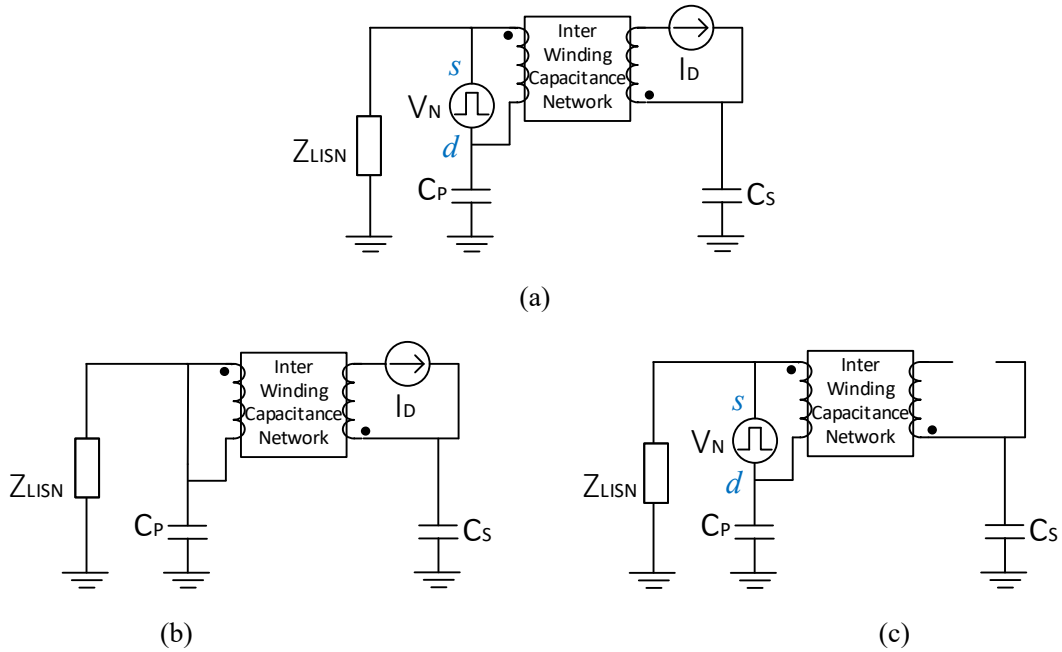


Fig.3.9 (a) CM equivalent circuit of a flyback converter including both noise sources by means of the substitution theory. (b) CM noise model diagram when considering the source I_D only. (c) CM noise model diagram when considering the source V_N only.

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TABLE 3-1 KEY PARAMETERS OF THE FLYBACK CONVERTER

Parameter	Symbol	Value
Input voltage (V)	V_i	48
Switching frequency (kHz)	f_{sw}	103
Turns ratio	m:n	14:7
Load (Ω)	R_o	10
Output voltage (V)	V_o	10

The details of the propagation route in the transformer interwinding are discussed in the next section. Table 3-1 indicates the key parameters of the flyback converter.

3.2.2 Accurate Model of Flyback Transformer

This section discusses the CM noise model of flyback transformer. The turns ratio of the primary and secondary winding is m and n , respectively. As the primary intrawinding capacitance is paralleled with noise source V_N , it would not affect the CM current that flows through LISN.

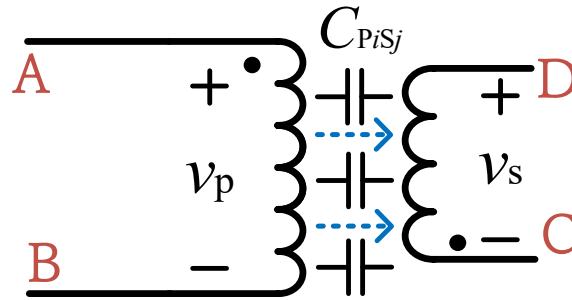


Fig.3.10 The transformer model with leakage inductance [54].

Fig.3.10 shows the preliminary model of flyback transformer in the former study, where C_{PiSj} represents the overall interwinding capacitances [54]. The effectiveness of leakage inductance can be ignored when analyzing the interwinding capacitance. Our study

Optimized CM Noise Analytical Modeling Method

researched the impact of the leakage inductance and improve the accuracy. The measurement of interwinding capacitance is given below. Here the leakage inductance can be ignored temporarily because the transformer is picked out from the circuit.

The primary voltage is V_P and the secondary is V_S . Assuming that each coil has the same voltage drop, then the voltage difference ΔV_P and ΔV_S can be calculated as:

$$\Delta V_P = \Delta V_S = \frac{V_P}{m} = \frac{V_S}{n} \quad (3.10)$$

Then the voltage of arbitrary turn V_{Pi} and V_{Sj} can be written as:

$$V_{Pi} = V_B + (i - 1)\Delta V_P \quad (3.11)$$

$$V_{Sj} = V_C + (j - 1)\Delta V_S \quad (3.12)$$

The total displacement current passing through C_{PiSj} is

$$i_{C_t} = \sum_{i=1}^m \sum_{j=1}^n C_{Pi-Sj} \frac{d(V_{Pi}-V_{Sj})}{dt} \quad (3.13)$$

Submitting (3.10)-(3.12) into (3.13), the displacement current can be simplified as:

$$\begin{aligned} i_{C_t} &= \sum_{i=1}^m \sum_{j=1}^n C_{Pi-Sj} \frac{d(V_{Pi}-V_{Sj})}{dt} \\ &= C_1 \frac{d(V_B-V_C)}{dt} + C_2 \frac{dV_P}{dt} \\ &= C_2 \frac{d(V_A-V_C)}{dt} + (C_1 - C_2) \frac{d(V_B-V_C)}{dt} \end{aligned} \quad (3.14)$$

where C_1 and C_2 are expressed as:

$$C_1 = \sum_{i=1}^m \sum_{j=1}^n C_{Pi-Sj} \quad (3.15)$$

$$C_2 = \sum_{i=1}^m \sum_{j=1}^n C_{Pi-Sj} \frac{i-j}{m} \quad (3.16)$$

Fig.3.11 gives the two-capacitor model of the interwinding capacitance, where C_1 and C_2 are utilized to represent the overall interwinding capacitance. Consequently, the

displacement current could be represented by two electrical potential differences: $V_A - V_C$ and $V_B - V_C$.

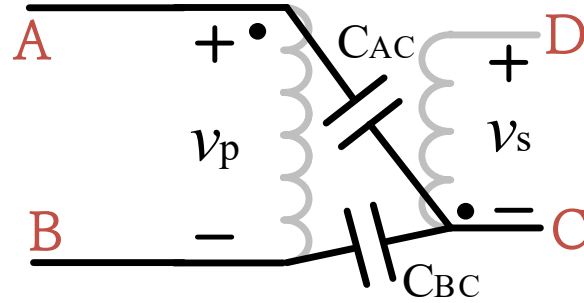


Fig.3.11 Two-capacitor model of the interwinding capacitance.

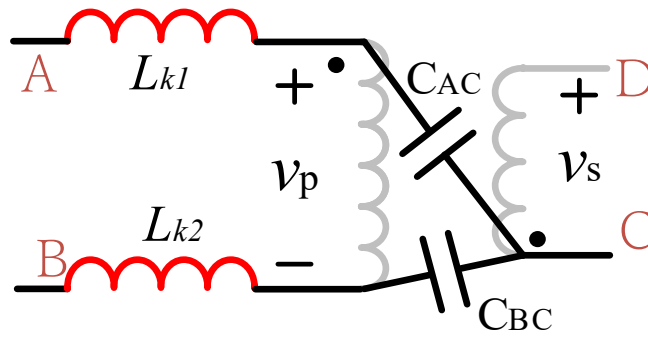


Fig.3.12 Proposed CM noise model of flyback transformer

The proposed model of flyback transformer is shown in Fig.3. 12. Where L_{K1} , L_{K2} represent the primary leakage inductance. This structure is based on the assumption that the leakage inductance is homogeneously distributed with the coils. Therefore, L_{K1} is the same as L_{K2} and is equal to half of the measured primary leakage inductance. Here the secondary leakage inductance is ignored as it is not included in the CM noise current propagation route, and the diode on the secondary side does not generate CM noise.

In conclusion, the identification and measurement of leakage inductance (L_1 & L_2) and the parameter (C_1 & C_2) should follow the three steps below:

Optimized CM Noise Analytical Modeling Method

1) Measure the leakage inductance: connect the primary side of transformer with the impedance analyzer while the secondary is shorted.

2) Confirm the total interwinding capacitance: short-circuit the primary ports (A&B) and the secondary ports (C&D) separately, and then connect both to the impedance analyzer (As is shown in Fig.3.13).

3) Confirm the capacitance ratio by a signal generator. The experimental set up is shown in Fig.3.14. The output impedance of the signal generator is set at high impedance because a smaller current may minimize the impact of leakage inductance.

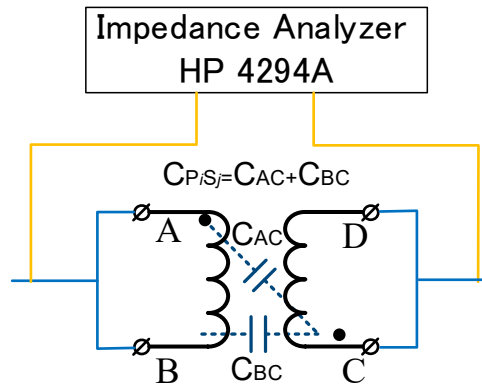


Fig.3.13 Total interwinding capacitance (C_{PISj}) measurement with impedance analyzer

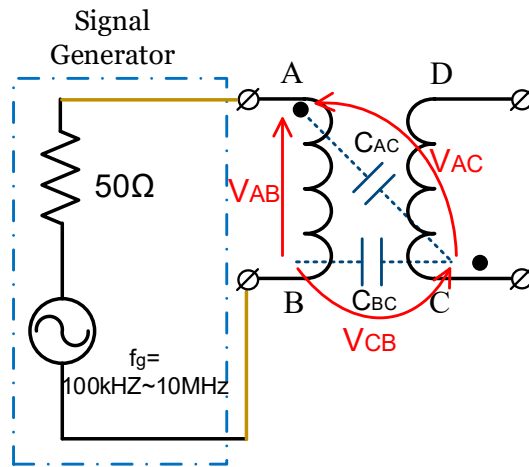


Fig.3.14 Capacitance identification with the signal generator.

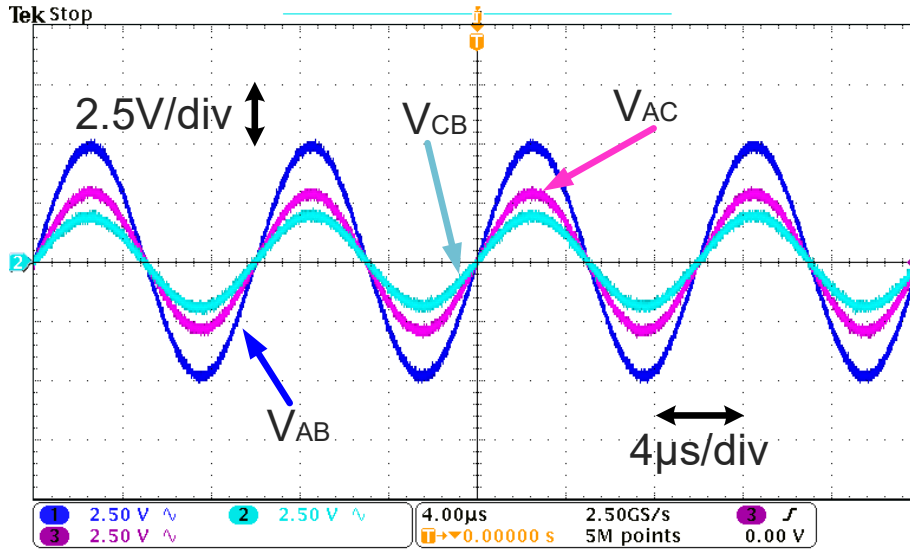


Fig.3.15 Experimental waveforms of V_{AB} , V_{AC} , and V_{CB} .

Take C_{AC} to represent the voltage difference between A and C, C_{BC} represents the voltage difference between B and C. Then the parameters C_1 and C_2 can be calculated accordingly. The relationships between C_{AC} , C_{BC} , V_{AC} , and V_{CB} can be expressed as follows:

$$\frac{C_{BC}}{C_{AC}} = \frac{V_{AC}}{V_{CB}} \quad (3.17)$$

$$V_{AB} = V_{AC} + V_{CB} \quad (3.18)$$

The measured voltage waveforms of V_{AB} , V_{CB} , and V_{AC} are shown in Fig.3.15, where the input voltage of V_{AB} is 5 V, and the voltage ratio of V_{AC} and V_{CB} is 2.7/2.3. Accordingly, the capacitance values are estimated as: C_{Pisj} equals to 23 pF, C_{AC} equals to 10.6 pF, and C_{BC} equals to 12.4 pF.

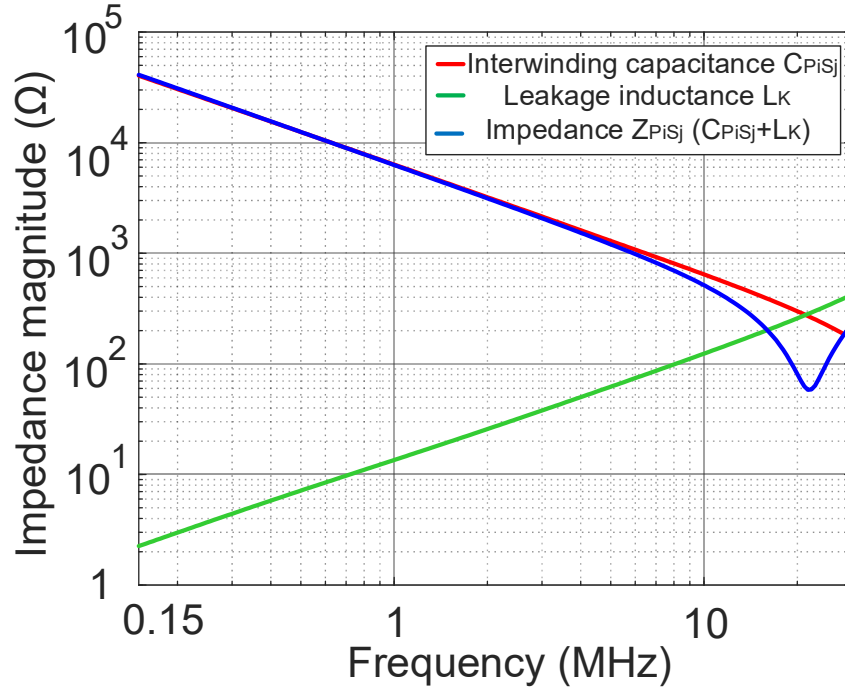


Fig.3.16 Magnitudes comparison with consideration of the leakage inductance.

The measurements show that the primary leakage inductance L_K is 1.9 μH . To emphasize the impact of the leakage inductance, a magnitude comparison is shown in Fig.3.16, where the capacitance C_{PiSj} is drawn in red, the leakage inductance L_K is drawn in green, and the impedance Z_{PiSj} after combining leakage inductance is drawn in blue. Noted that the impact of the leakage inductance becomes obvious at high frequency, which may influence the noise prediction furtherly.

3.2.3 Noise Source Extraction Based on Waveform Decomposition

Since the accuracy of the overall noise model is largely affected by the noise source, more dynamic characteristics should be considered in this noise source model. In order to improve the noise model accuracy and accelerate the mathematical calculation, the voltage waveform of the switch is divided into two parts: trapezoid and high-frequency ringing, and then the Fourier series are calculated separately [61].

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The decomposition diagram is shown in Fig.3.17. The front is the overall waveform, and the back is the ringing and trapezoid. Here the parameters are also marked for more clarification in the following calculation.

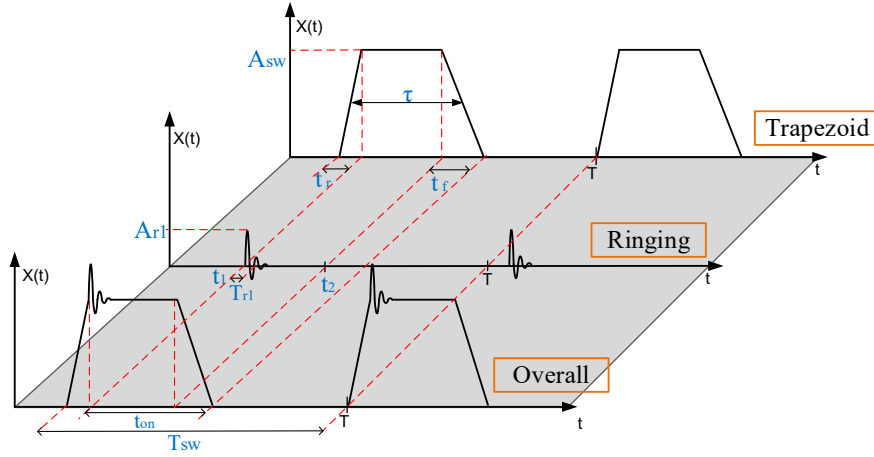


Fig.3.17 Decomposition diagram of switching voltage waveform.

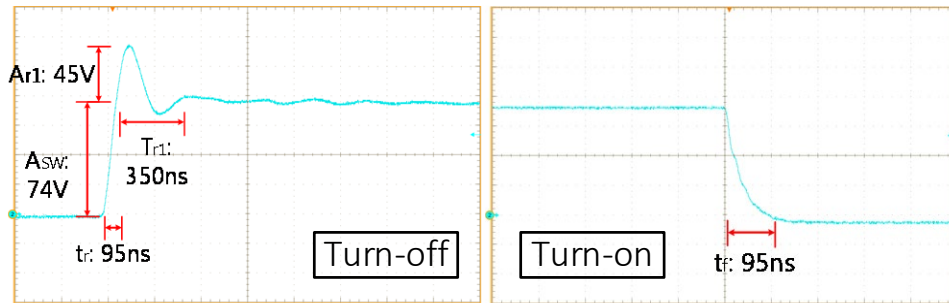


Fig.3.18 Turn-off and turn-on voltage waveforms of the MOSFET.

Fig.3.18 shows the turn-on and turn-off switching voltage waveforms. The parameters are identified as was represented in Fig.3.17. Noted that the turn-on ringing is small enough and therefore has been ignored.

After decomposing the voltage waveform, the overall Fourier series (C_n) could be obtained by calculating the Fourier series. According to the analysis of Section 3.1, the formulas of the trapezoid (C_{n_t}) and sinusoid (C_{n_r}) are presented in (3.19)-(3.21), separately:

Optimized CM Noise Analytical Modeling Method

$$C_n = C_{n_t} + C_{n_r} \quad (3.19)$$

$$C_{n_t} = -j \frac{A_{sw}}{2\pi n} e^{-\frac{jn\pi(\tau+t_r)}{T_{sw}}} \left[\text{sinc}\left(n \frac{t_r}{T_{sw}}\right) e^{-\frac{jn\pi\tau}{T_{sw}}} - \text{sinc}\left(n \frac{t_f}{T_{sw}}\right) e^{-\frac{jn\pi\tau}{T_{sw}}} \right] \quad (3.20)$$

$$C_{n_r} = \frac{1}{T} \int_0^T V_r(t) \cdot e^{-jn\omega} \cdot dt \quad (3.21)$$

Where the time-domain equation of ring $V_r(t)$ could be expressed as a piecewise function:

$$V_r(t) = \begin{cases} 0, & 0 < t < t_1 \\ \frac{A_{r1} \sin \omega_{r1}(t-t_r)}{e^{k_{r1}(t-t_r)}}, & t_1 \leq t < t_2 \\ 0, & t_2 \leq t < T \end{cases} \quad (3.22)$$

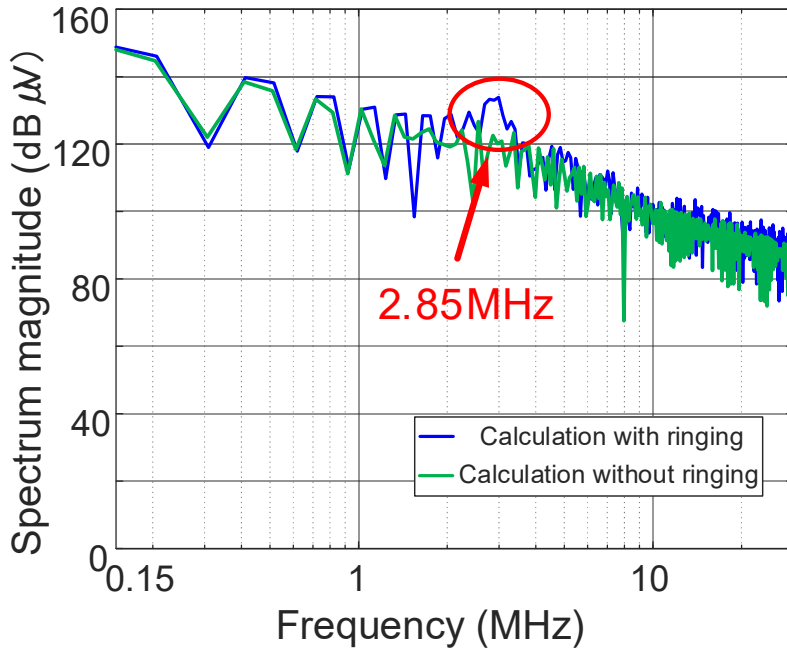


Fig.3.19 The calculated envelopes of the noise source V_N .

The calculated envelopes of noise source are shown in Fig.3.19 in frequency domain, where the blue curve is the result after considering the ringing voltage, and the green curve is the calculation result only with the trapezoidal waveform. The biggest difference is the noise peak at 2.85 MHz. Since the time period of ringing is 350 ns, this noise peak is proved to be caused by the ringing voltage. In addition, ignoring the ringing parts may

further cause the inaccuracy of the noise model.

3.2.4 Model Verification by Experimental Spectra Comparison

In previous sections, the impedance structure of flyback transformer is investigated, and the noise source is obtained in frequency domain as well. Based on the analysis above, the completed CM noise model of flyback converter is shown in Fig.3.20. Then the voltage across LISN could be expressed in (3.23), where the calculation is processed in the frequency domain. Here Z_{LISN} can be taken as 25Ω .

$$V_{CM}(j\omega) = \frac{Z_{LISN}[Z_{AC}(Z_{BC}+Z_{CP}+Z_{CS})+Z_{BC}Z_{CS}]V_N(j\omega)}{[Z_{LISN}(Z_{CP}+Z_{CS})+Z_{CP}Z_{CS}](Z_{AC}+Z_{BC})+Z_{AC}Z_{BC}(Z_{LISN}+Z_{CP})} \quad (3.23)$$

where:

$$Z_{AC} = Z_{L_{k1}} + Z_{C_{AC}} \quad (3.24)$$

$$Z_{BC} = Z_{L_{k2}} + Z_{C_{BC}} \quad (3.25)$$

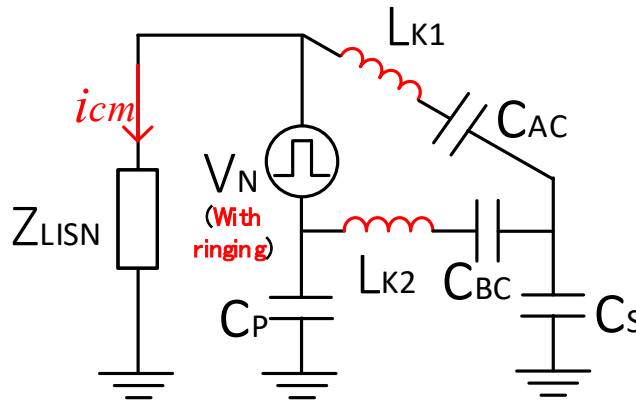


Fig.3.20 Proposed CM noise model of flyback converter.

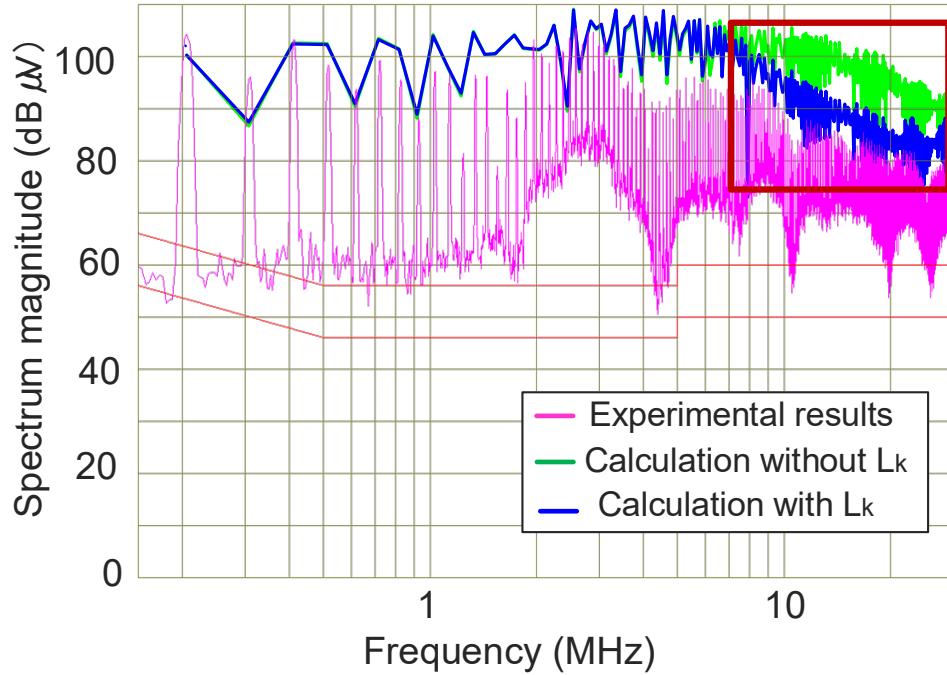


Fig.3.21 Spectra comparison between calculation and experimental results.

A comparison of the noise envelope of both the calculated and measured results is shown in Fig.3.21. The LISN measured results are plotted in pink while the calculation results of ignoring and considering leakage inductance are also given. Noted that the noise peak around 2.8MHz is caused by the ringing waveform, which can match with the calculation in the last section.

Compared to ignoring the leakage inductance, the calculations made after considering the leakage inductance are more consistent with the experimental results, especially in the high-frequency range. This comparison confirms the accuracy of the proposed noise modeling method. Meanwhile, the validity of the proposed interwinding impedance structure and the noise source calculation method are also verified accordingly.

3.3 Conclusions

This chapter introduces a developed CM noise modeling method for the flyback converter. Compared with the previous publications, the accuracy of the CM noise model is optimized mainly in two aspects: the leakage inductance is considered in the noise propagation route and the noise source is decomposed for easing calculation.

In terms of the impedance structure of the interwinding structure, this study not only calculates the displacement current but also discusses the impact of the leakage inductance. Besides, the spectrum of the noise source is obtained by decomposing the voltage waveform and calculating their Fourier series separately. Therefore the dynamic process of the switching voltage could be better considered.

The feasibility and effectiveness of the proposed modeling method are validated by comparison with experimental results. With the help of the proposed noise model, different noise attenuation methods could be predicted. The future research would focus on noise attenuation by utilizing and designing the transformer leakage inductance.

Chapter 4 Common Mode Noise Suppression Study of Impedance Balancing Technique

4.1 Introduction to Impedance Balancing Technique

Except for adding EMI noise filters at the input or output side of the power converter, there are other ways to attenuate the EMI noise. The impedance balancing technique is one of these and has become the focus of extensive discussion [63]. In brief, the noise source could be attenuated greatly in the balancing converter topologies. As an effective method to reduce the conducted noise, the impedance balancing technique has been investigated and implemented in many articles [80] – [92].

Researches in [80] had analyzed the noise generation mechanism of a conventional boost converter and proposed the impedance balanced topology. The noise current of boost circuit is mainly caused by the asymmetry, therefore the improvement is towards the symmetric topology. The implemented process included spilling the power inductor into two same parts, adding another diode at the neutral pole of the output side, and adjusting the parasitic capacitances. The given experimental waveforms showed that the voltages of the two stray capacitances have the same amplitudes and opposite phase angles, which well achieved the design purposes. However, the added diode may increase converter loss and the analysis of balancing methodology need further study.

Kong and Wang had deepened the research and studied the balance conditions in different DC-DC converters [82]-[87]. These balancing converter topologies were designed following the Wheatstone bridge theory [83]. Specifically, the semiconductor

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switches were surrounded by the impedance bridges while the impedance ratios of each two bridge arms remain the same. Therefore, the noise source could be greatly attenuated and the explanation can refer to the Wheatstone bridge theory. In addition, the influence of the unbalanced factors, like the coupled power inductors and the error of the bridges, were analyzed with the proposed noise model [85]. However, this balancing method was also troubled with the problems like high frequency noise peak and the mismatch of the impedance arms.

Furthermore, the balancing technique can be flexibly applied to more complex circuits by adjusting the propagation impedance [89]-[92]. In the case of DC-AC inverters, the general method is to coordinate with the external CM choke circuit [89], [90]. Authors in [92] also discussed the CM noise contributions of rectifier and inverter in an AC-DC-AC system separately, then the Wheatstone bridge balancing technique is implemented in the case of parallel-connected rectifiers. The noise level has been suppressed well in these situations, but there are still some issues that need further investigation. Firstly, serious noise peaks occurred because of the impedance resonance. Moreover, it lacks an accurate model that could effectively estimate attenuation effects.

Another direction of balancing methodology is the advanced topology. A novel balanced inverter topology was firstly introduced in [93]. The novel balanced inverter has three switches in series for each phase leg and this design could attenuate the CM voltage fluctuations during operation. A comparison of the conventional inverter, four-leg inverter, and the balanced inverter was shown in [95], which showed that the low-frequency band noise has been reduced well. Another kind of balanced topology, a system combined with two parallel inverters, was introduced in [96]. In conclusion, the above balanced solutions

CM Noise Suppression Technique Study of Impedance Balancing Technique

may reduce the conducted EMI noise well in low-frequency band, but the performance at high frequency needs further improvement and the load value must be exactly the same to keep symmetric.

In this chapter, a novel CM noise modeling method was introduced to analyze and optimize the impedance balancing technique [97]. Specifically, the attenuation degree of the noise source was investigated by comparing the impedance and spectra information, and the resonance problem was solved effectively by redesigning the bridge impedances. The CM noise model based on conventional boost converter would be investigated at first as the application of balance would not affect the operation of the switch. This accurate noise source method plays a vital role in improving the accuracy of the overall model.

4.2 CM Noise Analysis of Conventional Boost Converter

4.2.1 Propagation Path of CM Noise Current and the Equivalent Model

Similar to the CM noise modeling process of flyback converter which was introduced in Chapter 3, the boost converter noise model extraction method would be shown below.

Fig.4.1 shows the configuration of a conventional boost converter connecting with the LISN. The LISN is placed between the DC power supply and the boost converter circuit, which provides a standard impedance to the EUT and evaluates the conducted noise level by integrating with a spectrum analyzer. P and N are the positive pole and negative pole, separately. G is the frame ground of the circuit and LISN. In conventional boost converter, C_i and C_o are the input side and output side capacitances, and L is the power inductor. C_g is the capacitor that connects the ground and the drain of the MOSFET switch, which simulates the parasitic capacitors between the switch and the heatsink as well as the

capacitors between the heatsink and the frame ground.

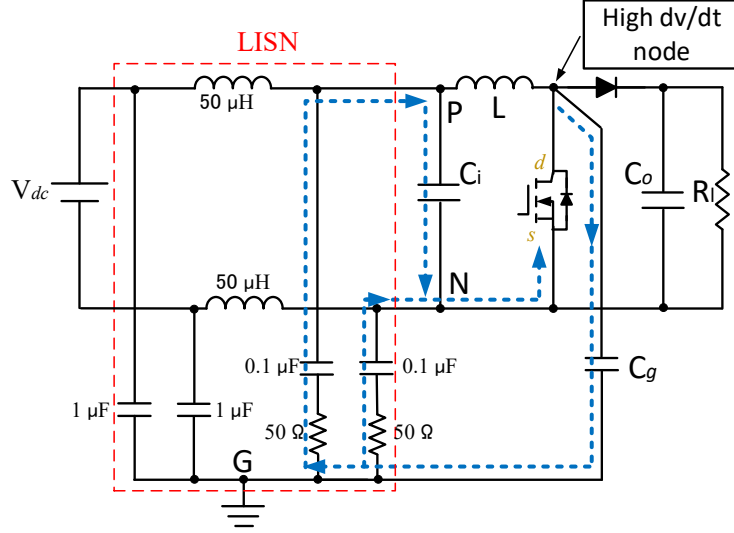


Fig.4.1 Configuration of the conventional boost circuit and LISN setup.

At the conducted noise frequency band, the relative impedance of the capacitor is small and the inductor is large. Therefore, C_i (100 μF) and L (500 μH) could be considered as short and open, respectively, in the CM noise model. The conducted noise was generated by the high-frequency voltage fluctuation between the drain and source of the switch. Here in the conventional boost converter, the source voltage is almost constant even when the switch is working. On the other hand, as marked as the high dv/dt node in Fig.4.1, the drain voltage changes rapidly and large CM current flows through C_g to the ground. In conclusion, the blue dotted line in Fig.4.1 indicates the flow direction of CM current.

The CM noise model resulted from the developed analysis is shown in Fig.4.2. Where Z_{LISN} represents the equivalent resistance of LISN, which could be taken as a 25 Ω resistor in CM noise calculation. V_N is the noise source, represents the voltage between drain and source of the switch (namely V_{ds}). Z_{C_g} is the impedance of C_g . According to the EMI measurement specification, the length of the connecting wire between EUT and LISN

CM Noise Suppression Technique Study of Impedance Balancing Technique

should be equal to 80 cm. Here, to improve the accuracy of the noise model, the impedance of the connecting wire is considered and represented as Z_w .

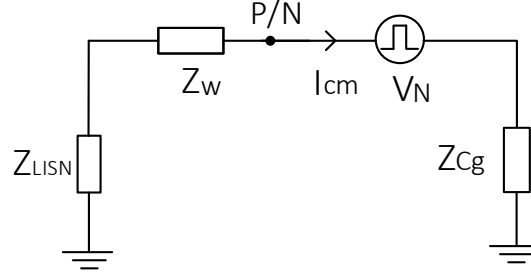


Fig.4.2 CM noise model of the conventional boost circuit.

According to the equivalent noise model, the CM noise voltage of conventional boost converter could be represented by (4.1):

$$V_{LISN} = \frac{Z_{LISN}}{Z_{LISN} + Z_w + Z_{Cg}} V_N \quad (4.1)$$

This equation is true in both time and frequency domains, and the calculations in this paper are performed in a frequency domain.

4.2.2 Noise Source Extraction and the Model Validation

As mentioned in the introduction, the precision of the noise model is largely affected by the noise source. Thus, plenty of methods have been proposed to obtain the noise source in time or frequency domain. In this article, a novel frequency domain modeling method is introduced. The computational complexity is reduced while the accuracy is guaranteed. Specifically, the switch voltage waveform during the experiment was observed and classified in time domain, and then, the overall spectrum could be obtained in frequency domain by calculating the Fourier series by parts. The accuracy of the noise source model is improved remarkably after considering the high frequency ringing.

The general steps for noise source modeling include:

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- (1) Capture the turning on and turning off waveforms of the switch by the oscilloscope, while adjusting the appropriate time scale in order to observe the parameters.
- (2) Divide the overall waveform into three parts: trapezoid, turn-off ringing, and turn-on ringing. Extract the necessary parameters for the Fourier transform.
- (3) Calculate Fourier series of each component separately and then add them together. Finally, the overall spectrum could be obtained.

To specify the modeling process, an experimental setup was established according to Fig.4.1. Details of the experimental configuration are given in the following section. The key parameters of the circuit are given in Table 4-1. In addition, 1nF Y-capacitor was adopted as C_g .

TABLE 4-1. KEY PARAMETERS OF CONVENTIONAL BOOST CONVERTER.

Parameter	Symbol	Value
Input voltage (V)	V_{dc}	12
Switch frequency (kHz)	f_{sw}	103
Duty ratio		0.75
Input capacitance (μF)	C_i	100
Power inductor (μH)	L	500
Load (Ω)	R_l	50

Fig.4.3 shows the voltage waveform of switch measured by an oscilloscope. The ringing caused by turn-on and turn-off could be seen clearly from the following magnified images. Note that the turn-on ringing is quite limited and thus, its effectiveness could be ignored.

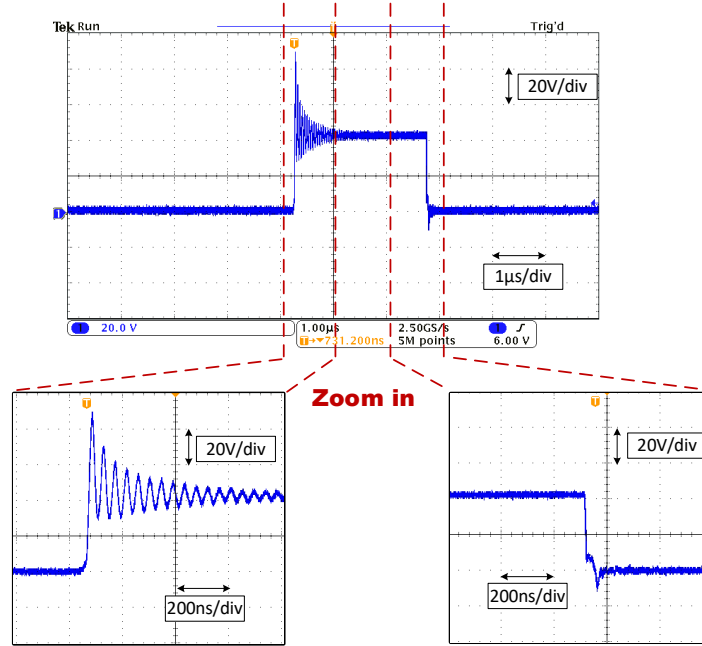


Fig.4.3 The experimental voltage waveform of switch measured by oscilloscope.

As the real voltage waveform is an irregular shape, it is difficult to calculate its spectrum directly. This paper decomposed the switch voltage waveform and the Fourier series could be done in steps. The corresponding decomposition process of the overall waveform is shown in Fig.4.3. The overall waveform is shown at the top of the figure, and it can be divided into the trapezoid component and the ringing component. This decomposition has reduced the complexity of Fourier transformation and improved model precision at the same time. The relative parameters are also marked in Fig.4.4, which could help to identify the corresponding symbols listed in Table 4-2.

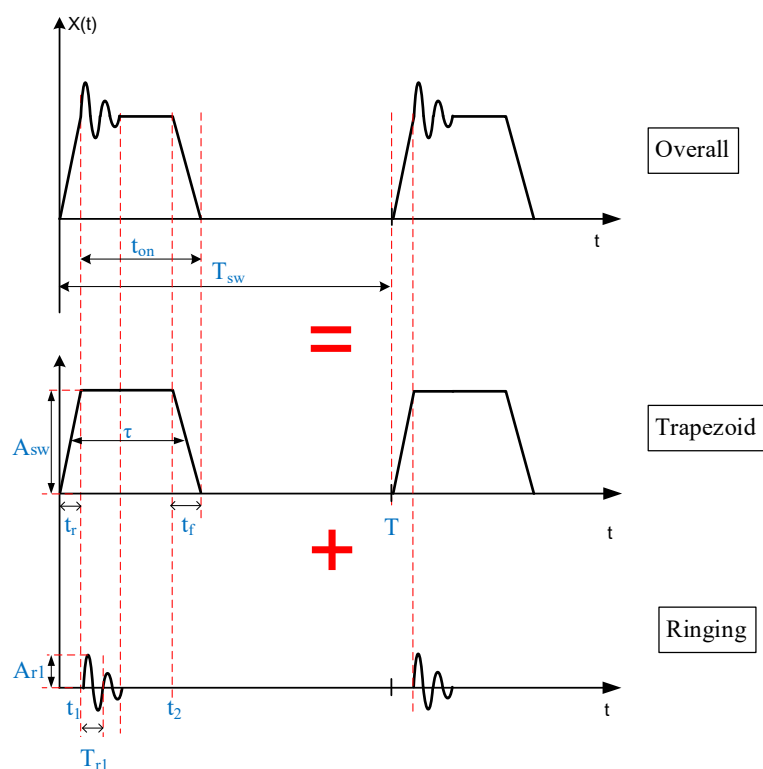


Fig.4.4 Decomposition diagram of the switching voltage waveform.

TABLE 4-2. CHARACTERISTIC PARAMETERS OF THE SWITCHING VOLTAGE WAVEFORM.

Parameter	Symbol	Value
Amplitude of trapezoid (V)	A_{sw}	42.3
Period of the voltage (μs)	T_{sw}	9.7
Switching time period (μs)	τ	2.47
Rising time (ns)	t_r	70
Falling time (ns)	t_f	29
Duration of the turn-off ringing (μs)	t_{on}	2.5
Amplitude of the turn-off ringing (V)	A_{rl}	41.5
Period of the turn-off ringing (ns)	T_{rl}	42.3

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According to the mathematical definition in chapter three, the formula for calculating the Fourier series corresponding to trapezoid could be expressed in (4.2).

$$C_{n_t} = -j \frac{A_{sw}}{2\pi n} e^{-\frac{jn\pi(\tau+t_r)}{T_{sw}}} \left[\text{sinc}\left(n \frac{t_r}{T_{sw}}\right) e^{-\frac{jn\pi\tau}{T_{sw}}} - \text{sinc}\left(n \frac{t_f}{T_{sw}}\right) e^{-\frac{jn\pi\tau}{T_{sw}}} \right] \quad (4.2)$$

In order to approximate the experimental waveform, the difference of turn-on time and turn-off time was also taken into consideration.

Before calculating the Fourier series of the ringing waveform, it is better to clarify the corresponding time-domain expressions. The ringing shown in Fig.4.4 begins after the switch turns off (t_1), and will last until the switch turns on (t_2). Therefore, it could be written as a piecewise function, as is presented in (4.3):

$$V_r(t) = \begin{cases} 0, & 0 < t < t_1 \\ \frac{A_{r1} \sin \omega_{r1}(t-t_r)}{e^{k_{r1}(t-t_r)}}, & t_1 \leq t < t_2 \\ 0, & t_2 \leq t < T \end{cases} \quad (4.3)$$

According to the definition, the formula for calculating the Fourier series can be written as in (4.4). Then, the Fourier series of the ringing part can be calculated by substituting (4.3) into (4.4).

$$C_{n_r} = \frac{1}{T} \int_{\langle T \rangle} V_r(t) \cdot e^{-jn\omega} \cdot dt \quad (4.4)$$

Finally, the Fourier coefficient of the overall switch waveform would be calculated by summarizing C_n and C_{n_r} .

According to (4.1)–(4.4) and the measured data, the Fourier series calculation was completed with Matlab 2015b. Fig.4.5 shows the noise source spectra comparison between experimental and calculation results. The calculation results are plotted in the blue line, and the red experimental results refer to the FFT series computed by the

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oscilloscope. The comparison shows that the calculation results match the experimental result in all frequency ranges. The high-frequency peak is caused by the ringing, and this frequency (around 23.5 MHz) agrees with the frequency of the ringing waveform in Fig.4.3 ($T_{r1} = 42.3$ ns). The effectiveness of this frequency-domain modeling method and the obtained parameters were well verified.

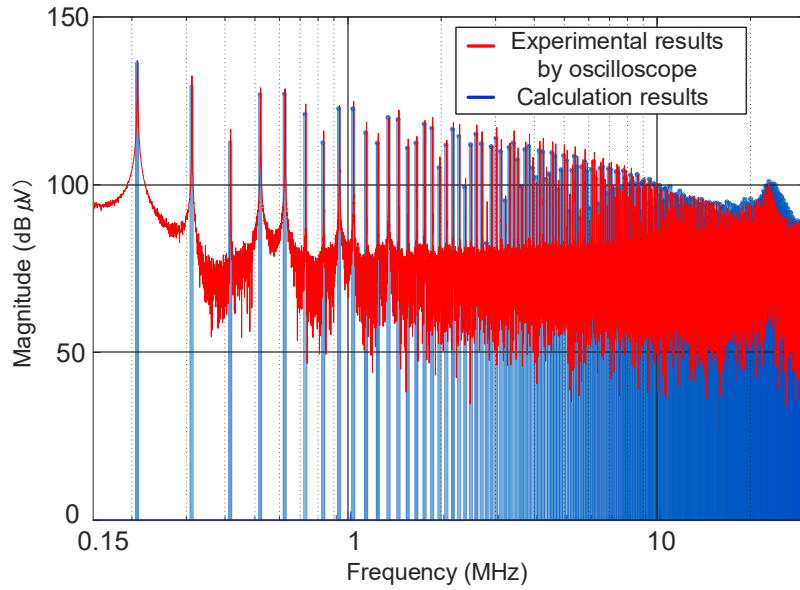


Fig.4.5 The frequency spectra of switch voltage (noise source) comparison between calculation and measured results.

Furthermore, based on (4.1), Fig.4.6 shows the prediction result of boost converter by utilizing the calculated noise source. The experimental result measured by LISN is also given in green. It can be observed that the prediction result matches the measurement in the whole frequency band, and this model is able to predict CM noise with good accuracy.

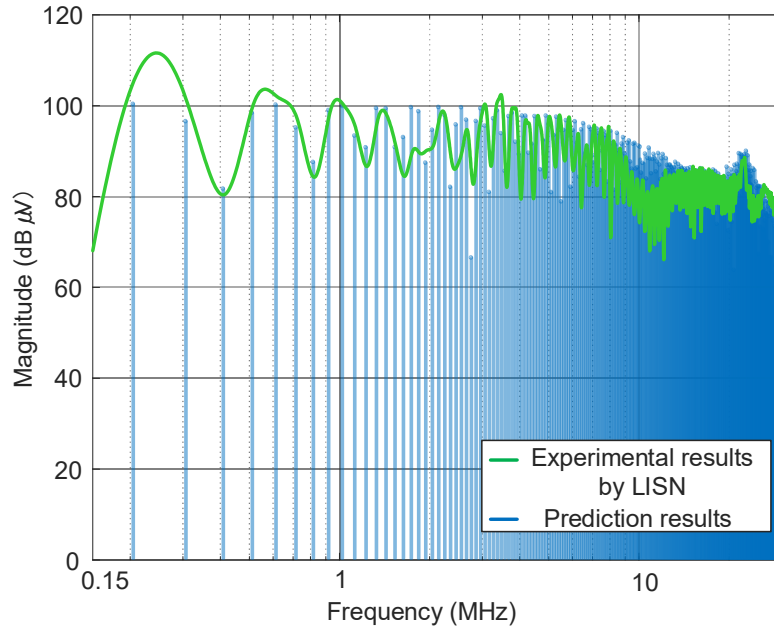


Fig.4.6 CM noise spectra comparison between prediction and measured results.

4.3 Mechanism Analysis of the Balanced Boost Converter

Unlike the conventional passive or active EMI filters, the impedance balancing technique could reduce the conducted noise by changing the topology of conventional boost converter. The configuration of balanced boost converter along with LISN is shown in Fig.4.7. Compared with Fig.4.1, the power inductor L was split into two parts (L_1 and L_2) and another capacitor C_2 was placed between the N phase and the ground. This action changes the path of the noise current without affecting the efficiency of the circuit.

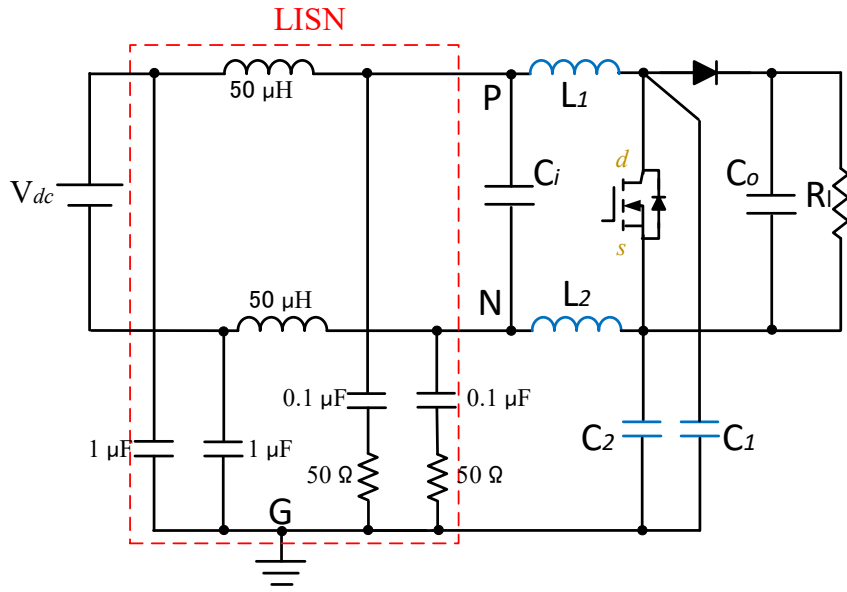
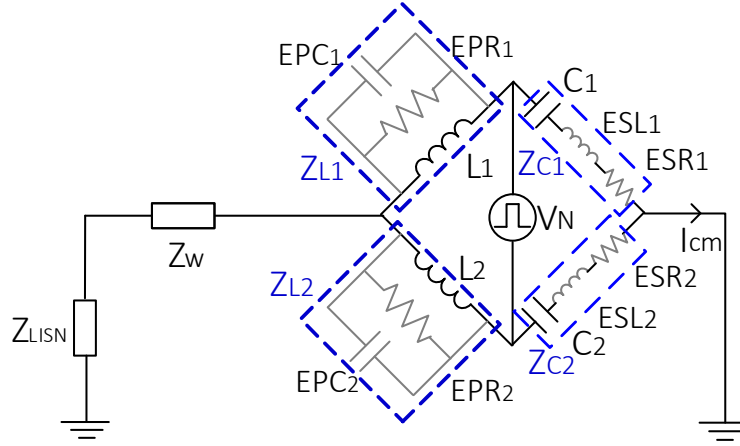


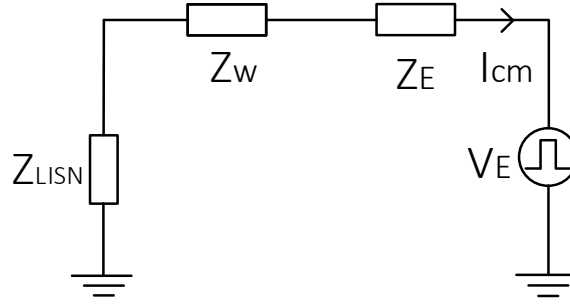
Fig.4.7 Configuration of the balanced boost converter and LISN.

4.3.1. CM Noise Model of Impedance Balanced Boost Circuit

The corresponding CM noise model of balanced boost circuit is shown in Fig.4.8(a). The model includes the equivalent parallel capacitor (EPC) and equivalent parallel resistor (EPR) of each inductor, as well as the equivalent series inductor (ESL) and equivalent series resistor (ESR) of each capacitor. These parameters may affect the self-resonant frequency and cause impedance unbalancing as well. In addition, a more simplified model obtained according to the Thevenin theorem is shown in Fig.4.8 (b).



(a)



(b)

Fig4.8 (a) CM noise model of the impedance balanced boost circuit; (b) simplified CM noise model by the Thevenin theorem.

V_E and Z_E in the figure are the equivalent noise source and the equivalent impedance, respectively. V_N is the same with conventional boost converter since the driver circuit of the switch is the same with the boost circuit, and the implementation of balancing impedance would not affect the efficiency of the circuit. The relationships between V_E , Z_E , V_N , and the impedance bridge arms are shown in (4.5)–(4.6).

$$Z_E = \frac{Z_{L1}Z_{L2}}{Z_{L1}+Z_{L2}} + \frac{Z_{C1}Z_{C2}}{Z_{C1}+Z_{C2}} \quad (4.5)$$

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$$V_E = \left(\frac{Z_{L2}}{Z_{L1} + Z_{L2}} - \frac{Z_{C2}}{Z_{C1} + Z_{C2}} \right) V_N \quad (4.6)$$

where Z_{L1} , Z_{L2} , Z_{C1} , and Z_{C2} are the impedance of each arm. Furthermore, the voltage of LISN can be expressed as in (4.7), which has a similar form as (4.1).

$$V_{LISN} = \frac{Z_{LISN}}{Z_{LISN} + Z_w + Z_E} V_E \quad (4.7)$$

Specifically, no CM current would flow through LISN if the impedance of each bridge arm could satisfy the condition in (4.8):

$$\frac{Z_{L1}}{Z_{L2}} = \frac{Z_{C1}}{Z_{C2}} = n \quad (4.8)$$

where n represents the impedance ratio.

According to the above analysis, the most important step of a balanced circuit is selecting the appropriate impedance arms and keeping the impedance ratio (n) stable during all the target frequency bands. The coupled inductor has been previously used in publications due to its smaller size [94]–[96]. However, unexpected problems have also arisen. Firstly, the coupling coefficient not only affects the inductance value after decoupling but also imposes specific requirements on the selection of core. In [95], the toroidal core has been replaced because it is hard to acquire a high coupling coefficient when the turn ratio is high. Secondly, the parasitic capacitors between the primary side and secondary side make the route of CM noise propagation exceedingly complicated. This component has greatly increased the difficulty of modeling analysis and may cause poor reduction results at the same time. Therefore, two isolated inductances were adopted in this study to prevent these consequences.

In this article, as the power inductor in the conventional boost circuit is 500 μH , two

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250 μH inductors (L_1 , L_2) and two 1nF Y-capacitors (C_1 , C_2) were utilized to build the balanced circuit. The magnitude spectra of impedance bridge arms are shown in Fig.4.9. The figure shows that both inductors and capacitances are well matched over the frequency range. The condition of balancing impedance is properly satisfied.

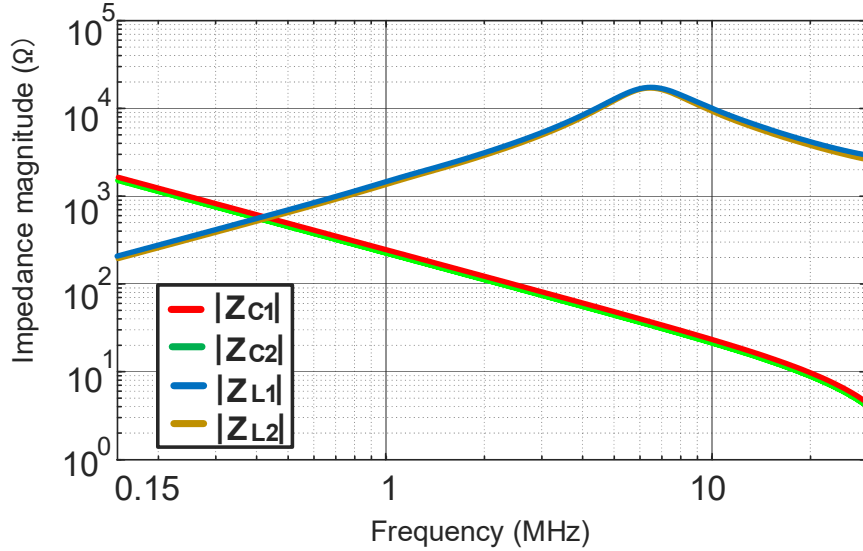


Fig.4.9 Magnitude comparison of bridge arms impedance (Z_{C1} VS. Z_{C2} and Z_{L1} VS. Z_{L2}).

Fig.4.10 shows the experimental CM noise spectrum of balanced boost circuit, which is plotted in red. Meanwhile, the green envelope is the CM noise of conventional boost circuit (same as in Fig.4.6). By utilizing the balanced technique, the noise level was well attenuated from 0.5 MHz and upwards. The effect of this technique is obvious, as much as 40dB reduction was done at some frequency points. Nevertheless, the noise peak at low-frequency bands degraded the performance. The relative analysis and solution are given in the following sections.

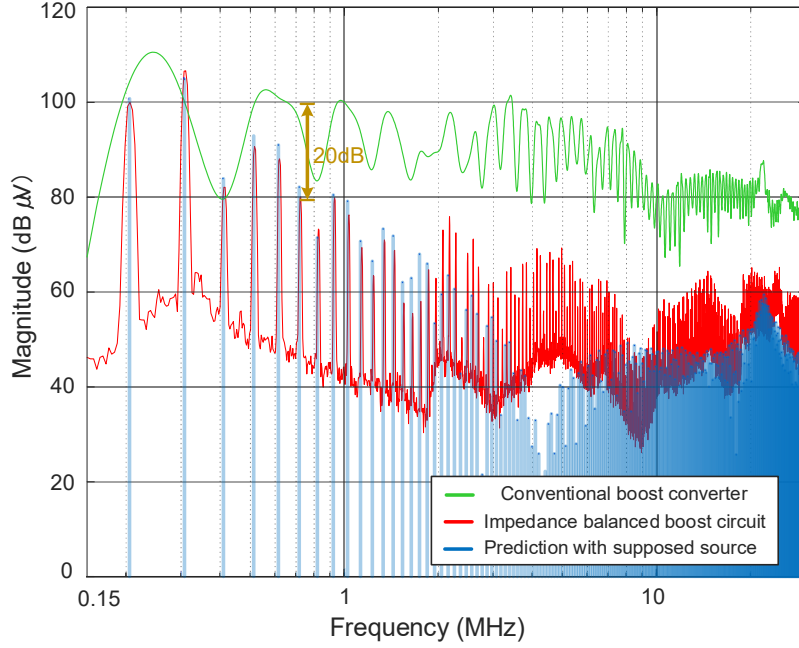


Fig.4.10 CM noise spectra comparison of balanced impedance circuit.

As is shown in Fig.4.10, the CM noise of balanced boost circuit did not drop to zero, even if the impedance arms are almost balanced. This means that (4.8) could not calculate the equivalent noise source accurately. This error may be caused by the interference of the PCB layout and the inevitable mismatch of the impedance. Therefore, this article proposed a novel method to evaluate the attenuation level of the noise source and give predictions accordingly.

Comparing the voltage of LISN in the conventional boost converter and the balanced boost circuits, which are represented in (4.1) and (4.7), respectively, it should be noted that the two formulas have the same form and the difference is the impedance Z_E and Z_{Cg} in the denominator. This also means that the first half of (4.1) and (4.7) are the same when Z_E equals to Z_{Cg} , and then the differences of the two noise sources can be directly deducted from the CM noise spectra, as shown in Fig.4.10.

The equivalent impedance of balanced boost circuit Z_E could be calculated by (4.5), or

CM Noise Suppression Technique Study of Impedance Balancing Technique

measured directly. The error between the two results is small. Referring to (4.5), Z_{E0} represents the entire impedance of 250 μH inductors (L_1 , L_2) paralleled with 1nF capacitors (C_1 , C_2). Fig.4.11 shows the impedance variation of Z_{Cg} and Z_{E0} through frequency. The Z_{Cg} is larger than Z_{E0} from 0.15 MHz to 0.69 MHz, and then the Z_{E0} becomes larger from 0.69 MHz.

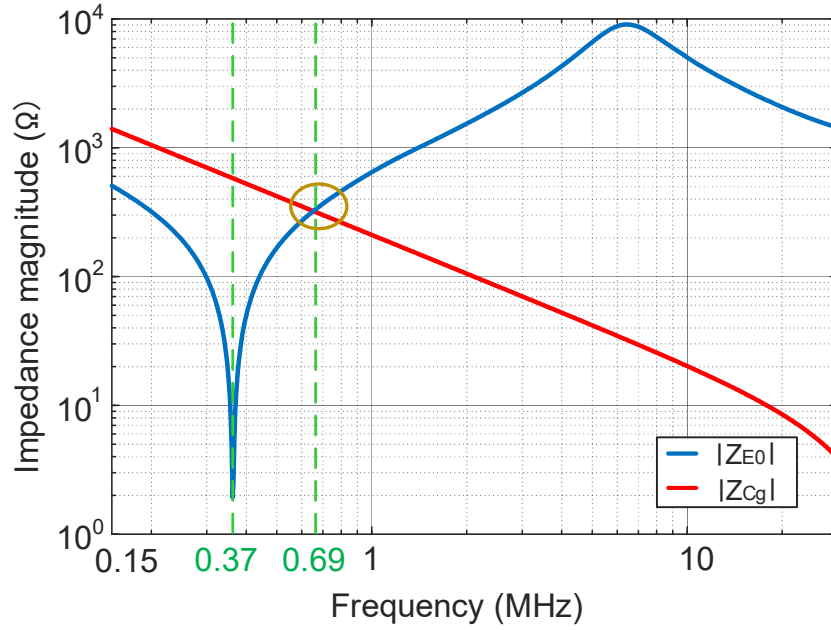


Fig.4.11 Impedance magnitude comparison between Z_{E0} (250 μH and 1 nF) and Z_{Cg} .

According to Fig.4.11, it should be noted that Z_{Cg} equals Z_{E0} at the point of 0.69 MHz, where the noise of balanced circuit is 80 dB μV and the conventional boost converter is about 100 dB μV . This spectra difference equals to 20 dB μV , which means that V_N is about 10 times of V_E according to the definition. Therefore, this study assumed that V_E is one-tenth of V_N and then substituted other measured impedance values into (4.7) for calculation.

The prediction results are plotted in translucent blue in Fig.4.10 and are basically consistent with the measured spectrum except for the medium frequency band. This

comparison confirms the hypothesis.

In addition, the resonance frequency of impedance arms could be calculated by (4.9).

$$f_c = \frac{1}{2\pi\sqrt{(L_1//L_2)(C_1//C_2)}} \quad (4.9)$$

Especially, when L_1 and L_2 equal to 250 μH , C_1 and C_2 equal to 1nF, the series resonance of Z_{E0} happens at 0.37 MHz, which agrees with the noise peak in Fig.4.10. Therefore, it is necessary to find corresponding countermeasures to optimize the known impedance balancing technique.

4.4 Optimization of Balanced Schemes with Redesigned Parameters

As shown in Fig.4.10, the noise peak occurs at the resonant frequency; thus, several researchers have proposed different methods to solve the spike problem. In [92], the authors changed the inductance value and added a pF-class capacitor to cooperate with the impedance ratio in (4.8). However, this may push the noise spike to the middle or high frequency band. The designing method of the inductor in [97] achieved satisfactory results, but this result is based on an additional CM choke filter. As opposed to previous methods, this paper adjusted the values of passive elements to move the resonance peak out of the conductive frequency range.

As was mentioned in the previous section, a low-frequency peak occurs because of the series resonance among the four impedance bridge arms. On the other hand, it is hard to further suppress the equivalent noise source because the impedance arms are well matched, especially at a low frequency. Thus, the more practical method is to redesign the value of inductors or capacitors. Without changing the balancing configuration shown

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in Fig.4.7. This paper has proposed two alternative schemes to attenuate the noise peak according to (4.9). One is increasing the inductor to 1.4 mH and the other is increasing the value of Y-capacitors to 4.7 nF. The following sections will investigate and compare the two approaches.

4.4.1. Balanced Circuit with 1.4mH Inductors and 1nF Y-capacitors

After calculation using (4.9), the inductance should be larger than 1.3 mH if the capacitors C_1 and C_2 are kept constant. Here, two 1.4 mH inductors were utilized to replace the 250 μ H inductors.

The experimental results were recorded and drawn with black in Fig.4.12, compared with the former balanced boost circuit in red and the green envelope of conventional boost converter. These experimental results show that the proposed balanced scheme with 1.4 mH inductors and 1nF capacitors could attenuate noise level overall frequency range compared with conventional boost converter. Furthermore, the spike at 0.37 MHz was effectively suppressed. However, the noise at high frequency increased in contrast with the former balanced boost circuit (250 μ H and 1 nF). The valley at 0.8 MHz is due to the self-resonance between the 1.4mH inductor and its EPCs.

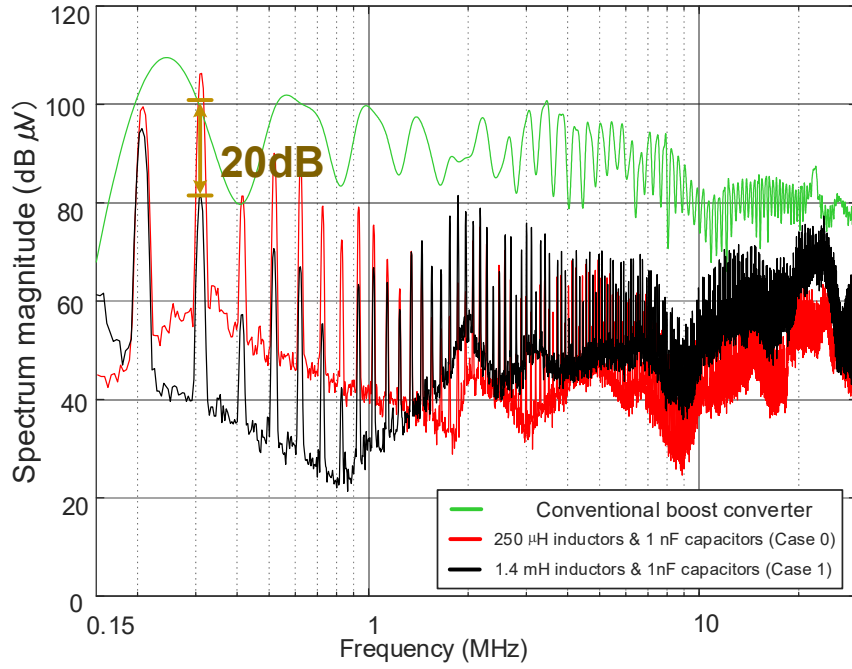


Fig.4.12 CM noise comparison after replacing the 250 μ H inductors with 1.4 mH inductors.

Referring to (4.5), Z_{E1} identifies the whole impedance including 1.4 mH inductors and 1nF Y-capacitors. Fig.4.13 gives its magnitude in black and makes a comparison with Z_{Cg} in red. Note that Z_{E1} equals to Z_{Cg} at 0.24 MHz, which means that the noise source attenuation degree could be speculated quickly With the same analysis method as the one in the last section.

According to Fig.4.12, the noise spectra difference around 0.3 MHz reached about 20 dB μ V, which indicates that the equivalent noise source V_E in the optimized scheme was attenuated well compared with conventional boost converter. However, a drawback of this approach is the increased weight and volume of the circuit.

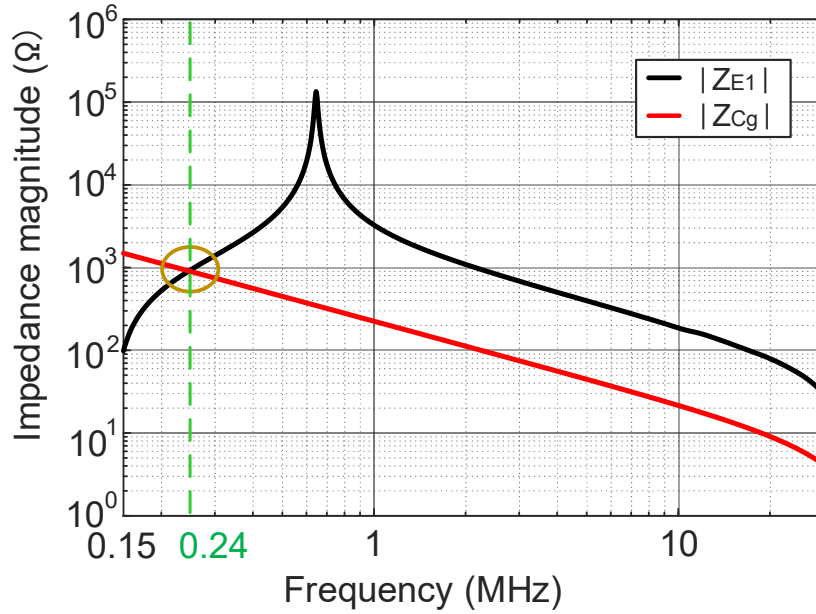


Fig.4.13 Impedance magnitude comparison between Z_{E1} (1.4 mH and 1 nF) and Z_{Cg} .

4.4.2. Balanced Circuit with 250 μ H Inductors and 4.7 nF Y-capacitors

Other than using a pair of larger inductors, this scheme utilized greater Y-capacitors to decrease the series resonant frequency. After calculation, 4.7 nF capacitors were selected to replace the former 1nF capacitors and the other experimental conditions were kept the same.

The CM noise spectra comparison is shown in Fig.4.14. The purple curve is for the case of an optimized solution after replacing the 1 nF Y-capacitors with 4.7 nF, the green one is the noise envelope of conventional boost converter, and the red one is for the former balanced circuit boost (250 μ H and 1 nF). The comparison shows that after utilizing the scheme with 4.7 nF capacitors, a low-frequency spike was also moved out from the frequency band of interest and the CM noise was significantly attenuated. Meanwhile, at a high frequency, the noise after adopting 4.7 nF capacitors is lower than the former case

using 1 nF capacitors.

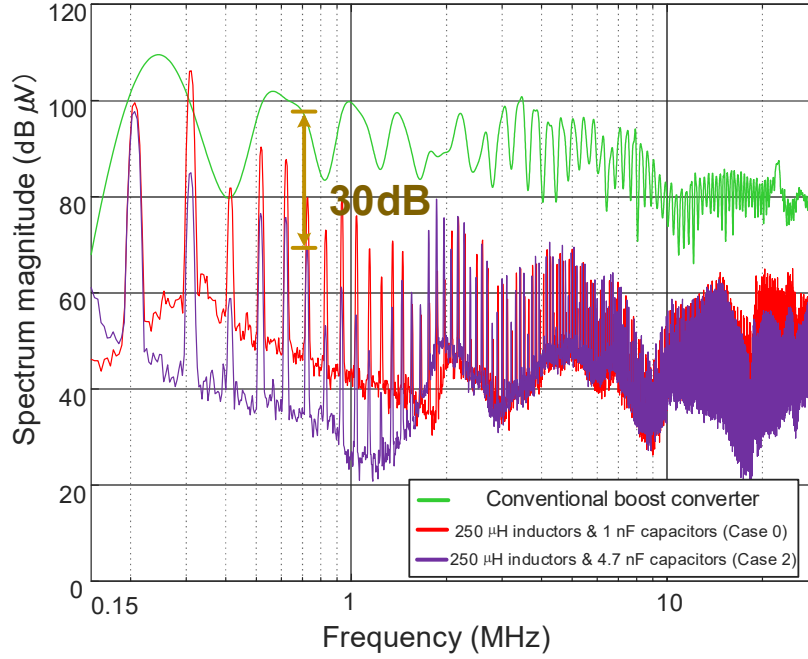


Fig.4.14 CM noise comparison after replacing the 1 nF Y-capacitors with 4.7 nF Y-capacitors.

Similar to Z_{E1} , Z_{E2} represents the whole bridge impedance of 250 μ H inductors and 4.7 nF Y-capacitors. The comparison between Z_{E2} (purple) and Z_{Cg} (red) is shown in Fig.4.15. The resonant frequency of optimized impedance Z_{E2} was removed to around 0.16 MHz and the performance at high frequency is almost the same compared to that in Fig.4.11. At 0.68 MHz, where Z_{E2} equals Z_{Cg} , about 30 dB reduction was achieved according to Fig.4.14. Compared with the former balanced boost circuit (250 μ H and 1 nF), the noise source was significantly attenuated and the performance was improved over the entire frequency range.

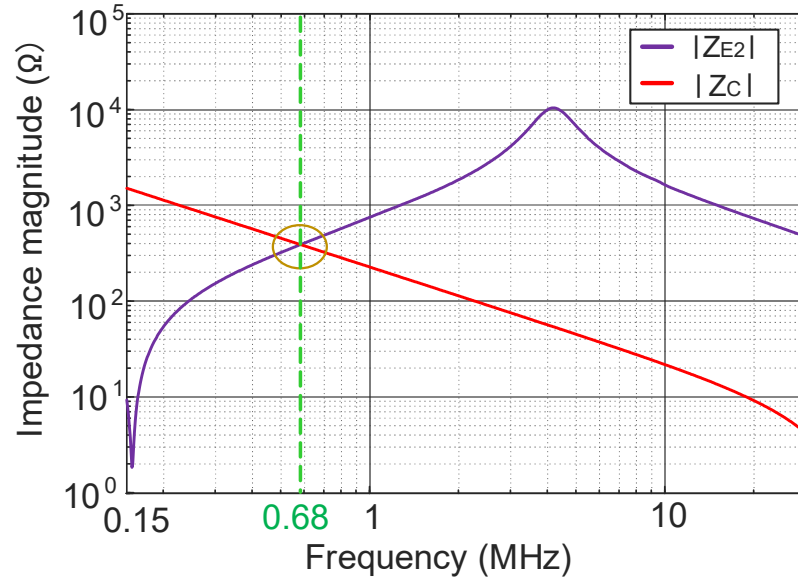


Fig.4.15 Impedance magnitude comparison between Z_{E2} (250 μ H and 4.7 nF) and Z_{Cg} .

There may be some concerns that the greater capacitor may increase the CM current and result in excess of relative standards. Fig4.16 shows the CM currents comparison between conventional boost converter and the optimized balanced boost circuit (250 μ H and 4.7 nF). The results show that with optimized schemes, the CM current becomes even smaller because the noise source has been effectively attenuated.

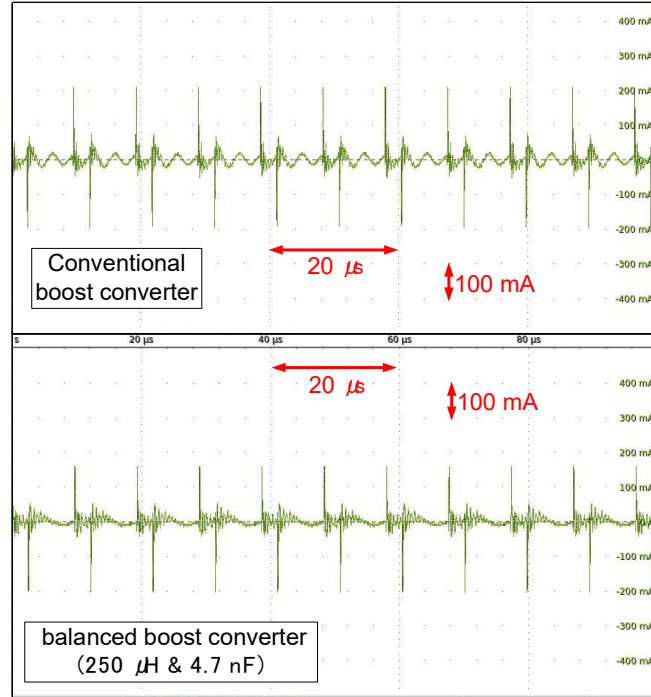


Fig.4.16 Experimental CM current waveforms comparison between conventional boost converter and optimized impedance balancing boost converter (250 μ H and 4.7nF).

In summary, the optimized scheme of changing the capacitors is more effective than replacing the inductors, which not only considers the attenuation results but also the cost and the difficulty of implementation.

A comparative investigation between the proposed work and the previously published relevant arts was carried out and is addressed in Table 4-3. The comparison reveals the effectiveness and practicability of the proposed schemes.

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TABLE 4-3. COMPARISON OF NOISE OPTIMIZATION EFFECTS WITH FORMER WORK.

	Noise Reduction Effect	Attenuated the Resonance Peak?	Required Additional Filter?	Practicability and Feasibility
[84]	Medium	No	Yes	Medium
[85]	Medium	No	No	Medium
[86]	Limited	No	Yes	Limited
[87,88]	Medium	Yes	Yes	Medium
Proposed work	Medium	Yes	No	High

4.4.3 Comparison of Balanced Technique and Passive Noise Filter

As was introduced in the front chapters, a choke circuit can increase the propagation impedance and is set at the front end of the converter. However, the shortage is bulky and weight are increased at the same time.

In this section, a CM noise filter consisted of a choke inductor and two Y-capacitances is utilized. Its effectiveness is compared with the proposed impedance balancing scheme. The choke inductance is 1.4mH and the Y-capacitance is 1nF. The configuration is shown in Fig.4.17 where the choke filter is implemented at the input side of the conventional boost converter. Here the effectiveness of passive noise filter and the optimized balanced scheme, which was introduced in 4.3.2 (250 μ H and 4.7nF) would be evaluated. The basic exoerimental conditions are the same with the Table 4.2.

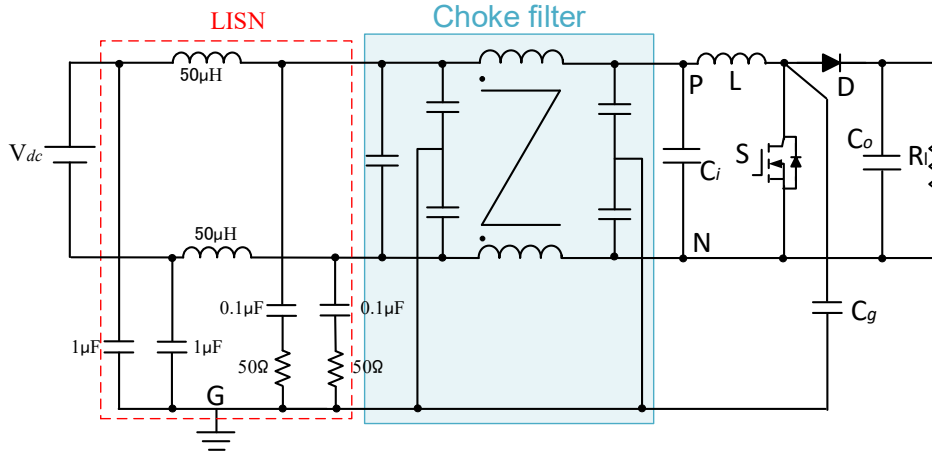


Fig.4.17 Configuration of conventional boost converter after installing the choke circuit.

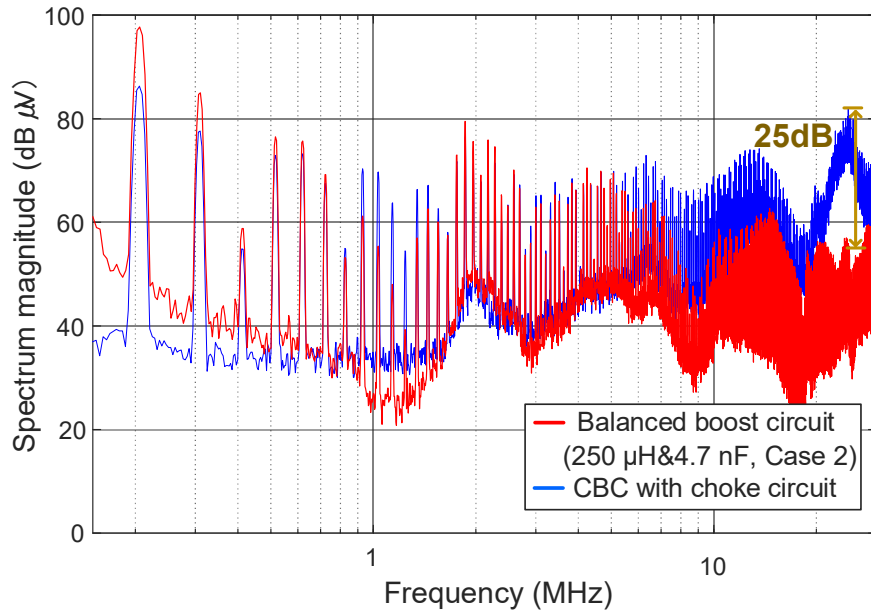


Fig.4.18 CM noise spectra comparison of optimized balanced boost circuit and the conventional boost converter after installing a choke filter.

Fig.4.18 gives the spectra comparison measured by LISN. The blue spectrum is the result of the choke filter, while the red one is the optimized balanced boost circuit. Although the choke filter is better in the low-frequency range, the balanced circuit could reduce the noise level more after 5MHz. The maximum difference is over 25dB. Therefore, the validity of the impedance balancing technique is verified.

4.5 Conclusions

Based on the basic impedance balancing configuration and the calculation formula of resonance frequency, two optimized schemes were proposed to deal with the noise peak in the low-frequency band. The size of the impedance arms was redesigned and the two schemes were investigated in different aspects. Finally, the experimental results were presented to demonstrate the analysis. The contribution of this research is enhancing the feasibility and effectiveness of the known impedance balancing technique.

In order to overcome the shortages of the traditional impedance balancing technique, this research tries to modeling the balanced boost converter and then improved the performance by redesigning the impedance parameters. Firstly, the modeling method is introduced with the conventional boost converter. The detail of noise source extraction is given, and the accuracy of the overall model is guaranteed.

Then the impedance balancing technique is adopted and the relative noise model is built up with the similar modeling method. Based on the model, the attenuation degree is investigated properly. Furthermore, the optimized scheme with redesigned parameters is proposed. Comparing with the traditional balanced converter and the choke filter circuit, this scheme can avoid the noise peak, and save the bulky at the same time. The experimental results are presented to demonstrate the feasibility and effectiveness of the proposed schemes.

Chapter 5 Conclusions and future work

5.1 Conclusions

This thesis has made a major improvement on the CM EMI noise modeling and reduction technique of power electronic converters. The contributions of this work are summarized below.

1. An improved behavioral modeling method on the basis of the terminal port theory is proposed. This modeling method is utilized in the DC-DC boost converter and DC-AC single-phase inverter to evaluate and optimize the noise filter design. The main features of the proposed modeling method are as follows:
 - a. The internal structure of the terminal port box is investigated. Compared to the former behavioral modeling methods which treated the converter as a “black box”, this proposed structure could help to identify the noise source and CM impedance.
 - b. The accuracy of the noise impedance extraction is improved with the interpolation algorithm. As the modeling process is processed in the frequency band, the magnitude and the phase angle of the impedance are most significant. However, the previous behavioral models were depended on the physical model from the impedance analyzer. The proposed interpolation algorithm could help to modify the measurement results for the frequency domain calculation directly.
 - c. Not only the EMI filter at the input side but also the harmonic filter at the

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output side are evaluated with the proposed model. The additional harmonic filters may change the route of the noise current, and then affect its performance of the conducted noise. The comparison between experimental results and the prediction results could verify that different kinds of filters could be evaluated with the proposed noise model.

2. A novel analytical modeling method is proposed to analyze the CM noise in flyback converter. The analytical modeling method is based on the analysis and understanding of the noise generation mechanism. Nevertheless, the previous work has certain limitations. The new scheme proposed in this study could optimize the modeling process and improve the accuracy of the noise model.

- a. The voltage across the switch is first decomposed into trapezoidal and ringing parts. Then calculate the corresponding Fourier series respectively. The necessary parameters for calculation are extracted from the experimental voltage.
- b. The inverse Fourier transform from the frequency domain to the time domain is also computed, which is utilized to compare with the experimental results. The error from waveform decomposition could be corrected in this way. Therefore, the complex switching transient characteristics and ringing are included in the noise source, and the accuracy of the calculated noise source is verified.
- c. The noise propagation route of flyback transformer is investigated. Not only the interwinding capacitance but also the leakage inductance of the primary side are considered in the noise model. The comparison between the

Chapter 5

experimental spectrum and the calculated results shows that the leakage inductance plays a certain role and cannot be ignored.

3. The EMI noise mitigation technique with balanced impedance is analyzed and optimized based on a boost converter. The previous balance method is deeply troubled by impedance mismatch, unclear noise source suppression effect, and noise peaks at the conducted frequency band. Therefore, this research proposes optimized schemes to solve these problems, and the characteristics are mainly reflected in the following aspects:
 - a. The analytical noise modeling method proposed in the third chapter is utilized to analyze the balanced boost converter at first, which could help to investigate the noise current propagation in the balanced boost converter.
 - b. The noise source attenuation degree is evaluated by combining the noise spectra and impedance information.
 - c. Based on the noise model, the reason that caused for noise peak is studied. Furthermore, two optimized schemes are proposed to improve the performance.

5.2 Future Work

1. Extend the application scope of the behavioral modeling method, especially to the more complex converter system.
2. In flyback converter, study how to use the leakage inductance of the transformer to suppress conducted noise.
3. Investigate the effectiveness of the balancing technique in the near field EMI noise research.

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