Electromagnetic interference prediction of power converter including nonlinear behavior of inductive component

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**ELECTROMAGNETIC INTERFERENCE PREDICTION OF
POWER CONVERTER INCLUDING NONLINEAR
BEHAVIOR OF INDUCTIVE COMPONENT**

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Summary

Electromagnetic compatibility (EMC) is one of the important design considerations for power convertor design. Development of electromagnetic interference (EMI) noise model in power convertor design forms a key part for EMI simulation and prediction purposes. With an accurate EMI noise model, the EMI filter can be systematically designed to comply with the relevant EMC standards with good confidence. In addition, the power quality of the power system has become a major concern due to more nonlinear loads connected to the power grid, especially the switching power supplies in most consumable electronics. To reduce power harmonics due to these power converters, passive or active power factor correction (PFC) circuits are part of power converter designs. Active PFC converter can achieve very high power factor but it requires a high frequency active switch, which increases the EMI generated by the convertor. Interleaved PFC converter is a popular topology in PFC converter due to its lower EMI noise and larger output power capacity. The main benefits of the interleaved boost converter when compared with the conventional non-interleaved boost converter are that the inductor volume, the current rating of the semiconductors, and the input current ripple can be reduced. Flyback PFC converter is a simple and yet efficient topology and therefore becomes a popular choice for most applications.

In any converter design, inductors and transformers are critical components. To achieve large inductance with smallest possible size, these inductors and transformers are wound on magnetic cores with very high permeability. To simplify the power convertor design process, they are usually considered to be linear with constant inductances. In reality, their inductances depend on the current as well as the frequency due to nonlinearity of the magnetic cores. In this thesis, a comprehensive mathematical model has been developed that takes into account the nonlinear behavior of the magnetic core so that the actual operating condition of the inductor or transformer in the power converter can be emulated. The model allows more realistic conducted EMI prediction of the power convertor, which prevents overdesign of EMI filter for EMC compliance purpose. Using the interleaved

boost converter and the flyback converter as design examples, the model that includes the nonlinearity of the inductor or transformer has been validated experimentally.

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List of Abbreviations

PFC	Power Factor Correction
EMI	Electromagnetic Interference
EMC	Electromagnetic Compatibility
EMS	Electromagnetic Susceptibility
SPMS	Switched Power Mode Supply
L	Line
N	Neutral
J-A	Jiles-Atherton
AC	Alternative Current
DC	Directive Current
DM	Differential Mode
CM	Common Mode
PCB	Printed Circuit Board
RF	Radio Frequency
CE	Conducted Emission
RE	Radiated Emission
CS	Conducted Susceptibility
RS	Radiated Susceptibility
LISN	Line Impedance tabilization Network
TDMS	Time Domain Measurement System
TDEMS	Time Domain EMI Measurement System

DSO	Digital Signal Oscilloscope
EUT	Equipment Under Test
IF	Intermediate Filter
RMS	Root Mean Square
ADC	Analog-to-digital converter
CCM	Continuous Conduction Mode
CRM	Critical Conduction Mode
DCM	Discontinues Conduction Mode
BCM	Boundary Conduction Mode
HF	High Frequency
CT	Current Transformer
AEF	Active EMI Filter
EPL	Equivalent Parallel Inductor
ESL	Equivalent Series Inductor
EPR	Equivalent Parallel Resistor
ESR	Equivalent Series Resistor
EPC	Equivalent Parallel Capacitor
ESC	Equivalent Series Capacitor
DE	Differential Evolution
ZCD	Zero Current Detection
DE	Differential Evaluation
PWM	Pulse Width Modulation
kHz	kilohertz

MHz	Megahertz
UPS	Uninterruptible Power Supply
IC	Integrated Circuit
CISPR	Committee International Special des Perturbations Radio
IL	Insertion Loss
MMF	Magneto-Motive Force
POL	Point Of Load
DFT	Discrete Fourier Transform
FFT	Fast Fourier Transform
dB	Decibel

List of Symbols

I_{DM}	DM current
I_{CM}	CM current
V_{Line}	Line noise voltage
$V_{Neutral}$	Neutral noise voltage
V_{CM}	Common mode noise voltage
V_{DM}	Differential mode noise voltage
V_{Noise}	Noise voltage
V_{Limit}	Noise limit
V_{DS}	Drain source voltage
V_{DIO}	Secondary diode voltage
V_{EMI}	EMI noise voltage
V_{rec}	Rectified input voltage
V_o	Output voltage
V_{ref}	Reference voltage
S_{IF}	IF signal
j	Simulation step
Z	Impedance
L	Inductor
L_{CM}	CM filtering inductor
L_{DM}	DM filtering inductor
τ	Transient time
ψ	Flux
f	Frequency

$L_{\lambda P}$	Primary leakage inductor
$L_{\lambda S}$	Secondary leakage inductor
L_m	Magnetic inductor
C_{GS}	Gate-source capacitance
C_{DS}	Drain-source capacitance
C_{GD}	Gate-drain capacitance
C_{in}	Input capacitor
C_{PG}	Parasitic capacitance from the drain node of switch to the ground.
C_w	Inter-winding capacitor
F	Magneto-motive force
B	Flux density
H	Magnetic field
H_0	Initial magnetic field
H_e	Effective magnetic field
H_{air}	Air gap magnetic field
μ_0	Free space or air permeability
A_c	Core area
M_{irr}	Irreversible magnetization
M_{an}	Anhysteretic magnetization
M_s	Saturation magnetization
M_{rev}	Reversible magnetization
M	Total magnetization
M_0	Initial magnetization
α	Inter-domain coupling factor
k	Domain wall pinning constant (irreversible magnetization component)

c	Domain wall bowing parameter(reversible magnetization component)
∂	Shape parameter
$H_{j,M}$	Measured magnetic field
$H_{j,S}$	Modeled magnetic field
$B_{j,M}$	Measured flux density
$B_{j,S}$	Modeled flux density
H_c	Coercive force
B_r	Remanence flux density
X_r	Remanence differential susceptibility
X_m	Maximum differential susceptibility
X_{in}	Initial differential susceptibility
D	Core diameter
n	Turn number
S	Switch
d	Duty cycle
t_{on}	Switch turn-on time
t_{off}	Switch turn-off time
R_d	Drain-to-source turn-on resistance
L_1, L_2	Boost inductor
L_d	Drain equivalent series inductor
L_s	Source equivalent series inductor
$A_{req,DM}$	Required filter attenuation
f_{DM}	DM emission frequency
$f_{DM,cor}$	Corner frequency
A_L	Inductance per turn

L_{air}	Air gap length
i_{L1}, i_{L2}	Boost inductor current
i_{DS}	Drain-source current

Chapter 1 Introduction

1.1 Background

Electromagnetic interference (EMI) is a disturbance that affects an electrical circuit due to either electromagnetic induction or radiation emitted from external source. The disturbance may degrade the functional performance of the victim circuit. The sources of EMI usually carry rapidly changing electrical voltages or currents [1-6]. Electromagnetic compatibility (EMC) is a situation when a system with several circuits co-exists together without exhibiting EMI phenomena. With higher operating frequencies of electronic devices, EMC has become an important design consideration for electronic circuits and systems [7-10].

Switching power converters, without exception, generate high frequency EMI that couples into the power mains shared by other circuits through conduction. For this reason, international EMC standards specify strict conducted EMI limits for power converters. The switching elements in power converters are usually semiconductor devices with operating frequency in the range of tens of kHz to several MHz. During the repetitive turn-on and turn-off operations, they give rise to very fast voltage and current transients (di/dt and dv/dt) and these transients are major sources of conducted EMI [11, 12].

The major sources of conducted EMI from power convertors have been well reported [13-15]. In order to quantify the contribution of each noise source to the resultant conducted EMI, simulation models that include parasitic elements of the circuit components have been studied [16-19] and it has been shown that the EMI noise models must be constructed correctly for accurate EMI prediction [20-23]. Besides propagation paths and components' parasitic, accurate EMI prediction also requires proper models of noise sources. For simplicity, most noise models are assumed to be linear but such assumption may not be always true, especially the noise models involve inductive components with magnetic cores.

The Jiles-Atherton (J-A) model [24-27] was adopted to develop the EMI noise prediction models for an interleaved PFC converter, and a flyback PFC converter have been carefully investigated. The ability of the J-A model to handle the nonlinear behavior of the inductive elements in both converters was validated with experimental results.

1.2 Motivation and Objectives

To accurately model a power convertor circuit for the purpose of EMI prediction and analysis, proper modeling of circuit components are critical. The ability to construct an accurate EMI noise prediction model before realization of a prototype is necessary so that early EMI problems can be identified and rectified to meet the required EMC specifications. Inductive components, such as inductors and transformers with magnetic cores are common sights in power convertor circuits. Accurate modeling of their behaviors plays a key role for accurate EMI prediction. Therefore, there is a need to develop a model that can handle the non-linear hysteresis of magnetic core so that the non-linearity of the inductive components can be included in power convertor circuit design.

The scope of this thesis focuses on EMI prediction and analysis of power factor correction (PFC) converters with the non-linear behaviors of inductive components included. Based on the literature survey, two commonly used techniques to model non-linear hysteresis of magnetic core are Preisach and Jiles-Atherton (J-A) methods. The J-A model is adopted in this thesis, as it has the ability to predict the magnetic hysteresis behavior accurately [28, 29].

The objectives of the thesis are as follows:

- Mathematical modeling of non-linear behavior of magnetic material;
- Integration of non-linear hysteresis model of inductive components in the PFC converters for conducted EMI prediction;
- Systematic EMI filter designs for PFC converters; and
- Experimental validation with simulated conducted EMI results.

1.3 Thesis Organization

The background and motivation of the thesis are discussed in this chapter.

Chapter 2 gives a comprehensive overview of EMI related to power converters.

Chapter 3 describes the noise sources, coupling paths and components modeling in the power converters.

Chapter 4 introduces the theory of J-A model and its numerical implementation to handle the non-linear hysteresis characteristic of the magnetic core.

Chapter 5 shows how the J-A model is implemented in an interleaved PFC converter with the non-linearity of the boost inductor considered in the simulation for EMI predication and analysis.

Chapter 6 extends the J-A model in a flyback PFC converter for EMI analysis and reduction.

Finally, Chapter 7 concludes the thesis and recommends possible future work.

Chapter 2 Review of Power Converter EMI

2.1 Introduction

Due to mandatory international EMC regulatory requirements, all electrical and electronic products, including power converters, have to comply with the relevant EMC standards [30-34]. There are three basic elements for an EMI occurrence, the source, the coupling path and the victim, as indicated in Fig. 2-1.

Fig. 2-1. Three basic elements for an EMI occurrence

An EMI situation occurred if the received interfering signal causes the victim circuit to malfunction. To limit the likelihood of EMI, the coupling path must be made as inefficient as possible for the disturbance source. In general, the coupling path can be in the forms of conduction and radiation, as illustrated in Fig. 2-2 (a) and (b), respectively. For the EMI coupling through conduction, there must be electrical conductors connecting the source and the victim. These electrical conductors can be shared power supply or grounding wires. For EMI coupling through radiation, the source transmits the disturbance to the victim through free space in the form of electromagnetic wave.

Fig. 2-2. EMI coupling paths: (a) conducted EMI (b) radiated EMI [35]

From Fig. 2-2, it is quite clear that the first step to suppress EMI generated by a power converter is to pay EMC design consideration so that the emission is minimized at the source (the power converter) and then a built-in EMI filter is designed to reduce the emission further for EMC compliance purpose.

2.2 EMC Standards

To ensure EMC, a system must fulfill both electromagnetic interference (EMI) and electromagnetic susceptibility (EMS) requirements, as indicated in Fig. 2-3.

Fig. 2-3. Classification of EMC [36]

Table 2-1 Overview of the electromagnetic disturbance phenomena

Conducted low-frequency phenomena	Harmonics, voltage fluctuations, voltage unbalance, power frequency variations, induced low-frequency voltages.
Radiated low-frequency field phenomena	Power frequency magnetic fields (continuous and transient), low frequency electric fields.
Conducted high-frequency phenomena	Directly coupled or induced voltages or currents (continuous wave and modulated wave).
Radiated high-frequency field phenomena	High frequency electric fields (continuous wave, modulated wave and transients)
Electrostatic discharge phenomena (ESD)	Human electrostatic discharge, machine electrostatic discharge.

EMI emission level refers to the magnitude of electromagnetic disturbance emitted by equipment-under-test (EUT) and EMS level refers to the magnitude of a given electromagnetic disturbance that begins to cause the EUT's performance to degrade. For

EMC assurance purposes, EMS limit specified in an EMC standard usually is much higher than the EMI limit. The difference between the EMI and EMS limits are referred to as the immunity margin, which is a key factor for EMC assessment of a EUT [37].

The Committee International Special des Perturbations Radio Electriques (CISPR) under IEC is mainly responsible for international EMC standardization so that it avoids discrepancies between different standards in different countries and regions. Different electromagnetic disturbance phenomenon is given in Table 2-1 [38].

For power converters, the lower frequency harmonic levels are controlled up to 50th harmonics of the mains frequency or 150 kHz. The RF conducted emissions are controlled from 150 kHz to 30 MHz [39]. As mentioned earlier, conducted EMI noise sources are usually responsible for EMC related problems in power converters [40, 41] and the recommended suppression techniques are reducing EMI at the source first [42] and then followed by adding EMI filter, such as inductive chokes and capacitors, to bring the EMI level below the required limit [43].

Fig. 2-4. EN 55015 conducted emission limit [44]

As an example, the applicable EMC standard for luminaires is EN55015 and the conducted emission limits are shown in Fig. 2-4. The red and black lines are the average and quasi-peak limits, respectively. Usually, the frequency range applied is from 150 kHz

to 30 MHz but if the switching frequency of the power converter is lower than 150 kHz, the limit below 150 kHz must also be applied [44].

2.3 EMI from Power Converters

2.3.1 Conducted EMI from Power Converters

High frequency switching techniques are being employed in power converters for size reduction and efficiency improvement. However, the rapid switching of the power transistors also generates much conducted emissions [45]. These conducted emissions are usually in the form of harmonics of the switching frequency, which can easily extend up to several tens of MHz. The conduction modes of conducted emissions as shown in Fig. 2-5 can be either common-mode (CM) or differential-mode (DM) in nature [46]. For EMI analysis purposes, the CM and DM conducted emissions are usually separated, which is quite well-defined and understood for the single-phase or the DC systems [47, 48]. For three-phase converter systems, they are more complicated and there are no universal CM and DM definitions. However, separation of CM and DM conducted emissions in three-phase systems is also possible if a symmetrical, linear and time invariant three-phase system is considered [49-52].

Fig. 2-5. CM and DM noise current paths in a single-phase system [35]

2.3.2 Power Factor Correction (PFC) Converter

Power converters using switching techniques are basically related to non-linear loads, which result in poor power factor. In many applications, a power factor correction (PFC) circuit is included in the power converter design to improve its power factor. In general, there are two types of PFC circuits, the active and passive PFC circuits [53, 54].

The most commonly used PFC circuit is the passive PFC circuit, where a capacitor is added at the AC input to correct the power factor. However, it only works for a specific AC input voltage and has limitation on the maximum achievable power factor. The active PFC circuit has the ability to correct the power factor up to 0.95, very close to the ideal power factor and hence reduces the total harmonic distortion significantly. In addition, its automatic AC input voltage correction enables the converter to work with large input voltage variation. However, all these merits are at the expense of higher cost due to the complexity of the PFC circuit. The most common topologies for the active PFC converters are the boost converter operating in continuous conduction mode (CCM), average current mode and critical conduction mode (CRM).

A power converter consists of a power conversion circuit and an electronic circuit. The power conversion circuit converts an unregulated AC or DC input power to a regulated AC or DC voltage or current and delivers it to a load. The electronic circuit monitors the converter's input and output voltages and/or currents and generates control signals for the power conversion circuit. Hence, a power converter is a power electronic system that controls the flow of electric energy based on the load demand. Power electronic systems can be found in practically all industries, as regulated DC power supplies are required for computers, telecommunications, motor drives, cars and alternative energy systems. The main purpose of a power electronic system is to deliver power to a load with maximum efficiency. Fig. 2-6 shows a typical power electronic system.

Fig. 2-6. Block diagram of a typical power electronic system [55]

As in any circuit design, the basic circuit elements in power converters are resistors, capacitors, magnetic components and transistors, as shown in Fig. 2-7. In most integrated

circuits (ICs), circuit elements such as resistors, capacitors and transistors are preferred, as they can be easily integrated into a chip. It is more difficult to include magnetic components into ICs as they are much larger in size compared with other passive elements. Also, for power converter designs, efficiency is of major concern. The transistors in switching mode converters operate either saturation or cut-off region instead of linear mode to avoid significant power dissipation in the transistors for efficiency and thermal reasons. However, by adopting switching mode in power converter design, the side effect is the generation of significant EMI. Hence, an EMI filter is an integral part of a power converter, as it is needed for harmonics and RF emissions reduction to comply with the required EMC standards [56] and therefore EMI noise source modeling has become crucial for proper EMI filter design purposes.

Fig. 2-7. Basic circuit elements in a power electronic system [57]

During the high frequency switching operation of a power converter, the DC voltage of few hundred volts is chopped by a power transistor switch in a fraction of microsecond. Thus, conducted EMI is a major issue in most power converter design. One can reduce dv/dt and di/dt by increasing the switching time but at the expense of higher losses [58]. During the power conversion process, an input voltage will be chopped in accordance with a pulse width modulation (PWM) control signal, which maintains a desired output voltage or current waveform for the load. The output voltage and/or current have

harmonics related to the switching frequency and they can be reduced by a low pass EMI filter. In power converters, passive EMI filters play a significant role in reducing the harmonics as well as high frequency conducted emissions at both the input and output of the converter [59]. Typically, a switching frequency of 75 kHz or more is quite common for higher power density and size reduction but it has a significant impact on EMI as the harmonic contents will be easily above the regulatory frequency of 150 kHz specified by EMC standards. In modern power converters, achieving both high efficiency and EMC compliance requires some tradeoff.

2.3.3 Differential Mode (DM) and Common Mode (CM) Conducted EMI

Fig. 2-8 shows a typical configuration of the flyback PFC converter powered through a line impedance stabilization network (LISN). There are several EMI coupling paths, paths 1-3 are major CM coupling paths and paths 4 and 5 are major DM coupling paths, as indicated in Fig. 2-8. The parasitic capacitances C_P , C_{PS} and C_{PG} can be found in the three CM paths. The switching voltage V_{DS} generated by the main transistor and V_{dio} caused by the secondary diode are the two main noise sources. In most cases, noise caused by V_{DS} dominates and therefore only the noise caused by MOSFET is considered in noise simulation model of this thesis.

Fig. 2-8. Major noise current paths in the flyback PFC converter

The parasitic capacitance that links the noise source to the ground constitutes the coupling path for CM noise. The parasitic capacitance depends heavily upon the circuit layout and how the power devices are mounted on the PCB. For example, the parasitic capacitance between the base plate of the device and the heat sink can be determined by:

$$C_{PE} = \frac{\epsilon_o \epsilon_r A}{d} \quad (2.1)$$

where ϵ_o is the permittivity of the air (8.85415×10^{-12}), ϵ_r is the relative permittivity of the mica, A is the area of contact, and d is the thickness of the mica layer.

For analysis of EMI problems in the power converters, there is a need to understand the source mechanisms of DM and CM noise currents. The switching action of the transistor switch causes currents with high di/dt flowing in both the input and the output circuits of the power converter, as shown in Fig. 2-9. These currents have well-defined current paths and are DM in nature.

Fig. 2-9. DM noise currents generated by the switching of MOSFET

Fig. 2-10 shows the DM noise model according to the two DM noise coupling paths as shown in Fig. 2-8. i_{DS} is the drain-source current which is a repetitive pulsed current, C_{in} is the input filter and V_{EMI} is the DM noise magnitude to be measured with the LISN.

Fig. 2-10. The DM equivalent circuit of flyback PFC converter

The switching of the power transistor also results in rapidly changing voltage across the transistor, which causes noise currents to the grounded chassis through the parasitic capacitance, as illustrated in Fig. 2-11. These currents through the grounded chassis will return through both line and neutral lines and therefore are CM in nature. Usually, the parasitic capacitance between the transistor switch and heat sink is the major contributor of the CM noise current, as the heat sink must be grounded to chassis for safety requirements [60]. Other contributors are stray capacitances between transformer primary and secondary windings, output rectifier and heat-sink, output rectifier and transformer core, etc.

Fig. 2-11. CM noise currents coupled to chassis through stray capacitance

The main source of CM noise in the PFC converter is the parasitic capacitance C_{PG} from the drain node of the PFC converter switch to the ground. Fig. 2-12 is the simplified noise propagation model for the calculation of the CM noise spectrum which shows the path of the CM noise current for the paths 1, 2 and 3 which are the major CM noise path. Due to the very small parasitic capacitance for the considered setup, the dc-blocking capacitors of the LISN and the PFC filter capacitor C_{in} can be modeled as short circuits at the frequencies of interest. For the same reason, the inductor of the LISN is assumed to be open circuit. This model is only valid for the case in which a filter capacitor is used across the input or output terminals of the rectifier.

Fig. 2-12. CM equivalent circuit of the flyback PFC converter, (a) noise from switch, (b) noise from diode

The resultant CM noise magnitude is V_{EMI} given by:

$$V_{EMI} = V_{EMI1} + V_{EMI2} \quad (2.2)$$

2.3.4 Conducted EMI Measurement

To measure conducted emissions of a power convertor, the EMC standards specify a well-defined test setup and a line impedance stabilization network (LISN), as shown in Fig. 2-13, which is usually needed. Fig. 2-13 shows that a LISN is basically a low-pass filter. The conducted emissions emitted by the power converter (connected to AC out) are terminated by a 50Ω impedance, the standard input impedance of a spectrum analyzer. For safety reason, the $1 \text{ k}\Omega$ resistor is for discharging the capacitor when the AC power is

switched off. The corner frequency of this high-pass filter is around 150 kHz. This high-pass filter has no impact on the 50 Hz or 60 Hz AC power supply, as the inductor offers very small impedance and the capacitor behaves like an open-circuit.

Fig. 2-13. LISN circuit

During the actual measurement, two identical LISN circuits are needed for the line and neutral power lines between the AC power source and the converter-under-test, as shown in Fig. 2-14 [61-63]. The conducted emissions are measured in terms of voltages across the 50 Ω termination. In all EMC standards, the noise voltage is usually expressed in dBμV, which is defined as the ratio in terms of dB with respect to 1 μV.

Fig. 2-14. Conducted emission measurement for line and neutral lines with two LISNs

In the past, conducted emissions are measured in frequency-domain with a spectrum analyzer directly. Recently, there has been reported work by measuring conducted emissions in time-domain with a digital oscilloscope and then convert them into frequency-domain using Fourier transform technique [48, 64]. The basic operating principle of a spectrum analyzer is shown in Fig. 2-15 [65, 66]. For a switching converter with a switching frequency of f_s , it generates conducted emissions at the fundamental frequency f_s and higher harmonics at $2f_s$, $3f_s$ and so on. The frequencies of these emissions will be down-converted to an intermediate frequency (IF) after going through a mixer stage. These emissions pass through an IF filter with a specified bandwidth, followed by an envelope detector and finally the quasi-peak value of the noise voltage is measured by a quasi-peak detector.

Fig. 2-15. Block diagram of a spectrum analyzer for EMI measurement [67]

Fig. 2-16 shows the measurement setup using a spectrum analyzer for conducted emissions in line and neutral lines generated by a power convertor, where CM and DM components of the conducted emissions can be discriminated by a noise separator. The relationships that govern the line and neutral emissions and the CM and DM components are given in equations (2.3) and (2.4).

$$V_{Line} = V_{CM} + V_{DM} \quad (2.3)$$

$$V_{neutral} = V_{CM} - V_{DM} \quad (2.4)$$

Fig. 2-16. Conducted EMI measurement with CM/DM noise separator

There are many different noise separators for CM and DM discrimination in frequency-domain measurement [47, 68] and Fig. 2-17 shows one of them that can be used to separate the CM and DM emissions. One of the separators is illustrated in Fig. 2-17, and the double-pole double-throw switch allows the CM and DM emissions to be obtained separately.

Fig. 2-17. CM/DM noise separator circuit [68]

Similarly, Fig. 2-18 shows the time-domain measurement setup for conducted emissions from a power converter. The line and neutral emissions can be fed into two channels of a digital storage oscilloscope (DSO) simultaneously. The adding and subtracting functions of the DSO can perform the calculations in equations (2.5) and (2.6) for CM and DM components separation, respectively.

Fig. 2-18. Block diagram of time-domain conducted EMI measurement

$$V_{CM} = (V_{Line} + V_{Neutral}) / 2 \quad (2.5)$$

$$V_{DM} = (V_{Line} - V_{Neutral}) / 2 \quad (2.6)$$

The time-domain measurement system (TDMS) uses the same LISN as shown in Fig. 2-14 but the emissions are fed into a two-channel DSO. The TDMS without a noise separator is shown in Fig. 2-19, which has an advantage of simpler setup.

Fig. 2-19. CM and DM conducted EMI measurement using DSO

Apart from the single-phase power electronic systems, there are also three-phase power electronic systems [69, 70] for such purpose. Fig. 2-20 shows a typical conducted emission measurement setup for a three-phase power electronic system. The emission voltage on each of the $50\ \Omega$ terminators are V_A, V_B and V_C , which are the vector sums of CM and DM emission voltages on each phase of the system. The CM and DM emission voltage can be obtained mathematically using equations (2.7)-(2.10).

Fig. 2-20. Conducted EMI measurement for a three-phase power electronic system

$$|V_{CM}| = \left| \frac{V_A + V_B + V_C}{3} \right| \quad (2.7)$$

$$|V_{DM_A}| = |V_A - V_{CM}| = \left| \frac{2V_A - V_B - V_C}{3} \right| \quad (2.8)$$

$$|V_{DM_B}| = |V_B - V_{CM}| = \left| \frac{2V_B - V_A - V_C}{3} \right| \quad (2.9)$$

$$|V_{DM_C}| = |V_C - V_{CM}| = \left| \frac{2V_C - V_A - V_B}{3} \right| \quad (2.10)$$

where V_{DM_A} , V_{DM_B} and V_{DM_C} are the DM voltage on the phase A, B and C separately.

The CM-DM separator introduces insertion loss and imbalance to the measured conducted emissions. By reviewing these conducted EMI measurement methods, the TDMS has the advantage of eliminating the need of a CM-DM separator for the discrimination of CM and DM conducted emissions, thereby making the measurement system more compact and cost-effective.

2.3.5 Radiated EMI from Converters

As mentioned earlier, EMI can be coupled through radiation. For example, a switching converter as shown in Fig. 2-21 has several RF current carrying loops that can be potential sources of radiated emissions. The radiated emissions from a power converter depend very much on the switching frequency of the power conversion process. Power converters with switching frequencies below 500 kHz tend to have their emissions up to several tens of MHz, which is usually not a concern for radiated emission compliance, where the regulated frequency range is above 30 MHz. However, it can be a problem for the internal circuit operation due to near-field couplings [71, 72].

Fig. 2-21. Circuit loops in a switching convertor

2.4 EMI Filters

To meet the required conducted emission limits in the EMC standards, a passive EMI filter is necessary, as shown in Fig. 2-22. The insertion of the EMI filter between the noise source and the LISN suppresses conducted emissions emitted by the power converter [73, 74].

Fig. 2-22. Passive EMI filter for conducted EMI suppression

Basically, a passive EMI filter is a low pass filter consists of shunt capacitors and series inductors and Fig. 2-23 shows some common topologies of the low pass filters. Fig. 2-24 (a) and (b) show the typical DM and CM EMI filters, respectively.

Fig. 2-23. Low pass EMI filter configuration, (a) single-stage L-C circuit, (b) π - circuit, (c) T - circuit, (d) multistage L-C circuit

Fig. 2-24. Typical passive EMI filter: (a) DM (b) CM

The active EMI filters use feedback or feed forward methods in the control topologies to suppress EMI noise [75]. Comparing to the passive EMI filters, active EMI filters have the advantage of wide switching frequency applications and more efficient to eliminate the noise, of course, at the expense of higher cost and circuit complexity.

Fig. 2-25 shows a CM active EMI filter with current canceling technique. It consists of a CM current transformer (CT) and a high performance video op-amp, which is connected to the AC input of the noise source [76]. Two output capacitors are used for the current injection to the AC line. To meet the leakage current safety specifications, the capacitance of these capacitors cannot be too large. Fig. 2-26 shows a DM active EMI filter with current canceling technique comprises a DM noise detection circuit and an amplifier that applies an amplified version of the sensed DM voltage across the driving choke.

Fig. 2-25. CM active EMI filter

Fig. 2-26. DM active EMI filter

In order to meet the stringent required conducted emission limit specified in the EMC standard, an EMI filter is an indispensable part of a power converter system for EMC compliance. Meeting the conducted emission limit with the least number of filter components is critical in the EMI filter design process for both size and cost reasons.

2.5 EMI Filters Modeling

The passive EMI filter can be evaluated by its insertion loss (IL) which can be defined as the ratio of the generator voltage and the measured voltage when the filter is inserted between the noise generator and the measuring instrument as shown in Fig. 2-27.

Fig. 2-27. Insertion loss measurement, (a) reference circuit, (b) filter inserted

The insertion loss can be generally expressed in terms of voltages as shown in equation (2.11):

$$IL(Insertion Loss) = 20 \log \left| \frac{V_1}{V_2} \right| = 20 \log \left| \frac{V_N}{V_2} \frac{R_2}{R_1 + R_2} \right| \quad (2.11)$$

where V_1 and V_2 are the measured voltages with and without filter insertion, respectively; and V_N is the input source voltage.

Parasitic elements of the EMI filter must be considered to predict filter's performance correctly and to facilitate the proper design process [77]. The EMI filter with the parasitic elements is shown in Fig. 2-28(a), DM inductor is modeled as L, EPR and EPC, and DM capacitor is modeled as C, ESR and ESL. Fig. 2-28(b) is the simplified circuit of the EMI

filter with parasitic elements included, where Z_L and Z_C are determined by equation (2.12) and (2.13), respectively.

$$Z_L = R + j\omega L + 1/(j\omega C) \quad (2.12)$$

$$Z_C = \frac{1}{1/R + 1/(j\omega L) + j\omega C} \quad (2.13)$$

Fig. 2-28. Parasitic components in an EMI filter

2.6 Proposed EMI Noise Modeling

The fast switching of the power transistors in a power converter generates significant conducted emissions in the form of harmonics of the switching frequency, which can easily extend up to several tens of MHz. The conduction modes of conducted emissions can be either common-mode (CM) or differential-mode (DM) in nature. The CM and DM conducted emissions are usually analyzed separately for identifying the dominant conduction mode of EMI. An accurate conducted EMI prediction model of a power converter before its realization is critical in the product design process, as it can identify potential EMI problems early so that they can be rectified before the final product design. Magnetic components such as inductors and transformers are critical parts in power converter design. Depending on their operating currents, these components can exhibit non-linear behaviors if they operate in the saturation region. Hence, proper modeling of these components that include the magnetic material characteristics will allow a more realistic conducted EMI with better accuracy.

Fig. 2-29. DM noise equivalent circuit with a boost inductor

As these magnetic components are in the DM current path, only DM noise source model will be discussed in details. To understand the impact of non-linearity of the magnetic components, Fig. 2-29 shows a simplified DM noise equivalent circuit of a boost converter with a boost inductor. The DM conducted emissions due to the noise source is given by:

$$V_{DM} = V_{Noise} \frac{50}{100 + Z_L} \quad (2.14)$$

The emission magnitude expressed in dB μ V across the 50 Ω termination of the LISN is:

$$V_{Noise} (dB\mu V) = 20\log(|V_{DM}| \times 10^6) = 20\log(50V_{noise}) - 20\log(100 + Z_L) + 120 \quad (2.15)$$

From (2.15), the general rule of designing the boost inductor to reduce DM emission is to choose high permeability magnetic core to achieve the necessary inductance with the least number of turns, so as to reduce its size as well as the parasitic capacitance. On the other hand, magnetic core with high permeability tends to saturate easily and hence the impedance of the boost inductor can vary significantly when its operating current is high. Therefore the non-linear behavior of the magnetic core has a direct impact on DM conducted emission.

Chapter 3 Conducted EMI Modeling

3.1 Introduction

To predict conducted EMI generated by a power converter with reasonable accuracy, the noise source, critical passive components, active components and magnetic components must be properly modeled. In addition, the influence of PCB layout, such as trace inductance, must also be well represented in the model, especially at higher frequencies. In general, there are several approaches to develop conducted EMI models for prediction purpose [11, 78, 79] and they are summarized in Table 3-1.

Table 3-1 Comparison of different EMI prediction approaches

Approach	Pros	Cons
Modular simulation approach	Efficient for functionality check	Poor accuracy
Complete circuit simulation approach	Good accuracy for linear circuit	Computational prohibitive and unable to handle nonlinearity of magnetic components
Hybrid circuit with magnetic module that including nonlinearity	Efficient and reasonable accuracy	Additional mathematical manipulation to handle magnetic nonlinearity

The most straight forward simulation approach is to treat the whole power converter as several functional modules. This approach is easy to implement and computationally efficient. However, such an approach is good for functionality check but lacks in accuracy. The second approach is to include all passive and active components of the power converter and perform a full-circuit simulation. If all the non-ideal effects, such as parasitic elements of the components are properly characterized and modeled, the simulation can be highly accurate but at the expense of prohibitive simulation time. The first two approaches usually assume all the components operate in linear region and therefore do not include the nonlinearity behaviors of magnetic components. In this thesis, the proposed prediction model adopts a hybrid approach. For examples, the control circuitry that does not affect the EMI results will be modeled as a module instead of a full

circuit. In addition, the magnetic components are modeled mathematically and interfaced with the rest of the circuits and modules seamlessly. By doing so, the proposed model achieve reasonable accuracy and simulation efficiency. Of course, it requires a comprehensive development of the nonlinear mathematical models for magnetic components, which is the major contribution of the thesis.

3.2 Noise Source Model

The power transistor in a switching mode power converter is the major cause of conducted EMI noise. The main excitation source of the noise is the repetitive pulses resulted from the high frequency switching of the power transistor, such as MOSFET. The repetitive pulses with a known switching frequency can be described by a pulsed waveform with a period T , rise time and fall time. For simplicity, the rise time t_r and fall time t_f are usually assumed to be the same, i.e. $t_r = t_f = \tau$, as shown in Fig. 3-1.

Fig. 3-1. Typical model of MOSFET switching pulse

The pulsed waveform is characterized by a duty cycle D with the period $T = 1/f_0$, where f_0 is the switching frequency of the power converter. As conducted EMI is measured in frequency-domain, repetitive pulses can be described in frequency-domain using Fourier analysis. The discrete frequencies at of the pulsed waveform occur at $n/T = n \cdot f_0$.

For ideal pulsed waveform with $t_r = t_f = 0$, the Fourier series is expressed as:

$$e(t) = 2V_{DC} \cdot D \cdot \sum_{n=1}^{\infty} \frac{\sin(n\pi \cdot D)}{n\pi \cdot D} \quad n = 1, 2, 3, \dots \quad (3.1)$$

For finite rise and fall times, where $t_r = t_f = \tau$, the expression in equation (3.1) becomes:

$$e(t) = 2V_{DC} \cdot D \sum_{n=1}^{\infty} \frac{\sin(n\pi \cdot D)}{n\pi \cdot D} \cdot \frac{\sin(n\pi \cdot \tau / T)}{n\pi \cdot \tau / T} \quad n = 1, 2, 3, \dots \quad (3.2)$$

Equation (3.2) indicates that the larger the rise/fall time, the lower amplitude of higher frequency spectral components.

Fig. 3-2. CM noise source model of an interleaved PFC converter

Fig. 3-3. DM noise source model of an interleaved PFC converter

Using an interleaved PFC converter as an example, the CM noise source can be modeled by the pulsed voltage across the MOSFET, as shown in Fig. 3-2; and the DM noise source can be described by the resultant pulsed current passes through MOSEFT, as shown in Fig. 3-3.

3.3 Passive Components model

3.3.1 Resistor

Fig. 3-4 (a) and (b) show an ideal resistor and a resistor with parasitic elements. The equivalent circuit model will include an equivalent series inductor (ESL) and an equivalent parallel capacitor (EPC). The frequency responses of the impedance for both ideal and non-ideal resistor models are shown in Fig. 3-5 (a) and (b), respectively. At lower frequencies, the resistor exhibits its ideal value and with increasing frequency, the parasitic effects due to ESL and EPC begins to kick in.

Fig. 3-4. Resistor model, (a) ideal resistor, (b) non-ideal resistor model with parasitic capacitance and inductance

Fig. 3-5. Frequency depended impedance, (a) ideal resistor, (b) non-ideal resistor

3.3.2 Capacitor

Capacitors are generally used for filtering and bypassing applications. An ideal capacitor as shown in Fig. 3-6 (a) exhibits a decreasing reactance with increasing frequency, which is governed by $X_C = 1/(2\pi fC)$, where X_C is capacitive reactance. With increasing frequency, the parasitic elements, equivalent series inductance (ESL) and equivalent

series resistance (ESR) as shown in Fig. 3-6 (b), begin to show their effects [80]. The impedance frequency responses for an ideal capacitor and non-ideal capacitor are shown in Fig. 3-7 (a) and (b), respectively.

Fig. 3-6. Capacitor model, (a) ideal model, (b) non-ideal model

Fig. 3-7. Impedance of capacitor, (a) ideal model, (b) non-ideal model

3.3.3 Inductor

Based on the measured impedance response of an inductor using an impedance analyzer, the parasitic components of the inductor can be extracted [81]. For operating frequency up to 30 MHz, a second order model as shown in Fig. 3-8 (b) is usually good enough. The equivalent circuit model includes the inductor itself and the parasitic effects are represented by equivalent parallel resistor (EPR) and the equivalent parallel capacitor (EPC). Fig. 3-9 (a) and (b) show the typical impedance versus frequency responses of an ideal inductor and an inductor with the parasitic effects included, respectively.

Fig. 3-8. Model of the inductor, (a) ideal model, (b) non-ideal model considering high frequency parasitic components

Fig. 3-9. Frequency impedance, (a) ideal impedance, (b) non-ideal impedance

3.3.4 Transformer

Fig. 3-10 shows the equivalent circuit model of a transformer with an inter-winding capacitance C_w and primary and secondary parasitic capacitances, C_p and C_s . The other parasitic components that need to be accounted for are the leakage inductances of the primary and secondary windings, $L_{\lambda P}$ and $L_{\lambda S}$. While most of the magnetic field is confined within the core, the leakage fields act as a magnetic dipoles radiating out into surrounding space and may induce disturbances on nearby components.

In many modern converters, the use of planar transformers [82-84] has become more common for higher power density packaging, stable performance from a mass production perspective and lower leakage inductances. The reduced leakage inductances are achieved by interleaving the primary and secondary layers so that the stray magnetic flux induced by the eddy currents is canceled. However, a planar transformer has higher intra-winding parasitic capacitances due to close proximity of the coils. The increase in these capacitances gives rise to higher CM noise.

Fig. 3-10. A transformer Equivalent circuit

3.4 Printed Circuit Board

The interconnecting tracks on the printed circuit board (PCB) have to be accounted for in EMI prediction model [85, 86], especially at high frequencies. For example, the equivalent circuit model of line and neutral tracks on the PCB is shown in Fig. 3-11.

Fig. 3-11. High frequency equivalent circuit of PCB tracks

At higher frequency, the current carried by the conductor tends to crowd closer to the outer periphery, a phenomena commonly known as skin effect [87]. The skin depth can be determined by:

$$\delta = \frac{1}{\sqrt{\pi f \mu_0 \sigma}} [m] \quad (3.3)$$

where μ_0 is the permeability and σ is the conductivity of the conductor ($5.96 \times 10^7 \cdot \Omega^{-1} \cdot m^{-1}$ for copper).

For a rectangular trace on PCB, the current crowds uniformly round the edges of the track and the resistance can be estimated by:

$$R_{tra} = \frac{1}{2\sigma\delta(w+h)}[\Omega] \quad (3.4)$$

where σ is the conductivity of the conductor ($\Omega^{-1} \cdot m^{-1}$) and w (m) and h (m) is the width and height of the PCB trace, respectively.

At high frequencies, the effect of inductance of a conductor is often more dominant than that of the resistance. The external inductance of a conductor due to external magnetic fields is usually referred to as the partial inductance and can in turn be divided into self and mutual inductances. The partial self-inductance comes from one segment of a wire due to external fields for a straight round wire at high frequencies can be expressed as :

$$L_{tra} = \frac{\mu_0}{2\pi} l [\ln(\frac{2l}{r}) - 1] (H / m) \quad (3.5)$$

where l (m) is the length of the wire and r (m) is the radius [88].

The mutual inductance arise due to interaction with the return current and if it needs to be considered, the geometry becomes important and the equations significantly more complex.

3.5 Active Component Model

An unwanted feature of the MOSFET structure is the presence of a body diode. To reduce the risk of turning on the parasitic BJT-transistor, the metallic source electrode that covers the n^+ region is lengthened to also cover the body region in order to obtain a short-circuited base connector on the BJT. The performance of the body diode is relatively slow compared with the MOSFET and if used in a switched converter it can cause high peak currents during diode turn off. Several measures can be taken in order to improve the performance of the diode and to make it more suitable for switched applications. This section presents the model for MOSFET [89].

As the gate-source voltage reaches above the threshold voltage, a strong inversion layer starts to form in the p -material which gives a free path for the current to flow from drain to source. The created inversion channel acts like a resistor, i.e. the drain current is proportional to the applied drain voltage. The MOSFET is said to be operating in the linear region; i.e. the drain current is proportional to the applied gate-source voltage.

When the applied voltage at the drain is so large that the inversion layer at the drain end becomes zero ($V_{DS} = V_{GS} - V_{th}$), the so called pinch-off point has been reached. The electron concentration at the drain end of the channel area becomes very low since only a depletion region will exist. Note that the existence of only a depletion region is no barrier to electron flow; the electric field pulls electrons into the drain. If $V_{DS} = V_{GS} - V_{th}$, the MOSFET is operating at the onset of saturation and if the drain-source voltage is increased even further, the MOSFET will operate in the saturated region.

The switching behavior of a power MOSFET is strongly determined by the internal parasitic capacitances that are formed due to internal build-up of charges. These parasitic capacitances are shown in Fig. 3-12 whose values vary strongly with the applied voltages. They depend on different physical properties and can be divided into three major parts [90].

The gate-source capacitance C_{GS} mainly constitutes of two parts: the first part C_{pp} is formed by the capacitive coupling between the gate electrode and the source electrode. This capacitance is geometry dependent and does not vary with the applied voltages. The second part is voltage dependent and can in turn be divided into three separate subdivisions. The first subdivision is the major contributor and consists of the gate oxide capacitance in the channel region ($C_{channel}$). The second subdivision is the diffusion capacitance in the n^+ -region which originates from the gradient in the carrier concentration and depends on the applied voltage. If a change in the junction current is desired, the stored minority carriers close to the space charge region must be removed, hence the name diffusion capacitance. The third subdivision is the spread of the space

charge region in the drift region. All constituents are connected in parallel which results in an internal capacitance that does not only vary with the applied gate-source voltage but also with the resulting drain-source voltage.

Fig. 3-12. Internal capacitances in a vertically diffused MOSFET [91]

The drain-source capacitance (C_{DS}) is independent of the gate voltage, but is dependent on the applied drain-source voltage. It originates from the depletion layer capacitance in the drain-source $p-n$ -junction and can be divided into two parts; the depletion capacitance (C_j) and the diffusion capacitance (C_d). However, in the internal MOSFET $p-n$ -junction, the diffusion capacitance (C_d) is negligible; hence the drain source capacitance is mainly made up of the depletion capacitance (C_j).

The gate-drain capacitance (C_{GD}) is perhaps the most important parameter that determines the switching characteristics of a MOSFET. Simplified, it can be seen as the field oxide capacitance ($C_{field-oxide}$) connected in series with the depletion capacitance ($C_{depletion}$) in the drift region. The depletion capacitance shows a strong voltage dependence since the drain-gate voltage varies strongly. As long as the MOSFET is turned off, the voltage across gate-drain is high which gives a large depletion layer and

consequently a low gate-drain capacitance ($C_{GD} \approx C_{depletion}$). As the MOSFET is turned on and operating in the linear region, the voltage across it is low, an accumulation layer is present under the gate electrode. The gate-drain capacitance is then dominated by the field oxide capacitance ($C_{GD} \approx C_{field-oxid}$).

Fig. 3-13 shows a typical MOSFET model which can be applied in the power electronic converters simulation based on the above description [92, 93].

Fig. 3-13. Switch model

3.6 Non-linear Behavior of Magnetic Component

3.6.1 Conventional models

An inductor or a transformer is a magnetic component that stores magnetic energy when a current passing through it [94] and its ability to store magnetic energy is quantified by its inductance. Typically, a magnetic component is wound with many turns and the higher the number of turns, the larger the inductance. Magnetic components are commonly used in power converters for their abilities to store energy as well as act as filters for harmonics suppression [95]. The magnetic component can be either without core (air-core) or with a ferromagnetic core.

Usually, core material with higher permeability is preferred as it increases the magnetic induction and thereby provides higher inductance. At high frequencies, soft ferrites are widely used as magnetic cores, as they incur much lower losses at high frequencies compared to ordinary iron alloys. Most of magnetic components are constructed using

enamel coated wire wrapped around a ferromagnetic core with wire exposed on the outside, while some enclose the wire completely and are referred to as "shielded".

Ferromagnetic-core magnetic components use iron as a magnetic core to increase the inductance [96]. Another kind of magnetic material commonly used to construct inductors is ferrite where the main applications are in the frequencies of MHz range. A magnetic core can increase the inductance of a coil by a factor of several thousand due to its higher permeability. However, the magnetic properties of the core material also cause several side effects, which require special design attention. If the current through a ferromagnetic cored coil is high enough that the inductance will not remain constant anymore but will change with the current through the coil [97], which leads to nonlinearity and results in distortion of the signal. For example, audio signals can suffer inter-modulation distortion due to saturated inductors. To prevent nonlinearity, the currents through these inductors must be limited so that the core will not be saturated.

There are many attempts to construct mathematical models that can accurately analysis experimental data for magnetic materials in the past. However, only several of these have been proven to be feasible. These model are the Preisach model [98], the Stoner-Wohlfarth model [99], and the Globus model. Before the chosen model can be used in circuit-base analysis, the merits and constraints of each of these models need to be understood, as each of them is only capable of giving an approximate description for the magnetic materials under certain conditions. To be able to model the EMI noise accurately in power electronic circuits, accurate modeling of the inductive components with magnetic cores is necessary [100].

Preisach proposed a model to describe the hysteresis based on switching mechanisms in 1935. In This model, it assumes that the material is consisted of an assembly of elemental volumes that each has a characteristic pair of forward or reverse switching field. A sample of iron, for example, may have randomly distributed magnetic domains, resulting in a net magnetic field of zero. Fig. 3-14 shows a rectangular hysteresis loop [101] that can be used to represent the local magnetization curve of elemental volume. The classical

Preisach model assumes that the ferromagnetic material is constructed by a large number of switching volumes. Each elementary volume is a rectangular hysteresis loop consists of two parameters, the critical field h_c and the bias field (or offset field) h_o .

$$h_A = h_o + h_c, h_B = h_o - h_c \text{ [A/m]} \quad (3.6)$$

where h_A and h_B are switching field parameters.

Fig. 3-14. Magnetization curve described by Preisach

3.6.2 Piecewise linear magnetization model

Consider an inductor or transformer with an inductance of L_{nor} is designed to operate at a nominal voltage and a nominal frequency. When the voltage exceeds its nominal value, the inductor begins to saturate and its inductance reduces to L_{sat} . The nonlinear flux-current characteristic is shown in Fig. 3-15, where the scales of flux and current are expressed in terms of per unit quantity. The nominal voltage and current are chosen as base values for the per-unit system.

Fig. 3-15. Flux-current characteristic of the nonlinear inductance

The current in the inductor i is a nonlinear function of flux linkage ψ that, in turn, is a function of voltage across the inductor v . They are related by:

$$v = L \frac{di}{dt} = \frac{d\psi}{dt}, \quad \psi = \int v \cdot dt \quad (3.7)$$

$$i = \frac{\psi}{L(\psi)} \quad (3.8)$$

Fig. 3-16. Model of a nonlinear inductor

With the above equations, an inductive element with nonlinear inductance can be modeled as a current source, where the current i is a nonlinear function of voltage v , as shown in Fig. 3-16. In this model, the relationship between i and v must be known first so that it can be implemented in a circuit-based simulation tool.

For ease of circuit analysis, the characteristic of the magnetic core are usually considered to be linear or piece-wise linear, as shown in Fig. 3-17. In Fig. 3-17 (a), the hysteresis and saturation are ignored, while in Fig. 3-17 (b), the saturation is included. The linearized relationship between B and H is given by (3.9):

$$B = \begin{cases} B_{sat} & \text{for } H \geq B_{sat} / \mu \\ \mu H & \text{for } |H| < B_{sat} / \mu \\ -B_{sat} & \text{for } H \leq -B_{sat} / \mu \end{cases} \quad (3.9)$$

Fig. 3-17 Approximation of the B-H characteristics of a magnetic core material, (a) hysteresis and saturation ignored, (b) hysteresis ignored

3.6.3 Proposed model

In order to simply scale up the prediction of models that are based on one or two domains, one needs to develop equations that can represent the behavior of the materials. These models use statistical thermodynamic principles to describe the magnetization behavior, which includes a large number of magnetic domains [102]. The Langevin-Weiss model which considers an array of magnetic moments in thermal equilibrium at a particular temperature [103] is a classical approach for magnetism on this scale.

The Jiles-Atherton model was developed based on the Langevin-Weiss approach [104] and it was based on statistical mechanisms. It works well for material with low

anisotropy for which the main mechanism is domain boundary movement. But For highly anisotropic materials, a simple analytic anhysteretic equation [105] cannot be developed for it, then the mathematical approximation becomes less realistic.

1. Anhysteretic magnetization description

The extension of the Langevin-Weiss theory describes the ferromagnetic materials by incorporating a coupling among magnetic moments acting as a strong magnetic field to align the magnetic moments in a domain parallel to each other. A mean field α is introduced to quantify this coupling, which is proportional to the bulk magnetization M . so that the effective field is $H_e = H + \alpha M$, which includes coupling to the magnetization can be used to replace the classical magnetic field H .

According to the energy dissipation law, the energy supplied to the materials is equal to the change of magneto-static energy and it is called anhysteretic magnetization. Equation (3.10) represents the anhysteretic function of a magnetic material.

$$M_{an} = M_s L\left(\frac{\mu_0 m (H + \alpha M)}{k_B T}\right) [A / m] \quad (3.10)$$

where k_B is the Boltzmann constant, T is the absolute temperature, and $L(x)$ is mainly depended on the material anisotropy.

For the isotropic material, the anhysteretic function is:

$$M_{an} = M_s [\coth(x) - 1/x] \quad (3.11)$$

Equation (3.12) is the anhysteresis function for materials exhibiting axial anisotropy.

$$M_{an} = M_s \tanh(x) \quad (3.12)$$

2. Extension to hysteresis

From the anhysteretic magnetization description, it is possible to develop an analytical model of hysteresis by considering the energy dissipation mechanism. The energy loss is assumed to be proportional to the change in magnetization, and this situation can happen under both domain wall motion and rotation. Thus, the reversible and irreversible magnetization can be derived in analytical formulations and then the two components can be combined to give an equation that describes the total magnetization M .

The differential irreversible magnetization can be described as below:

$$\frac{dM_{irr}}{dH} = \frac{M_{an} - M_{irr}}{k \cdot \delta - \alpha(M_{an} - M_{irr})} \quad (3.13)$$

where δ is +1 when H increases in the positive direction and -1 when H increases in the negative direction.

In the process of magnetization, the reversible magnetization M_{rev} results from the reversible domain wall bowing, translation or rotation. It is proportional to the difference between M_{an} and M_{irr} with a domain wall bowing parameter (reversibility coefficient) c , which represents the fraction of reversible magnetization changes and ranges from 0 to 1. Then the total magnetization M is the sum of irreversible and reversible magnetizations:

$$M = M_{rev} + M_{irr} = (1-c)M_{irr} + cM_{an} \quad (3.14)$$

The total differential magnetization which includes both irreversible and reversible components of magnetization, is then derived as below,

$$\frac{dM}{dH} = (1-c) \cdot \frac{M_{an} - M}{\delta \cdot k \cdot (1-c) - \alpha \cdot (M_{an} - M)} + c \cdot \frac{dM_{an}}{dH} \quad (3.15)$$

The solutions of these equations with a suitable anhysteretic function can describe a typical hysteresis loop. In these equations, the anhysteretic and hysteretic descriptions represent the components of magnetization along the magnetic field direction. The total

resultant magnetization represents the component of magnetization paralleling to the applied magnetic field direction [106].

A magnetic core with high permeability can increase the inductance of a coil by a factor of hundreds or thousands times. However, the magnetic properties of the core material also caused several other undesirable effects that require special design attention. If the current through a ferromagnetic cored coil is high enough, the inductance will no longer be constant but change with the current through the coil, which leads to nonlinearity and results in distortion of the signal. In this condition, it is necessary to find a method to describe this phenomenon accurately.

Table 3-2 summarizes the various possible magnetization models of the magnetic core. Preisach model can be employed to describe the nonlinear magnetization mechanism accurately but is more complicated to implement and integrate with circuit simulation model. On the other hand, the piecewise linear model is easy to be implemented in the circuit analysis but unable to handle the nonlinearity of the magnetization. The Jiles-Atherton (J-A) model is chosen for its accurate nonlinear magnetization description as well as ease of implementation in the circuit analysis. However, it requires additional steps to extract the parameters for simulation purpose, which is the focus of the next chapter.

Table 3-2 Magnetic component models comparison

Model	Pros	Cons
Preisach	Accurate nonlinear magnetization description	Difficult to implement in circuit simulation
Piecewise linear	Easy to implement in circuit simulation	Unable to handle nonlinear magnetization property
Jiles-Atherton	Accurate magnetization description and ease of implementation in circuit simulation	Additional steps to extract nonlinear magnetization model parameters

Chapter 4 Magnetic Material Modeling

4.1 Introduction

Accurate circuit modeling plays an important role as it shortens the product design and development process. For proper EMI modeling of power converter to predict conducted emissions, accurate modeling that includes dynamic operating condition is critical, especially for circuit components with nonlinear behaviors. Power converters contain several such non-linear components, such as ferromagnetic cored inductors and transformers. The nonlinearity characteristics of these components are useful especially for examination of the impact of dynamic circuit phenomena on EMI.

Inductors are realized by winding coils around a magnetically permeable core. Besides the resistance of the wire and the parasitic capacitance caused by the winding technique, the overall performance of an inductor also depends on the characteristic of the magnetic core material.

Several models have been adopted to describe the nonlinear behavior of magnetic material. Base on the literature review in Chapter 3, the Jiles-Atherton (J-A) model has been found to be most suitable for modeling ferrite-core based inductors and transformers. The J-A model is a group of first-order equations describing the hysteresis mechanism in ferromagnetic and ferri-magnetic materials.

It is based on the existing ideas of domain wall motion including both bending and translation. It can be used to model the dynamics of a magnetic cored component that can exhibit saturation and hysteresis behaviors. As a time-domain and history-dependent model, the J-A model is ideal for time-domain based circuit simulation because the group of first-order equations can be incorporated into other integro-differential equations describing other parts of a circuit. Hence, the dynamics of the circuit can be analyzed by solving a combination of all the equations so that the impact of nonlinearity of these magnetic components on EMI can be analyzed with ease.

J-A model is described by several first-order differential equations and they can be easily incorporated into other first-order equations that describe the behavior of a complete circuit with inductive elements. The fundamental requirement to achieve accurate circuit modeling is to extract the parameters of J-A model accurately. This chapter provides an introduction of J-A model and explains the extraction of the parameters for circuit simulation purposes. Then the application of J-A model in the power converter circuits is demonstrated.

4.2 Review of magnetic material modeling

4.2.1 J-A model

For a magnetic core as shown in Fig. 4-1, the magneto-motive force (MMF) in a close path of length l_m is equal to the total current passing through the center of the core. The current is the source of MMF and equation (4.1) relates the current to the MMF [57].

$$F(t) = H(t) \cdot l_m = i(t) \quad (4.1)$$

Fig. 4-1. Magnetic field in a close path

The relationship between flux density B and magnetic field H , or equivalently Φ and F , is determined by the magnetic core material characteristics. Fig. 4-2 (a) shows the B - H curve of free space with a slope μ_0 , which is the permeability of free space and equal to $4\pi \cdot 10^{-7}$ A/m. Fig. 4-2 (b) shows the B - H curve of a magnetic core excited by a sinusoidal magnetic field, where the relationship is highly nonlinear with a hysteresis loop.

Fig. 4-2. B - H characteristics, (a) free space or air, (b) typical magnetic material

To enable the implementation of nonlinearity of an inductive component in a circuit-based simulation tool, it is necessary to express an inductive component with magnetic core in terms of its current and voltage. Fig. 4-3 summarizes the relationship between the current and voltage of an inductive component. The winding voltage $v(t)$ is related to flux and flux density as described in equation (4.2) and the current in the winding $i(t)$ is related to magnetic field. The relationship between flux and magnetic field is dependent on the core material characteristics.

Fig. 4-3. Procedure to determination of a magnetic core's i - v characteristics

$$v(t) = \frac{d\psi(t)}{dt} = \frac{A_c \cdot dB(t)}{dt} \quad (4.2)$$

where A_c is the core area.

As mentioned earlier, the relationship between magnetic field and magnetization is assumed to be linear for ease of analysis. In reality, such relationship is nonlinear and the J-A model is required to handle this relationship. J-A model consists of a group of first-

order ordinary differential equations, which can be solved numerically to provide the relationship between the magnetic core's magnetization and magnetic field so that its nonlinear characteristics and magnetic hysteresis can be properly modeled.

Basically, it describes the B - H curve using five parameters. The magnetization is composed of two terms in the J-A model of hysteresis, an irreversible and a reversible component, given by:

$$M = M_{irr} + M_{rev} \quad (4.3)$$

The irreversible component M_{irr} represents the irreversible domain wall motion and the reversible magnetization M_{rev} corresponds to the reversible domain wall bending.

The relationship between M_{irr} and the magnetic field H is governed by:

$$\frac{dM_{irr}}{dH} = \frac{M_{an}(H_e) - M_{irr}}{k \cdot \text{sig}\left(\frac{dH}{dt}\right) - \alpha [M_{an}(H_e) - M_{irr}]} \quad (4.4)$$

$$\frac{dM_{rev}}{dH} = c \left(\frac{dM_{an}}{dH} - \frac{dM_{irr}}{dH} \right) \quad (4.5)$$

where M_{an} is the anhysteretic magnetization, k and c are the wall motion parameters, the function $\text{sig}(\cdot)$ is defined as $\text{sig}(x) = 1$ for $x \geq 0$ and $\text{sig}(x) = -1$ for $x < 0$. M is the total magnetization, which is the summation of M_{rev} and M_{irr} [24].

H_e is the effective magnetic field:

$$H_e = H + \alpha \cdot M \quad (4.6)$$

where H is the magnetic field in the core and α is an inter-domain coupling factor.

The flux density B is given by:

$$B = \psi / A_c = \mu_0(H + M) \quad (4.7)$$

where ψ is the flux which is the integration of the voltage, B is the flux density and μ_0 is the permeability of free space.

The anhysteretic M_{an} magnetization is described by the Langevin function:

$$M_{an} = M_s \left(\coth \frac{H_e}{\partial} - \frac{\partial}{H_e} \right) \quad (4.8)$$

$$\frac{dM_{an}}{dH} = \frac{M_s}{\partial} \left[1 - \coth^2 \left(\frac{H_e}{\partial} \right) + \left(\frac{\partial}{H_e} \right)^2 \right] \quad (4.9)$$

where M_s is saturation magnetic moment of the core material and ∂ is a shape parameter.

From equations (4.4) - (4.9), one can derive:

$$M_{rev} = c(M_{an} - M_{irr}) \quad (4.10)$$

$$\frac{dM}{dH} = (1 - c) \cdot \frac{M_{an} - M_{irr}}{\text{sig}(\dot{H}) \times k - \alpha(M_{an} - M_{irr})} + c \cdot \frac{dM_{an}}{dH} \quad (4.11)$$

From the viewpoint of numerical solutions for hysteresis modeling, firstly is to solve equation (4.4) for the irreversible magnetization and then adds the reversible magnetization by solving equation (4.5). However, when the magnetic field is reduced from the extremity of the loop such as loop tip, if the irreversible magnetization M_{irr} is smaller than the anhysteretic magnetization M_{an} in the first quadrant or larger than the M_{an} in the third quadrant, direct solution of equation (4.4) leads to a negative differential susceptibility as the field is reduced. Under these conditions, the domain walls actually

remain pinned on the defect sites, so $dM_{irr}/dH = 0$. However, the reversible magnetization is still operative so that a bulged domain wall will relax as the field reduces. It means that as the field is reduced from the loop tip, until the magnetization crossed the anhysteretic, the change in magnetization is almost reversible, therefore $dM/dH = dM_{rev}/dH$ in this region and equation (4.11) has to be modified in the form of [26]:

$$\frac{dM}{dH} = (1-c)\delta \frac{M_{an} - M}{\text{sig}(\dot{H})k(1-c) - \alpha(M_{an} - M)} + c \cdot \frac{dM_{an}}{dH} \quad (4.12)$$

where

$$\delta = \begin{cases} 0, & \text{if } \text{sig}(\dot{H}) \times (M_{an} - M) \leq 0 \\ 1, & \text{otherwise} \end{cases} \quad (4.13)$$

Fig. 4-4 shows the curve obtained from the J-A model. It describes the relationship between magnetic field and magnetization, which is related to flux or flux density. In equation (4.13), the coefficient δ is introduced. This is to take into account the fact that for the segments ab and cd , as shown in Fig. 4-4, all magnetization is reversible as $\delta = 0$ for these two segments and $M_{irr} = 0$ and $M_{rev} = c \cdot M_{an}$.

Fig. 4-4. Solid: Magnetization against magnetic field; Dotted: Anhysteretic magnetization against magnetic field

4.2.2 Parameters to Establish J-A Model

1) Review on Differential Evolution Method

Based on the description in the previous section, there are five parameters M_s, ∂, α, k and c need to be extracted before application. This section will describe how to extract these five parameters based on measured data using differential evolution (DE) method. Below is a brief review of DE method. Fig. 4-5 shows the mutation of different generations of parameters.

Fig. 4-5. Mutation of parameters using DE

The equation for mutation is given by:

$$v_{i,g} = x_{r0,g} + F \cdot (x_{r1,g} - x_{r2,g}) \quad (4.14)$$

where the scale factor $F \in (0, 1+)$, $x_{r0,g}$ is the base vector, and $x_{r1,g}$ and $x_{r2,g}$ are randomly selected for every mutation. During the optimization process, the difference between the mutation vectors and the base vector becomes smaller, which means the members of the population are getting closer to the optimal solution [107-109].

When applying the DE method to solve a problem, a cost function is needed. In this chapter, the primary purpose is to find the five parameters of the J-A model from a measured hysteresis loop. It can be achieved by using the minimization of the summation

of root of square of the difference between simulated and measured magnetic field as indicated by:

$$M_1 = \sum_{j=1}^{OneLoop} \sqrt{(H_{j,M} - H_{j,S})^2} \quad (4.15)$$

One can also target at minimizing the following product:

$$M_2 = \left[\sum_{j=1}^{OneLoop} \sqrt{(H_{j,M} - H_{j,S})^2} \right] \cdot \left[\sum_{j=1}^{OneLoop} \sqrt{(B_{j,M} - B_{j,S})^2} \right] \quad (4.16)$$

where, $H_{j,M}$, $H_{j,S}$, $B_{j,M}$ and $B_{j,S}$ are the measured magnetic field, modeled magnetic field, measured flux density and modeled flux density at each point within the hysteresis loop, respectively.

Fig. 4-6. Influence of parameter k on hysteresis loop (Solid: $k=44$; Dotted: $k=22$)

Fig. 4-7. Influence of parameter M_s on hysteresis loop (Solid: $M_s = 4 \times 10^5$; Dotted: $M_s = 5 \times 10^5$)

To observe how the five parameters influence the shape of the hysteresis loop modeled by the J-A model, one parameter is changed at a time with other four parameters remain unchanged.

Fig. 4-6 shows the influence of k and it can be clearly observed that k is equal to the difference of the positive and negative magnetic fields or the coercive forces on the horizontal axis when the hysteresis loop crosses horizontal axis. Fig. 4-7 illustrates the influence of saturation M_s . Based on these two figures, one can see that it is important to identify a narrow range for both parameters M_s and k using the measured saturated hysteresis loop at low frequency so that the other three parameters in the J-A model can be determined with ease.

Although the initial values of the parameters do not really affect the process of DE optimization process, the boundaries for parameters are important. Without proper boundaries, the optimal solution can be out of searching range of the algorithm. Moreover, the searching process may take a very long time or even get stuck in some local optimum solutions that are too far away from the global optimal solution. Hence, it is essential to identify the boundaries of critical parameters. The saturation parameter M_s and parameter k can be determined by a saturated hysteresis loop at low frequency so that the ranges for them are properly determined. To set the ranges for the remaining three

parameters in the J-A model, one has to refer to the manufacturers' data sheet for each type of magnetic material under study.

In this chapter, the ranges of the five parameters in J-A model for the magnetic core (RS Stock No. 232-9561) are listed below:

- $\partial = 2 \sim 50$;
- $\alpha = 4 \times 10^{-5} \sim 8 \times 10^{-5}$;
- $c = 0.1 \sim 0.8$;
- $k = 2 \sim 10$;
- $M_s = 4 \times 10^5 \sim 5 \times 10^5$.

The strategy to develop the B-H hysteresis loop is as follows:

- 1) Extract the five parameters in the J-A model using measured data; and
- 2) Use the five extracted parameters to predict inrush current and validate with experiment results.

By following steps 1 and 2, all the five parameters are obtained as follows:

- ∂ : 20.32;
- α : 4.002×10^{-5} ;
- c : 0.7997;
- k : 2.085;
- M_s : 4.9787×10^5 ;

Fig. 4-8 compares the modeled and measured B - H curves at 10 kHz, where the solid curve is the modeled hysteresis loop based on the five extracted parameters using DE method and the dotted curve is the measured hysteresis loop. In general, both the measured and modeled B - H curves agree well, which shows the J-A model has the ability to model the nonlinearity of the magnetic material with reasonable accuracy. The slight difference between the modeled hysteresis loop and the measured one could be due to the

assumption of uniform magnetic field in the toroidal core but in reality the magnetic field near the outer radius of the toroidal core is always less than that near its inner radius.

Fig. 4-8. Comparison of B-H curves at 10 kHz (Solid: modeling by J-A model with extracted five parameters; Dotted: measured)

2) Inverse J-A Model Parameters with Datasheet

The hysteresis parameters M_s , k , ∂ , α and c can be obtained from the following procedure [27, 110, 111]. The wall motion parameter k can be estimated as H_c , where H_c is the coercive force in A/m when $M = 0$.

Fig. 4-9. Hysteresis curve

Relationship between wall motion parameter c and shape parameter ∂ can be derived

from:

$$\chi_{in} = \left. \frac{dM}{dH} \right|_{(H=0, M=0)} = \frac{(1-c)M_{an}}{k(1-c) - \alpha M_{an}} + c \frac{dM_{an}}{dH} \quad (4.17)$$

where χ_{in} is the initial differential susceptibility with $H = 0$ and $M = 0$. By limiting $H \rightarrow 0$ into series expansion of equation (4.8), the expression becomes:

$$\begin{aligned} \lim_{H \rightarrow 0} M_{an} &= \lim_{H \rightarrow 0} M_s \left(\coth \frac{H}{\partial} - \frac{\partial}{H} \right) \\ &= \lim_{H \rightarrow 0} M_s \left(\frac{\partial}{H} + \frac{1}{3} \frac{H}{\partial} - \frac{1}{45} \left(\frac{H}{\partial} \right)^3 + \dots - \frac{\partial}{H} \right) = \lim_{H \rightarrow 0} M_s \frac{1}{3} \frac{H}{\partial} \end{aligned} \quad (4.18)$$

Thus, equation (4.17) can be rewritten as:

$$\begin{aligned} \chi_{in} &= \left. \frac{dM}{dH} \right|_{(H=0, M=0)} = \frac{(1-c)M_{an}}{k(1-c) - \alpha M_{an}} + c \frac{dM_{an}}{dH} \\ &= 0 + c \frac{dM_{an}}{dH} = c \frac{d \left(M_s \frac{1}{3} \frac{H}{\partial} \right)}{dH} = c \frac{M_s}{3\partial} \end{aligned} \quad (4.19)$$

$$c = 3\partial \chi_{in} / M_s \quad (4.20)$$

From equations (4.8) and (4.9) lead to:

$$M_{an}(M_r) = M_s \left(\coth \frac{\alpha M_r}{\partial} - \frac{\partial}{\alpha M_r} \right) \quad (4.21)$$

$$\left. \frac{dM_{an}}{dH} \right|_{M=M_r} = \frac{M_s}{\partial} \left[1 - \coth^2 \left(\frac{\alpha M_r}{\partial} \right) + \left(\frac{\partial}{\alpha M_r} \right)^2 \right] \quad (4.22)$$

$$\chi_r = \left. \frac{dM}{dH} \right|_{M_r, (H=0, \delta=-1)} = (1-c) \frac{M_{an}(M_r) - M_r}{-k(1-c) - \alpha [M_{an}(M_r) - M_r]} + c \left. \frac{dM_{an}}{dH} \right|_{M_r} \quad (4.23)$$

$$M_r = M_{an}(M_r) + \frac{k}{\frac{\alpha}{1-c} + 1 / (\chi_r - c \frac{dM_{an}(M_r)}{dH})} \quad (4.24)$$

where $M_r = B_r / \mu_0$ and remanence flux density B_r and saturation magnetization M_s can be obtained from the magnetic material specification. χ_r is the remanence differential susceptibility and is defined as the slope of the magnetization loop at the remanent point (when $M = M_r$) with $H = 0$. The secant method introduced in [112, 113] can be used to calculate the parameter α . Assuming that:

$$f(\alpha) = M_{an}(M_r) - M_r + \frac{k}{\frac{\alpha}{1-c} + 1 / (\chi_r - c \frac{dM_{an}(M_r)}{dH})} \quad (4.25)$$

$$\alpha_n = \alpha_{n-1} - \frac{\alpha_{n-1} - \alpha_{n-2}}{f(\alpha_{n-1}) - f(\alpha_{n-2})} f(\alpha_{n-1}) \quad (4.26)$$

$$\text{If } \left| \frac{\alpha_n - \alpha_{n-1}}{\alpha_n} \right| < \text{tolerable error, } \alpha = \alpha_n.$$

It is useful to include some redundancy by incorporating the coordinates of the loop tip M_m , H_m and the slope of the initial magnetization curve at the loop tip χ_m . Considering the differential susceptibility along the initial magnetization curve at the loop tip with $\delta = 1$ and the loop tip is sufficiently close to saturation, then the differential susceptibility of the initial magnetization curve at the loop tip will approach the differential susceptibility of the anhysteretic $dM / dH = dM_{an} / dH$. This can be used as an approximation to obtain an equation relating the hysteresis parameters. Using the general result $M_{rev} = c(M_{an} - M_{irr})$, it can be observed that the above approximation also implies that $dM_{irr} / dH = dM / dH = dM_{an}(H_m) / dH$.

From equations (4.6) and (4.12):

$$\chi_m = \left. \frac{dM}{dH} \right|_{(M_m, H_m)} = \frac{M_{an}(M_m) - M_m}{k(1-c) - \alpha[M_{an}(M_m) - M_m]} \quad (4.27)$$

$$M_m = M_{an}(M_m) - \frac{(1-c)k\chi_m}{\alpha\chi_m + 1} \quad (4.28)$$

where M_m is the maximum value of magnetization at tip of hysteresis loop, χ_m is the differential susceptibility at the tip of hysteresis loop that approaches saturation .

Using the secant method as introduced in equation (4.26) to solve equation (4.29), the shape parameter ∂ can be obtained from:

$$g(\partial) = M_{an}(M_m) - M_m - \frac{(1-c)k\chi_m}{\alpha\chi_m + 1} \quad (4.29)$$

With the above-mentioned procedures, all the five J-A model parameters can be calculated.

With the extraction of five model parameters, which is described in the above procedures, the J-A model is scalable for different conditions. Unlike other models, the J-A model is established by a group of first-order equations describing the hysteresis mechanism in ferromagnetic and ferri-magnetic materials. Because of its history-dependence characteristic, the J-A model is ideal for time-domain circuit simulation, as the group of first-order equations can be easily incorporated into other integro-differential equations describing other parts of a circuit.

4.3 Modeling Isolation Transformer in DC-DC Converter

To validate the J-A model for circuit simulation purpose, the inrush current phenomenon in a DC-DC converter with isolating transformer will be studied. The isolation transformer uses J-A method to model its magnetic core. Any load change will result in rapid change of switching duty-cycle and the magnetic core will experience a sudden change of flux density, which causes a large inrush current. Although the operating point

of the magnetic core is still well below the saturation point, it will be shown that that by introducing an intermediate duty-cycle stage, such kind of inrush current can be significantly reduced.

A distributed power supply point of load (POL) method has been widely adopted in practice in large-scale computer system and a typical DC-DC converter for such kind of system is shown in Fig. 4-10. It uses a full-bridge converter module with a synchronous rectifier. It was reported that a large inrush current is expected due to fast response to dynamic load change [114].

Fig. 4-10. Circuit under study

For simplicity, many power electronics circuits use linearized model for inductive elements with magnetic cores. In the linearized model, the hysteresis and magnetic core losses are not considered. To model the dynamic response and transient phenomena, such as inrush current, accurate hysteresis of the magnetic core is indispensable and the J-A model described in this chapter will be useful.

4.3.1 Simulation of Inrush Current

The inrush current phenomenon is analyzed using the circuit shown in Fig. 4-11, where the load current is not considered.

Fig. 4-11. Circuit used in modeling

In the J-A model of the transformer shown in Fig. 4-11, the resistor R is the resistive loss of the winding. In the simulation, $R = 8.85\Omega$. The transformer core is toroidal in shape (average diameter $D = 2.4 \times 10^{-2}$ m and cross-sectional area $A = 4.54 \times 10^{-5}$ m²). The number of turn of the winding is $n = 40$. The magnetic material has the following hysteresis parameters: $M_s = 2.821 \times 10^5$, $k = 30$ (A/m), $\partial = 27$ (A/m), $\alpha = 5 \times 10^{-5}$ and $c = 0.55$. The DC source voltage is 20 V and the system switching frequency is 10 kHz.

The voltage output at the switch of S_1 through S_2 is:

$$v_{tr}(t) = v_R(t) + v_L(t) \quad (4.30)$$

where $v_{tr}(t)$ is the voltage output at the switches of S_1 through S_4 ; $v_R(t)$ and $v_L(t)$ are voltages across the resistor R and the transformer, respectively.

According to the Ampère's law:

$$H\pi D = ni \quad (4.31)$$

$$v_L(t) = \mu_0 n A \frac{d}{dt} (H + M) = \frac{\mu_0 n^2 A}{\pi D} \cdot \frac{di(t)}{dt} \cdot \left[1 + \frac{dM}{dH} \right] \quad (4.32)$$

From equations (4.30) and (4.32):

$$\frac{di}{dt} = \frac{\pi D}{\mu_0 n^2 A} \cdot \frac{v_{tr}(t) - iR}{1 + dM / dH} \quad (4.33)$$

From equations (4.32) and (4.33),

$$\frac{dH(t)}{dt} = \frac{1}{\mu_0 n A} \cdot \frac{v_{tr}(t) - iR}{1 + dM / dH} \quad (4.34)$$

Equation (4.34) can be numerically discretized as follows:

$$\frac{H(j+1) - H(j)}{\Delta t} = \frac{1}{\mu_0 n A} \cdot \frac{v_{tr}(j+1) - i(j) \times R}{1 + (dM / dH)(j)} \quad (4.35)$$

where $j=1,2,3,\dots$ is the numerical simulation step.

Based on equation (4.30), integrating both sides of $v_{tr}(t) - iR = \mu_0 n A \frac{d}{dt} [H(t) + M(t)]$ results in:

$$M(t) = \int_{t_0}^t \frac{v_{tr}(t) - iR}{\mu_0 n A} dt - H(t) + M_0 + H_0 \quad (4.36)$$

where M_0 and H_0 are the initial magnetization and magnetic field values, respectively; and t_0 is the starting moment of modeling and $H(t) = ni / (\pi D)$.

Equation (4.36) can be also be numerically discretized as:

$$M(j+1) = \sum_{m=1}^j \frac{v_{tr}(m) - i(m) \times R}{\mu_0 n A} \times \Delta t - H(j) + M_0 + H_0 \quad (4.37)$$

The numerical solution can be obtained by the following steps:

1. Initialize current, magnetic field and magnetization;

2. Calculate the effective magnetic field H_e , anhysteretic magnetization M_{an} and dM_{an}/dH using equations (4.6), (4.8) and (4.9); respectively;
3. Determine the sign of dH/dt with $H(j) - H(j-1)$ where $j = 1, 2, 3, \dots$ is the simulation step;
4. Determine the sign of $M_{an}(j) - M(j)$ and calculate the value of δ using equation (4.13);
5. Calculate dM/dH using equation (4.12) (if it is first step of iteration, $dM/dH = 0$);
6. Calculate $H(j+1)$ using equation (4.35) and update the value of current $i(j+1)$ using equation (4.31);
7. Compute magnetization at $j+1$ step using equation (4.37);
8. Repeat steps 2 through 7 until the iteration ends.

4.3.2 Numerical Results of Inrush Current

Scenario 1: large inrush current due to sudden change of duty ratio from low to high

This scenario emulates a very large inrush current due to the rapid change of duty ratio from 0.1 to 0.4 in order to adapt the sudden change of load. The duty ratio of each positive voltage and negative voltage half cycle is 0.1 or 0.4.

Fig. 4-12 show the time-domain simulated voltage $V_{tr}(t)$ and Fig. 4-13 shows the zoom in voltage waveform. The change of duty ratio occurs at $t = 4 \times 10^{-3}$ s. Fig. 4-14 shows the simulated current waveform, where the change of duty ratio and a large inrush current occurs at $t = 4 \times 10^{-3}$ s. The inrush current is almost 1.2 A right after the moment of change of duty ratio and the eventual peak value of the current is about 0.12 A.

Fig. 4-15 shows the anhysteretic magnetization versus time. Fig. 4-16 and Fig. 4-17 show magnetization and magnetic field versus time, respectively. It is clearly observed from Fig. 4-14 that the large inrush current results in a large surge of magnetic field right after the change of duty ratio at $t = 4 \times 10^{-3}$ s.

Fig. 4-12. Voltage $v_r(t)$ against time t

Fig. 4-13. Zoomed-in voltage $v_r(t)$ against time t

Fig. 4-14. Current against time t

Fig. 4-15. Anhysteretic magnetization against time t

Fig. 4-16. Magnetization against time t

Fig. 4-17. Magnetic field against time t

Fig. 4-18. Magnetization against magnetic field from the beginning to the end of simulation

Fig. 4-19. Comparison of flux density against magnetic field in the steady state for each stage

Fig. 4-20. Comparison of magnetization against magnetic field in the steady state for each stage

Fig. 4-18 shows the magnetization against magnetic field from the start to the end of the simulation. Fig. 4-19 and Fig. 4-20 show the steady-state flux density and magnetization against magnetic field, before and after the change of duty ratio, respectively. All these simulation results show that the J-A model of the isolation transformer works correctly in the time-domain simulation.

Scenario 2: Inserting an intermediate stage to reduce inrush current

From the earlier example, there is a very large inrush current due to sudden change of duty ratio. To reduce the surge in inrush current, an intermediate stage with a duty ratio of 0.2 between 0.1 and 0.4 is introduced. This is to check if J-A model response to such change accordingly. The time-domain simulated voltage $V_{ir}(t)$ after inserting an intermediate stage is shown in Fig. 4-21 to Fig. 4-23. Fig. 4-24 shows the time-domain current and the relationship between magnetic field and magnetization can be found in Fig. 4-25 to Fig. 4-30.

Fig. 4-21. Voltage $v_{ir}(t)$ against time t

Fig. 4-22. Zoomed-in voltage $v_r(t)$ against time t : 1

Fig. 4-23. Zoomed-in voltage $v_r(t)$ against time t : 2

Fig. 4-24. Current against time t

Fig. 4-25. Anhyseretic magnetization against time t

Fig. 4-26. Magnetization against time t

Fig. 4-27. Magnetic field against time t

Fig. 4-28. Magnetization against magnetic field from the beginning to the end of simulation

Fig. 4-29. Comparison of flux density against magnetic field in the steady state for each stage

Fig. 4-30. Comparison of magnetization against magnetic field in the steady state for each stage

The simulations have shown that the inrush current phenomenon due to rapid change of switching duty ratio in a DC-DC converter with the insulating transformer based on J-A model provide useful results for analysis purposes. It is found that the inrush current is ten times higher than the normal current when there is a sudden change of duty ratio from 0.1 to 0.4. It is also found that by inserting an intermediate stage, the inrush current can be effectively reduced.

4.4 Modeling of Inductor in Resonant Tank Circuit

In the earlier example, the J-A model has been successfully applied to DC-DC converter to predict the inrush current due to sudden change in load. In this section, the J-A model is further extended to simulate the dynamics of a resonant inverter for driving a nonlinear load, a fluorescent lamp.

The inverter circuit is driven by a high frequency square wave and a semi-theoretical fluorescent lamp model developed from HID lamp model is adopted in the simulation. The simulated time-domain voltage across the lamp and current through it are simulated and compared to the measured results reported in [115].

The full bridge inverter connected to a parallel-loaded resonant tank is the most common topology used as high-frequency inverter for electronic ballast. There are four different type resonant tanks [116] and the LC topology as shown in Fig. 4-31 is adopted here. The circuit comprises an inductor in series with a capacitor and fluorescent lamp; and is driven by a 10 kHz square wave. Again, the J-A model is chosen for modeling the inductor.

Fig. 4-31. The circuit model

4.4.1 Fluorescent Lamp Model

In general, there are two types of fluorescent lamp models, the physical [117] and semi-theoretical models. Comparing with the physical model, semi-theoretical model is developed based on the curve-fitting process and easier to be incorporated into circuit simulation computation. There are five physical equations developed from the HID lamp model [118], which can be used to model the V - I characteristic of a fluorescent lamp.

In a low-pressure discharge and positive column, the electron temperature can be described by:

$$\frac{dT_e}{dt} = a_1(i^2 R - P_{elastic} - P_{rad}) \quad (4.38)$$

where T_e is the electrode temperature, t is the time, a_1 is a model constant, $i^2 R$ is the input power, $P_{elastic}$ is the elastic collision loss and P_{rad} is the UV radiation loss.

The total radiation loss can be expressed by the following exponential function of the electron temperature:

$$P_{rad} = a_2 \exp(-ea_3 / kT_e) \quad (4.39)$$

where both a_2 and a_3 are adjustable model constants, e is the charge on an electron and k_B is the Boltzmann constant.

The elastic collision loss is proportional to the electron energy and this energy loss is a linear function of electron temperature given by:

$$P_{elastic} = a_4(T_e - T_g) \quad (4.40)$$

where a_4 is an adjustable constant and T_g is gas temperature, which is 350K in the simulation.

The electrical conductivity of the discharge is proportional to the ionization rate. In a fluorescent lamp, the dominant ionization process is called “two-stage ionization” [119]. The mathematical expression of ionization rate for the two-stage ionization contains both algebraic and exponential functions of electron temperature. Therefore, the lamp resistance value can be expressed by the following empirical expression:

$$R = a_5 T_e^{-3/4} \exp(ea_6 / 2k_B T_e) \quad (4.41)$$

where a_5 and a_6 are adjustable model parameters.

Finally, the lamp current can be computed by solving the ballast circuit equation:

$$v(t) = i \times R + v_{ele} \quad (4.42)$$

where $v(t)$ is the input voltage of the fluorescent lamp, R is the lamp resistance and v_{ele} is the electrode voltage drop, which is 10V for the simulation.

The proposed lamp model is for a lamp. With the help of a set of measured lamp current and voltage, a Genetic Algorithm [120] can be developed to obtain the adjustable model constants of the lamp, as given in Table 4-1.

Table 4-1 Adjustable model constant for the lamp

a_1	a_2	a_3	a_4	a_5	a_6
38019	829000	6.45	0.055	3966.03	0.33

4.4.2 Dynamics of Resonant Tank with the Fluorescent Lamp

In the following simulation, the J-A model for the inductor and non-linear model of fluorescent lamp described earlier are adopted. For the inductor, the core is in toroidal shape and its average diameter $D = 3 \times 10^{-2}$ m, its cross-sectional area $A = 60 \times 10^{-5} m^2$ and the number of turns of the winding $n = 10$. The voltage source is a square wave with peak voltage of $\pm 110V$ and frequency $f = 10kHz$.

Applying Kirchhoff's voltage law:

$$V_0 = V_L + V_{FL} \quad (4.43)$$

where V_0 is the source voltage and V_{FL} is the fluorescent lamp voltage, which is also equal to the capacitor voltage V_C and V_L is the inductor voltage.

Equation (4.38) can be discretized as:

$$\frac{T(j+1) - T(j)}{\Delta t} = a_1 (i(j))^2 \cdot R(j) - P_{con}(j) - P_{rad}(j) \quad (4.44)$$

where $j = 1, 2, 3, \dots$ is the step of numerical simulation.

According to equation (4.41), $R(j+1)$ can be calculated.

The capacitor voltage can be expressed as:

$$\frac{dV_c}{dt} = \frac{i_c}{C} = \frac{i - i_{FL}}{C} = \frac{i - V_c / R}{C} \quad (4.45)$$

where i, i_c, i_{FL} are the inductor, capacitor and fluorescent lamp currents respectively.

Equation (4.45) can be discretized as follows:

$$\frac{V_c(j+1) - V_c(j)}{\Delta t} = \frac{i_c(j)}{C} = \frac{i(j) - V_c(j) / R(j)}{C} \quad (4.46)$$

Applying Ampère's law:

$$H = \frac{ni}{\pi D} \quad (4.47)$$

$$V_L = \mu_0 n A \frac{d}{dt} (H + M) = \frac{\mu_0 n^2 A}{\pi D} \cdot \frac{di}{dt} \cdot \left[1 + \frac{dM}{dH} \right] \quad (4.48)$$

From equations (4.47) and (4.48), one gets:

$$\frac{di}{dt} = \frac{\pi D}{\mu_0 n^2 A} \cdot \frac{V_0 - i_{FL} R}{1 + dM / dH} = \frac{\pi D}{\mu_0 n^2 A} \cdot \frac{V_0 - V_c}{1 + dM / dH} \quad (4.49)$$

Hence,

$$\frac{dH}{dt} = \frac{1}{\mu_0 n A} \cdot \frac{V_0 - i_{FL} R}{1 + dM / dH} = \frac{1}{\mu_0 n A} \cdot \frac{V_0 - V_c}{1 + dM / dH} \quad (4.50)$$

which can be discretized as:

$$\frac{H(j+1) - H(j)}{\Delta t} = \frac{1}{\mu_0 n A} \cdot \frac{V_0(j) - V_c(j)}{1 + (dM / dH)(j)} \quad (4.51)$$

Integrating both sides of $V_L = V_0 - V_c = \mu_0 n A \frac{d}{dt} (H + M)$ leads to:

$$M(t) = \int_{t_0}^t \frac{V_0 - V_c}{\mu_0 n A} dt - H(t) + M_0 \quad (4.52)$$

where M_0 is the initial magnetization and t_0 is the start moment of the numerical simulation.

Equation (4.54) can be discretized by:

$$M(j+1) = \sum_{m=1}^j \frac{V_0(m) - V_c(m)}{\mu_0 n A} \times \Delta t - H(j) + M_0 \quad (4.53)$$

The numerical simulation can be described by taking the following steps:

1. Initialize magnetization, current, electron temperature and capacitor voltage. Set the initial value of $dM / dH = 0$;
2. Calculate the effective magnetic field H_e , anhysteretic magnetization M_{an} and dM_{an} / dH using equations (4.8) and (4.9);
3. Calculate the fluorescent lamp resistance R using equation (4.41);
4. Compute the electron temperature $T(j+1)$ using equation (4.44) and then update the fluorescent lamp resistance $R(j+1)$ using equation (4.41);
5. Evaluate the capacitor voltage $V_c(j+1)$ using equations (4.45) and (4.46);
6. Determine the value of $M(j+1)$ using equation (4.53);
7. Calculate the sign of dH / dt by using $H(j) - H(j-1)$.
8. Determine the sign of $M_{an}(j) - M(j)$ and then calculate the values of δ and dM / dH using equation (4.12);
9. Compute $H(j+1)$ according to equation (4.51) and then update the current $i(j+1)$ according to equation (4.47);
10. Repeat steps 2 through 9 until iteration stops.

Fig. 4-32 shows the simulated time domain waveforms of the current, voltage, magnetic field and magnetization of the inductor. Fig. 4-33 show the simulated hysteresis loops of the resonant inductor at power and switching frequencies based on the J-A model, respectively. Both the main loop at power frequency and the minor loop at the switching frequency indicate that the J-A model works well at both low and high frequencies.

Fig. 4-32. Current, Voltage, magnetic field and magnetization against time

Fig. 4-33. Magnetization against magnetic field

Fig. 4-34 shows the zoom-in view of the inductor current and voltage under steady-state condition. Fig. 4-35 shows the dynamic change and eventual steady state of lamp resistance and lamp electron temperature.

Fig. 4-34. Zoom in inductor current and voltage

Fig. 4-35. Lamp resistance and electro temperature

Fig. 4-36 (a) and (c) show the dynamic lamp current and voltage versus time. It can be observed that the fluorescent lamp is excited by the high voltage at the beginning and then it settles into the steady-state. Fig. 4-36 (b) and (d) show the zoom-in view of lamp current and voltage versus time under steady-state condition. Fig. 4-37 compares the inductor, lamp and capacitor currents versus time under steady-state condition, where the dotted line is the inductor current and solid line is the lamp current.

Fig. 4-36. Dynamic lamp current and voltage

Fig. 4-37. (a) The comparison of inductor and lamp currents against time, (b) capacitor current against time

The simulation results of the resonant inverter circuit with fluorescent lamp has again shown that the J-A model of hysteresis can be easily incorporated into the circuit that

describes the voltage across the lamp and the current through all the components in both transient and steady states.

4.5 Summary

This chapter describes in details the implementation of J-A model by extracting the necessary parameters based on measured data. Two circuit examples, the DC-DC converter circuit and the inverter electronic ballast circuit have demonstrated the ability of J-A model to provide realistic simulation results in both transient and steady states of operation. It gives us the confidence to apply J-A model in power electronic circuit modeling for both harmonics and conducted EMI analyses. In the next chapter, the impact of nonlinearity of inductive component on EMI will be evaluated using the J-A model.

Chapter 5 Modeling Conducted EMI for Interleaved PFC Boost Converter

5.1 Introduction

PFC converters have become highly popular as they produce much lower power frequency harmonics in the lower frequency spectrum. However, the high frequency switching of the converters continue to generate RF conducted emission to the power mains above 150 kHz [47, 121]. In order to comply with the required conducted EMI standards, accurate EMI noise prediction is important, as it facilitate the correct design of EMI filter for EMI suppression purposes.

For simplicity, the inductance of the boost inductor in power electronic circuit is usually assumed to be constant without its core hysteresis and losses taken into consideration. In reality, the actual inductance varies with current and frequency. There are some methods to describe the nonlinear inductance in the power electronic circuits [3] but none can be used to for predicting the behavior of in the situation a complex system.

In this chapter, a conducted EMI model for an interleaved power factor correction (PFC) converter that takes into considering the nonlinearity of boost inductors is developed and analyzed. The J-A model described in Chapter 4 will be adopted to model the interleaved boost inductors so that their nonlinearity can be included in the circuit simulation. Once the conducted EMI noise model is accurately modeled, an EMI filter will be designed and installed between the input AC power mains and the PFC converter. Finally, experimental results are collected for validation purposes.

This Chapter is organized into six sections. Section 5.2 introduces the interleaved PFC boost converter configuration and the modeling of circuit components modeling. Section 5.3 develops the conducted EMI model of the PFC converter. Section 5.4 shows the prediction results of conducted emission and the experiment results are given in Section 5.5. Finally, Section 5.6 concludes this chapter.

5.2 Conducted EMI Model of PFC Boost Converter

5.2.1 PFC Boost Converter Topology

There are mainly three types of PFC boost converter topologies depending on its boost inductor current. They are continues conduction mode (CCM) [122, 123], discontinues conduction mode (DCM) [124, 125] and boundary conduction mode (BCM) [126, 127]. CCM has less switching losses but requires more complicated control topology, while BCM and DCM has a simple control topology but generate lager current ripples. As technology develops, the CCM topology becomes easier to implement for its obvious advantage.

Fig. 5-1 Interleaved PFC boost converter

Fig. 5-1 shows a typical interleaved PFC boost converter. The diode bridge which consists of four PB66 (D1 to D4) diodes rectifies the AC voltage and the inductors L(L1 and L2), MOSFET switches (S1 and S2) and the diodes (D5 and D6) configure an interleaved boost converter. The capacitor C added at the output reduces the voltage ripple. The MOSFET power switch turns on or off at a fixed frequency. The boost inductor current ($i_L = i_{L1} = i_{L2}$) and the duty cycle in every switching period d are related by the following equations [128].

$$L \frac{di_L}{dt} = V_{in-rec}, 0 \leq t < dT_s \quad (5.1)$$

$$L \frac{di_L}{dt} = V_{in-rec} - V_o, dT_s \leq t < T_s \quad (5.2)$$

where V_{in-rec} is the rectified DC input voltage, V_o is the output voltage and T_s is the switching period. Fig. 5-2 shows the average equivalent circuit model of the interleaved PFC converter, the circuit can be modeled with two differential equations that include the boost inductor current and output capacitor given below:

$$L \frac{di_L}{dt} = V_{in-rec} - V_o(1-d) \quad (5.3)$$

$$C \frac{dV_o}{dt} = 2i_L(1-d) - \frac{V_o}{R} \quad (5.4)$$

where $(1-d)$ is the off time of the power switch.

From equation (5.3), the duty d can be derived as follows:

$$d = \frac{t_{on}}{t_{on} + t_{off}} = 1 - \frac{V_{in-rec}}{V_o} = 1 - \frac{V_{in-rec}}{V_o} + \frac{L}{V_o} \frac{dI_L}{dt} \quad (5.5)$$

where t_{on} and t_{off} are the on and off times of the MOFET switch, respectively.

Fig. 5-2. Average equivalent circuit of the interleaved PFC converter

For an interleaved CCM PFC converter, the average current control method is adopted for the PFC controller, as it is less sensitive to the switching losses as compared to peak current or hysteresis current control method [129, 130]. Fig. 5-3 shows the implementation of the control block. There are two control loops in the controller, a slower response ‘outer loop’ that controls average DC output voltage and the faster inner

loop shapes the input current. The two average boost inductor currents are controlled so that they are proportional to the utility line-voltage waveform and the summation of the two interleaved inductor currents is the system input current. The two controllers generate the PWM duty ratios for the two power switches.

Fig. 5-3. Average current control configuration of the PFC converter

Fig. 5-4. Switching signals and boost inductor currents of the interleaving topology

For the boost average current control method, the average output voltage V_o is higher than the input peak voltage at any time. The input current drawn from the AC line at any instant must be proportional to the line voltage, which is controlled by the inner loop of the controller. The boost inductors L_1 and L_2 are interleaved in the circuit and the control signal of the switch S_1 is delayed half switching period from another switch S_2 , as shown in Fig. 5-4. It can be found that this topology has the advantage of reducing current ripple but incurs higher cost due to additional auxiliary circuits and power devices

[131].

5.2.2 Nonlinear Boost Inductor Modeling

For a given output power, higher current ripple reduces the size of the boost inductor but increases the core losses. Of course a lower current ripple requires a larger boost inductor size but has an easier design process. Hence, the boost inductor design is a trade-off between size and ripple performance. Using low-loss core magnetic materials such as Ferrites, Molypermalloy Powder, Ferrous Alloy Powder or Iron based Amorphous can reduce the core losses significantly.

In the PFC converter, the input current variation is mainly dependent on the input voltage, output power and boost inductors [132]. In order to predict the dynamics of the PFC converter more precisely, the J-A model has been chosen for the boost inductor hysteresis behavior simulation. An accurate representation for the boost inductor results in more accurate input current prediction, which leads to more accurate conducted EMI prediction.

The J-A hysteresis model with the five extracted parameters mentioned in Chapter 4 can be used to model the boost inductor nonlinear behavior accurately. The hysteresis parameters for the magnetic core material which is CTX16-17309 are given in Table 5-1.

Table 5-1 Adjustable J-A model constant

M_s (A / m)	k (A / m)	$\hat{\phi}$ (A / m)	α	c
1.436×10^6	138	174	5×10^{-5}	0.55

5.3 Conducted EMI Modeling of the PFC Converter

5.3.1 Modeling of Components

1) Boost Inductor

Based on the measured impedance versus frequency response of the boost inductor using a HP 4194A impedance analyzer, the parasitic components of the inductor can be

extracted. Fig. 5-5 shows the equivalent circuit model of the boost inductor with its equivalent parallel resistance (EPR) and equivalent parallel capacitance (EPC). Fig. 5-6 shows the frequency response of the boost inductor from 100 kHz to 30 MHz.

Fig. 5-5. Equivalent circuit model of the boost inductor with parasitic components

Fig. 5-6. Impedance frequency response of the boost inductor

2) DC Filter Capacitor

Based on the measured impedance versus frequency response of input DC filter capacitor using a HP 4194A impedance analyzer, the parasitic components of the input filter are extracted. Fig. 5-7 shows the equivalent circuit model of the input filter with its equivalent series resistance (ESR) and equivalent series inductance (ESL). Fig. 5-8 shows the frequency response of the input filter capacitor from 100 kHz to 30 MHz.

Fig. 5-7. Equivalent circuit model of the input filter with parasitic components

Fig. 5-8. Impedance frequency response of the input filter

5.3.2 DM Conducted EMI

The interleaved PFC converter topology to be implemented is shown in Fig. 5-9. The main source of the DM conducted EMI noise is very much dependent on the time-domain input current of the PFC converter. The input current of the PFC converter is dependent on the output power of the converter, inductance of the boost inductor and input voltage of the mains. Although a cycle by cycle method can be used to determine the PFC converter's input current waveform for a half period, some form of modulation method is employed in the actual circuit and the current passing through the boost inductor depends on the modulation function. Therefore, a simulation model including the controller should be included in the input current calculation.

The nonlinearity of the two boost inductors, L_1 and L_2 , are modeled with the hysteresis behavior described in details in the next section. In the power electronic circuits, high frequency switching of the power semiconductor is the major cause of conducted EMI [133, 134]. For the non-idealities of the power transistor switch (IR840), drain-to-source turn-on resistance ($R_d = 0.8 \Omega$), equivalent series inductances of the drain and source terminals ($L_d = 3.5 \text{ nH}$ and $L_s = 7.5 \text{ nH}$) and junction capacitances ($C_{gs} = 185 \text{ pF}$, $C_{gd} = 185 \text{ pF}$ and $C_{ds} = 200 \text{ pF}$) are obtained from the power transistor datasheet [135] and

included in the simulation model for more realistic prediction of i_{sum} , the summation of the two boost inductor currents. Similarly, the equivalent model of the boost inductor with the parasitic components, as shown in Fig. 5-5, is also included in the simulation model as shown in Fig. 5-9.

Fig. 5-9. Configuration of the simulated PFC converter system

As the PFC converter to be designed is a two-wire system (line and neutral wires without earth wire), differential-mode (DM) conducted EMI will be dominated. Fig. 5-10 shows the high-frequency DM noise current path, which superimposes on the power frequency input current of the PFC converter. The AC power line current contains the power frequency component due to AC-DC rectifier and a superimposed triangular waveform due to high frequency switching. The noise current spectrum in the frequency-domain

will depend on how accurate the input current waveform can be simulated. The conducted emissions will be measured by the LISN.

Fig. 5-10. DM current paths in the interleaved PFC converter

Fig. 5-11. Topology of an LISN

As the actual test setup uses the EMCO 3721/2M LISN for conducted emission measurement, the circuit model of the LISN, as shown in Fig. 5-11, is also included in the conducted emission prediction model. In the actual measurement, the DUT side is connected to the interleaved PFC converter and conducted emissions on the line and neutral lines are measured with channel 1 (CH1) and channel 2 (CH2) of the digital oscilloscope, respectively.

Once the time-domain conducted emissions emitted from the PFC converter at the line (L) and neutral (N) outputs of the LISN are simulated, the CM and DM conducted emissions are obtained as follows:

$$V_{CM} = (V_{Line} + V_{Neutral}) / 2 \quad (5.6)$$

$$V_{DM} = (V_{Line} - V_{Neutral}) / 2 \quad (5.7)$$

For the measurement, the actual emissions from the line and neutral of the LISN are fed into a digital storage oscilloscope (DSO). The adding and subtracting functions of the DSO can be used to implement equations (5.6) and (5.7) to measure the CM and DM conducted emissions, respectively. The frequency-domain spectrum of the time-domain measurement can be obtained via Fourier Transform function, as given in equations (5.8) and (5.9). The use of DSO has an advantage of measuring CM and DM conducted emissions without a CM-DM separator [136], so that the measurement results will not be influenced by the separator circuit.

$$X(k) = \sum_{j=1}^N x(j) w_N^{(j-1)(k-1)} \quad (5.8)$$

$$w_N = e^{(-2\pi i)/N} \quad (5.9)$$

where X returns the discrete Fourier transform (DFT) of vector x , computed with a fast Fourier transform (FFT) algorithm.

Fig. 5-12. DM noise propagation model

Once the noise current i_{in} is obtained from the circuit simulation as shown in Fig. 5-9, the

DM conducted emission can be predicted using the DM equivalent circuit model shown in Fig. 5-12. The noise current through the LISN can be calculated and conducted emission noise voltage V_{EMI} can be determined readily.

As two channels of the PFC converter have the same design and switching frequency, the impedances of the two boost inductors are the same. The two source voltages are shown in Fig. 5-13, where V_1 and V_2 are the harmonic components of V_{s1} and V_{s2} .

Fig. 5-13. Time-domain waveform of the source voltages, (a) Fundamental component, (b) 2^{nd} - order component

The fundamental component of the two noise sources will cancel each other. So the noise at the fundamental switching frequency in the 2-channel PFC converter is very small. For the 2^{nd} order harmonic component of noise sources, they are in phase and add up together and therefore, the frequency of $2f_s$ emission dominates. For the same reason, all the odd harmonics cancel each other and even order harmonics stay the same as those in the non-interleaving topology. Therefore, the DM conducted emission analysis and the subsequent EMI filter design are based on the even order harmonics for the interleaved topology.

5.4 Simulation Results

5.4.1 Simulation Set-Up

There are many circuit simulation tools available in the market, for examples, Candence and ADS. These are commonly used electronic design automation (EDA) tools. Candence is popular for integrated circuit and printed circuit board (PCB) designs. It also has many IPs covering a broad range of standard circuits; for examples, interfaces, memory, analog, data processing unit, etc. ADS is well established for RF circuit designs for mobile, wireless and high-speed data link applications. These two tools are versatile for RF, microwave and high speed digital designs, and they can also be extended for power converter design. However, as the main focus of the thesis is to integrate the control circuit and mathematical J-A model for EMI prediction purpose, direct simulation using these existing EDA tools can be cumbersome. Hence, MATLAB platform is adopted for ease of implementation of the mathematical functions of the J-A model.

The interleaved PFC converter topology illustrated in Fig. 5-9 is integrated with the nonlinear boost inductor model and the complete interleaved PFC converter system is simulated with MATLAB Simulink. The simulation was conducted based of the PFC converter specifications given in Table 5-2. The calculated boost inductor value is $200\mu H$ if its nonlinear behavior is ignored [137].

Table 5-2 PFC converter specifications and parameters

Specifications	
Input voltage (rms)	85-265V
Input frequency	50Hz
Output voltage	390V
Switching frequency	116.5kHz
Output power	195W

The simulated input current with the nonlinear behavior of the boost inductor included are implemented based on equations (5.10)-(5.16). Once the simulated time-domain input current is obtained, the DM conducted emissions can be determined with the models shown in Fig. 5-12. The frequency-domain conducted emission spectrum can be derived in accordance with equations (5.8) and (5.9). With the know conducted emission spectrum, the required attenuation of the EMI filter can be estimated by comparing the

emission spectrum and the conducted emission limit of the respective EMC standard.

According to the Ampère's circuital law $\pi DH = ni$, the relationship between current and magnetic field is given by,

$$H\pi D = ni \quad (5.10)$$

Then

$$V_L = \mu_0 n A \frac{d}{dt} (H + M) = \frac{\mu_0 n^2 A}{\pi D} \cdot \frac{di}{dt} \cdot \left[1 + \frac{dM}{dH} \right] \quad (5.11)$$

Integrating both side of equation (5.11) results in,

$$M(t) = \int_{t_0}^t \frac{V_L}{\mu_0 n A} dt - H(t) + M_0 + H_0 \quad (5.12)$$

where M_0 is the initial magnetization, t_0 is the start moment of the integration, and V_L denotes the boost inductor voltage respectively. Equation (5.12) can be discretized as below:

$$M(j) = \sum_{m=1}^j \frac{V_L(m)}{\mu_0 n A} \times \Delta t - H(j) + M_0 + H_0 \quad (5.13)$$

Form equations (5.10) and (5.11),

$$\frac{di}{dt} = \frac{\pi D}{\mu_0 n^2 A} \cdot \frac{V_L}{1 + dM / dH} = \frac{\pi D}{\mu_0 n^2 A} \cdot \frac{V_L}{1 + dM / dH} \quad (5.14)$$

Hence

$$\frac{dH}{dt} = \frac{1}{\mu_0 n A} \cdot \frac{V_L}{1 + dM / dH} = \frac{1}{\mu_0 n A} \cdot \frac{V_L}{1 + dM / dH} \quad (5.15)$$

where i is the boost inductor current.

Equation (5.15) can be discretized as

$$\frac{H(j+1) - H(j)}{\Delta t} = \frac{1}{\mu_0 n A} \cdot \frac{V_L(j)}{1 + (dM/dH)(j)} \quad (5.16)$$

Then the nonlinear boost inductor model can be built in the Embedded MATLAB Function of SIMULINK [68] with the following six numerical procedures,

- 1) Initialize magnetization, current and set the initial value of dM/dH as zero.
- 2) Calculate the effective magnetic field H_e , anhysteretic magnetization M_{an} and dM_{an}/dH using equations (4.7), (4.9) and (4.10).
- 3) The boost inductor voltage $V_L(m)$ is captured from the overall circuit simulation at each time step, the value of $M(j)$ can then be calculated using equation (5.13).
- 4) Calculate the sign of dH/dt using $H(j) - H(j-1)$, determine the sign of $M_{an}(j) - M(j)$ and value of δ , and calculate the value of dM/dH using equation (4.13) at last.
- 5) Calculate the $H(j+1)$ using equation (5.16). Update the current $i(j+1)$ using equation (5.10).
- 6) Repeat steps 2 through 5 in the Embedded MATLAB Function until simulation stops.

5.4.2 Dynamics of Power Converter

The modeling of nonlinearity of the boost inductor for PFC converter has been described in the earlier section. In addition to the nonlinearity behavior of the inductor, the parasitic elements of boost inductor and capacitor are also taken into account in the simulation model to ensure good prediction accuracy. The impedance responses of the boost inductor and filter capacitor are measured with their parasitic effects extracted and modeled. Fig. 5-5 shows the equivalent circuit model of the boost inductor with its equivalent parallel resistance (EPR) and equivalent parallel capacitance (EPC) and Fig.

5-7 shows the equivalent circuit model of the input filter with its equivalent series resistance (ESR) and equivalent series inductance (ESL).

Fig. 5-14 shows the simulated hysteresis curve of the boost inductor based on the J-A hysteresis model. The main loop in power frequency and the minor loop in the switching frequency show that the J-A model could be applied in the magnetic model at both low and high frequency appropriately.

Fig. 5-14. hysteresis behavior of boost inductor

Fig. 5-15 shows the simulated steady state output voltage and input current waveforms of the PFC converter with the boost inductors' nonlinear behavior included, which clearly demonstrates that the J-A model governed by a group of complex differential equations works seamlessly with the various circuit blocks of the PFC converter.

From the observed waveforms as shown in Fig. 5-16, it can be found that the ripple currents in the two inductors cancel each other due to the interleaving topology, and result in much smaller ripple in the total input current. Thus the interleaved topology has an ability to reduce the inductor magnetic core size compared with the traditional single boost inductor topology.

Fig. 5-17 (a) and (b) show the boost inductor currents without and with J-A model included. Fig. 5-17 (c) shows the zoom in current comparison when the boost inductor nonlinearity is ignored and considered. It is clearly observed that the current ripple in Fig. 5-17 (b) is smaller than that shown in Fig. 5-17 (a), a clear indication that J-A model handles the boost inductor nonlinearity very well in the simulation. The accurate prediction of the input current including the inductor's nonlinearity will lead to more accurate conducted EMI prediction for the proper selection EMI filter, which will be discussed in the following section.

Fig. 5-15. PFC converter input voltage, current, and output voltage, $V_{in-rms} = 110V$

Fig. 5-16. Individual boost inductor currents and input current with J-A model consideration,

$$V_{in_rms} = 110V$$

Fig. 5-17. Boost inductor current, (a) current without J-A model consideration, (b) current with J-A model consideration, (c) zoom in boost inductor current comparison

5.5 Experiment Results

To validate the conducted emission results from the prediction model described earlier, a PFC converter is designed in accordance with the specification given in Table 5-2. It is designed with a UCC28258/UCC28220 dual interleaved PFC pre-regulator and two boost inductors with toroid core CTX16-17309. The core has a diameter D of $2.3 \times 10^{-2} m$, a cross-sectional area A of $1 \times 10^{-4} m^2$, and the number of turns n is 100. These geometrical parameters of the magnetic core are applied in the J-A model to describe the nonlinear behavior of the boost inductors.

To observe both the time-domain and frequency-domain conducted EMI, a time-domain measurement setup (TDMS) using digital signal oscilloscope (DSO) is used for collecting the conducted EMI measurement results [48]. The TDMS is consists of a 3725/2M LISN and a TDS7054 oscilloscope as shown in Fig. 5-18. The frequency-domain conducted emission can be obtained using FFT function easily. One channel of the DSO is connected to line (L) output and another channel of the same DSO is connected to neutral (N) output of the line impedance stabilization network (LISN). By using the adding or subtracting function of the DSO, the common-mode (CM) or differential-mode (DM) components of the conducted EMI can be measured separately. A DM EMI filter will be designed and implemented between the input power and interleaved PFC converter to bring the emission level below the limit.

Fig. 5-18. Set-up of the conducted noise emission measurement

5.5.1 Noise Spectrum Comparison

Fig. 5-19 (a) shows the measured conducted emission from 100 kHz to 30 MHz. Fig. 5-19 (b) shows the comparison between simulated conducted emissions without J-A model and measured conducted emission. Fig. 5-19 (c) shows the comparison between simulated conducted emission with J-A model and measured conducted emission. For ease of comparison, Table 5-3 indicates the measured and simulated conducted emission of the first few harmonics of the switching frequency.

Fig. 5-19. Conducted emission, (a) measurement result, (b) comparison with simulation result with J-A model, (c) comparison with simulation result without J-A model

Table 5-3 Conducted emission comparison

Frequency (kHz)	Measured magnitude (dBμV)	Simulated magnitude without J-A model (dBμV)	Simulated magnitude with J-A model (dBμV)
116.5	94	90	95
233	102	96	104
349.5	70.7	66	71
466	77	72	77
582.5	65	62	67

It clearly indicates that the simulated emission without the boost inductors nonlinearity consideration under-estimates the emission by 4 to 6 dB. The comparisons clearly show that with J-A model included, the simulated results agree very well with the measured results. Without the J-A model, the nonlinearity of the boost inductor will not be included in the simulation model, which leads to under-estimation of conducted emission and incorrect selection of EMI filter to meet the respective EMI limit. However, at higher frequencies, the parasitic capacitance of the boost inductor begins to dominate and the simulated emissions with and without J-A model do not show much difference.

Fig. 5-20 shows the common mode (CM) measured conducted spectrum from 100 kHz to 30 MHz, as it is a two-wire power electronic system, the CM emission is much lower than the DM emission. Thus only DM EMI filter is needed for the PFC converter.

Fig. 5-20. Conducted common mode (CM) emission

5.5.2 EMI Filter Design

As the interleaved PFC converter is AC powered by two wires (line and neutral) without ground wire, the noise analysis will be mainly on DM EMI noise in the PFC converter with a DM filter as shown Fig. 5-21. For ease of design, the filter inductors $L_{D1} = L_{D2}$ and capacitors $C_{X1} = C_{X2}$. The values of these filter components are mainly dependent on the required conducted EMI attenuation and the corner frequency of the filter. Based on the corner-frequency method [138], where the corner frequency is determined by a

40dB/decade slope line which is tangential to the required attenuation, as shown in Fig. 5-22.

Fig. 5-21. Typical DM EMI filter topology

Fig. 5-22. Attenuation and corner frequency for filter design

For the interleaved PFC converter, which operates at a switching frequency $f_s=116.5$ kHz, the emission is higher at its 2nd harmonic 233 kHz, as explained in Section 5.3. With predicted emission level at 233 kHz and the required limit to be met, the attenuation required from the EMI filter with a 6dB margin is given by:

$$A_{req,DM} = V_{DM} - V_{limit} + 6dB = 40 \log \frac{f_{DM}}{f_{DM,cor}} \quad (5.17)$$

$$f_{DM,cor} = \frac{1}{2\pi \sqrt{L_{DM} C_{DM}}} = \frac{1}{2\pi \cdot \sqrt{2L_{D1} C_{X1}}} \quad (5.18)$$

Based on equation (5.17), the values of the inductor and capacitor are chosen accordingly to meet the required filter attenuation using equation (5.18). The selected inductor and capacitor with their parasitic components, as shown in Fig. 5-23 are included to ensure more accurate filter attenuation prediction.

Fig. 5-23. EMI filter with parasitic components included

Fig. 5-24. Insertion loss of the EMI filter with parasitic components considered, (a) $L_D = 85 \mu H$, (b)

$$L_D = 210 \mu H$$

With the boost inductor nonlinearity ignored, the required filter attenuation of the DM component $A_{req,DM}$ is 37 dB. The DM filter should provide $A_{req,DM}$ at 233 kHz, f_{DM} is the

frequency of DM emission at 233 kHz. Based on equation (5.17), $f_{DM,cor}$ is 27.7 kHz. Substituting the filter capacitor $C_{X1} = C_{X2} = 200nF$ and $f_{DM,cor}$ into equation (5.18), the needed DM choke $L_{D1} = L_{D2}$ is about $82.5\mu H$, choose $L_{D1} = L_{D2} = 85\mu H$ for the EMI filter. With the parasitic component considered, the attenuation is 31.3 dB according to Fig. 5-24 (a). It can be found that the parasitic component can degrade the attenuation of the EMI filter, so the margin of 6 dB is necessary in the design process. Fig. 5-25 (b) shows that the measured conducted emission with the EMI filter has exceeded the limit at 233 kHz.

Fig. 5-25. DM noise spectrum, (a) without EMI filter, (b) with EMI filter

With the boost inductor nonlinearity considered, $A_{req,DM}$ is 45 dB, $f_{DM,cor}$ is 17.5 kHz, and the needed DM choke is $206.8\mu H$. $L_{D1} = L_{D2} = 210\mu H$ are chosen for the EMI filter. The chosen inductor has an equivalent parallel resistance (EPR) of 4 k Ω and equivalent parallel capacitance (EPC) of 5pF. The attenuation is 39.2 dB according to Fig. 5-24 (b). The measured conducted emission with the EMI filter is shown in Fig. 5-26 (b), and now the emission at 233 kHz is below the limit.

Fig. 5-26. DM noise spectrum, (a) without EMI filter, (b) with EMI filter

5.6 Summary

Interleaved PFC converter has a potential to improve the power factor and reduce the current ripple as compared to the traditional PFC converter. However, as a high frequency switching converter, it produces significant EMI. Hence, EMI prediction and suppression methods are necessary to ensure the compliance of PFC converter with the respective EMC standard. In this chapter, the non-linearity of the boost inductor is considered by adopting J-A model, which has been demonstrated to describe the hysteresis behavior of the inductors very well. The experimental results have also been collected to verify the accuracy of the DM conducted EMI prediction model. Based on the accurate EMI prediction, the DM EMI filter can be designed properly with confidence.

Chapter 6 Modeling Conducted EMI for Flyback PFC Converter

6.1 Introduction

Flyback PFC converter is another popular topology for its simplicity in implementation. Besides the power transistor switch and the high-frequency power transformer, there is no additional inductor required. Also, it provides wide-range and multiple output voltage with good transient response [139].

This chapter extends the use of the J-A model to account for the nonlinear effect of the flyback transformer with the aim to obtain more accurate conducted EMI prediction. This chapter is organized as follows: Section 6.2 introduces the flyback PFC converter and its detailed transformer model. In Section 6.3, EMI noise models are introduced. In Sections 6.4 and 6.5, simulation and experimental conducted EMI results are presented and compared. Finally, conclusion is drawn in Section 6.6.

6.2 Description of the Flyback PFC Converter

6.2.1 Converter Architecture

A typical flyback PFC converter is shown in Fig. 6-1, which has an advantage of providing a wide range of output voltage and isolation function to protect connected loads. This topology requires a small size magnetic core but produce high peak current. The high peak and high switching frequency current in the flyback PFC converter result in high conducted EMI. So it is necessary to construct an accurate conducted EMI model to predict the emission and allow the design of an EMI filter to suppress emission below the required limit.

The isolation transformer can be represented by modeling the physical transformer with an equivalent circuit consisting of an ideal transformer and the magnetizing inductance. The magnetizing must follow and hold the rule of flux when the whole system operates at steady state.

Fig. 6-1. Flyback PFC converter

Fig. 6-2. Flyback converter with transformer equivalent circuit model

When the transistor switch is on, the converter circuit is shown in Fig. 6-2(b) and the primary winding voltage is:

$$V_L = V_{rec} \quad (5.19)$$

When the switch is in off-state, the equivalent circuit is given in Fig. 6-2(c) and the primary winding voltage is:

$$V_L = -nV_o \quad (5.20)$$

Applying the inductor's flux balance in the on and off states yield,

$$\psi_{on} + \psi_{off} = V_{rec} \times KT + (-nV_o) \times (1-K)T = 0 \quad (5.21)$$

Solution for equation (5.21) leads to the following equation for output voltage. It shows that flyback converter is basically a buck-boost converter and the required output voltage can be obtained by varying the transformer turn ratio and duty ratio of the gate driving signal.

$$V_o = \frac{V_{rec} K}{n(1-K)} \quad (5.22)$$

Fig. 6-3 shows the converter controller topology, which operates at a boundary between continuous and discontinuous mode. There are two control loops in this controller. The slower outer loop controls the average output voltage with a specific value and the faster inner loop shapes the input current. A zero current detection (ZCD) circuit is used to start a new switch cycle and keep the system operates in boundary conduction mode. An S-R flip flop is used to generate the required PWM signal to drive the power MOSFET. Fig. 6-4 shows the control signal of the switch and the switch is turned on when its current decrease to zero.

Fig. 6-3. BCM controller

Fig. 6-4. Switcher signal, voltage and current

The high-frequency ripple of input current generates DM or CM conducted EMI.

Typically, the DM EMI is dominant for a two wire system without ground wire. The high-frequency ripple of the input current is the main contribution to the DM conducted EMI and an additional DM EMI filter is often required to attenuate the conducted EMI below the required limit.

6.2.2 Modeling of Isolation Transformer and Input Capacitor

1) Isolation transformer modeling

In the flyback converter, the isolated transformer is an essential part and it can be easily saturated due to its first quadrant B - H loop. It also consists of some parasitic elements that can affect accurate conducted EMI prediction. The J-A model will be adopted to describe the transformer magnetic behavior as shown in Fig. 6-5. In practice, there will be leakage flux and $L_{\lambda P}$ and $L_{\lambda S}$ denote the leakage inductances of primary and secondary coil, respectively. R_1 and R_2 are the resistances of primary and secondary coils, respectively.

Fig. 6-5. Transformer model

The extracted J-A model parameters and transformer parasitic components are listed in Table 6-1.

Table 6-1 Parameters for modeling

Parameter	M_s (A/m)	k (A/m)	$\hat{\sigma}$ (A/m)	α	c	R_m	C_m
Value	3.9×10^5	138	310	4×10^{-4}	0.5	47 kΩ	40 pF

Fig. 6-6 shows the actual core of the isolation transformer and its magnetic characteristics are given in Table 6-2. The turn number is calculated according to the required magnetizing inductance and A_L value.

$$n = \sqrt{\frac{L}{A_{L-value}}} \quad (5.22)$$

where n is the turn number, L is the magnetizing inductance and A_L is the inductance per turn in the core.

Fig. 6-6. ETD 29 core size

Table 6-2 Magnetic core parameters

Symbol	Parameter	Value	Unit
V_e	Effective volume	5470	mm^3
L_e	Effective length	72	mm
A_e	Effective area	76	mm^2
$L_{air} = 2 \times L_g$	Air gap length	0.1	mm

Fig. 6-7. Impedance frequency response of the flyback transformer magnetic inductor

The parasitic components of the magnetic inductor are measured with an impedance analyzer. Fig. 6-7 shows the frequency response of the transformer impedance from 100 kHz to 30 MHz. This circuit model will be used in the flyback PFC converter conducted EMI simulation model to predict DM conducted emission.

2) DC Filter Capacitor

Based on the measured impedance versus frequency response of input DC filter capacitor using a HP 4194A impedance analyzer, the parasitic components of the input filter are extracted. Fig. 6-8 shows the equivalent circuit model of the input filter with its equivalent series resistance (ESR) and equivalent series inductance (ESL). Fig. 6-9 shows the frequency response of the input filter capacitor from 100 kHz to 30 MHz.

Fig. 6-8. Equivalent circuit model of the input filter with parasitic components

Fig. 6-9. Frequency response of the input filter

6.3 EMI Noise Model Description

The flyback PFC converter topology including the nonlinear transformer and parasitic components is shown in Fig. 6-10. The input current of the PFC converter is dependent on the output power of the converter, inductance of the transformer and input voltage of the mains. The nonlinearity of the transformer is modeled with the hysteresis behavior described in details in the next section. Similarly, the equivalent model of the transformer with the parasitic components, as shown in Fig. 6-5, is also included in the simulation model as shown in Fig. 6-10.

For non-idealities of the power transistor switch (STP5NK80ZFP), drain-to-source turn-on resistance ($R_d = 1.9 \Omega$), equivalent series inductances of the drain and source terminals ($L_d = 3$ nH and $L_s = 6.5$ nH), and junction capacitances ($C_{gs} = 910$ pF, $C_{gd} = 98$ pF and $C_{ds} = 20$ pF) are obtained from the datasheet [140] and included in the simulation model for more realistic prediction of input current i_{in} .

Fig. 6-10. Configuration of the simulated flyback PFC converter system

Fig. 6-11. Circuit model of TDMS

The time-domain EMI measurement system (TDEMS) [48] as shown in Fig. 6-11 will be used to measure the conducted EMI. With the features of adding and subtracting, the CM and DM of conducted EMI can be measured without a noise separator. Once the time-

domain measurement results are collected, the frequency-domain conducted emission is digitally computed via the fast Fourier transform (FFT).

For the flyback PFC converter, the equivalent DM EMI model is shown in Fig. 6-12, which can be used to calculate the DM conducted emission spectrum. Once the noise current i_{in} is obtained from the circuit simulation as shown in Fig. 6-10, the noise current through the LISN can be computed and conducted EMI voltage V_{EMI} can be determined easily.

Fig. 6-12. Modified DM noise model

6.4 Simulation Results

The flyback PFC converter circuit shown in Fig. 6-10 is simulated with MATLAB SIMULINK. Simulation is conducted for the specification given in Table 6-3. The magnetization inductance of the transformer is 1.25mH if the magnetic nonlinearity is ignored. When the switch is on, the energy is stored in the inductance of the transformer. When the switch is off, the energy is transferred to the output section of the system.

As shown in Fig. 6-10, the nonlinear magnetic inductor model has been incorporated with the flyback PFC converter to simulate the system dynamics. Based on the previous introduction for the J-A model, the five first order differential equations can be implemented in the embedded MATLAB function, where the input is the transformer

voltage and the output is the transformer current. Such electrical model can be used to describe the transformer nonlinearity in the SIMULINK.

Table 6-3 Flyback pfc converter specifications

Specifications	
Input voltage (rms)	85-265V
Input frequency	50Hz
Output voltage	48V
Output power	24W
Transformer turn ratio	3.8:1

In order to apply the J-A model in the SIMULINK, a derivation for the magnetic inductor's voltage and current is introduced. According to the Ampère's circuital law:

$$HL_e + H_{air}L_{air} = HL_e + \frac{B}{\mu_0}L_{air} = H(L_e + L_{air}) + ML_{air} = ni \quad (5.23)$$

where i is the transformer primary current, L_{air} is the air gap length of the magnetic core, H is transformer core magnetic field and H_{air} is the air gap magnetic field. The flux density in the air and magnetic core are the same.

Then

$$V_L = \mu_0 nA \frac{d}{dt}(H + M) = \mu_0 nA \cdot \frac{dH}{dt} \cdot \left[1 + \frac{dM}{dH} \right] \quad (5.24)$$

Integrating both side of equation (5.24) results in,

$$M(t) = \int_{t_0}^t \frac{V_L}{\mu_0 nA} dt - H(t) + M_0 + H_0 \quad (5.25)$$

where M_0 is the initial magnetization, t_0 is the start moment of the integration, and V_L denotes the transformer primary voltage. Equation (5.25) can be discretized as below:

$$M(j) = \sum_{m=1}^j \frac{V_L(m)}{\mu_0 n A} \times \Delta t - H(j) + M_0 + H_0 \quad (5.26)$$

Form equation (5.24), dM / dH can be described by:

$$\frac{dH}{dt} = \frac{1}{\mu_0 n A} \cdot \frac{V_L}{1 + dM / dH} = \frac{1}{\mu_0 n A} \cdot \frac{V_L}{1 + dM / dH} \quad (5.27)$$

Equation (5.27) can be discretized using

$$\frac{H(j+1) - H(j)}{\Delta t} = \frac{1}{\mu_0 n A} \cdot \frac{V_L(j)}{1 + (dM / dH)(j)} \quad (5.28)$$

Fig. 6-13 shows the flowchart of the J-A hysteresis model which will be used to describe the transformer in the flyback converter. In order to ensure the stability of the hysteresis model in the simulation, the simulation step should be much smaller than the period of switching frequency, usually it is set less than one tenth of period of the switching frequency. The smaller the time step, the better the system stability but at the expense of simulation time. As a compromise between system stability and computational efficiency, the simulation step is suggested to be between one tenth and one hundredth of the period of the switching frequency.

Fig. 6-13. Flowchart of the J-A hysteresis model

With the transformer's voltage and current relationship, the nonlinear inductive model can be built into the Embedded MATLAB Function of SIMULINK using the following six numerical procedures:

- 1) Initialize magnetic field, magnetization and set the initial value of dM / dH as zero.
- 2) Calculate the effective magnetic field H_e , anhysteretic magnetization M_{an} and dM_{an} / dH using equations (4.7), (4.9) and (4.10).
- 3) The transformer voltage $V_L(m)$ is captured from the overall circuit simulation at each time step, the value of $M(j)$ can then be calculated using equation (5.26).
- 4) Calculate the sign of dH / dt using $H(j) - H(j-1)$, determine the sign of $M_{an}(j) - M(j)$ and value of δ , and calculate the value of dM / dH using equations (4.13) and (4.14) at last.
- 5) Calculate the $H(j+1)$ using equation (5.28). Update the current $i(j+1)$ using equation (5.23).
- 6) Repeat steps 2 through 5 until simulation stops.

Fig. 6-14 shows the input voltage, current and average output voltage, which is kept at 48V. It can be seen that the power factor is close to utility due to PFC circuit. It shows that the flyback PFC converter in steady condition state. It shows that the J-A model does not affect the system stability in the simulation and fits seamlessly into the complex mathematical model described by MATLAB scripts in SIMULINK environment with other Simulink functional blocks.

Fig. 6-14. Input voltage and drain-source current and output voltage, $V_{in-rms} = 110V$

Fig. 6-15. (a) Drain voltage, (b) drain current

Fig. 6-16. Zoom in drain voltage and current

Fig. 6-15 shows the drain-source voltage and current with the J-A model included. The drain-source voltage is the caurce of the CM noise and the drain current is the major contribution of the DM noise. Fig. 6-16 shows the zoom in drain source currant and voltage.

Fig. 6-17 shows the simulated hysteresis curve of transformer core, which shows the J-A hysteresis model works well in the transformer simulation model. Fig. 6-17 (a) and (b) show the simulated magnetic field and magnetization against time; and Fig. 6-17 (c) shows the main loop at power frequency and Fig. 6-17 (d) shows the minor loop at switching frequency. It means that the J-A model can be applied in the power transformer magnetic model at low and high frequency appropriately. The observed results are useful to choose the appropriate magnetic core and number of turns of the transformer to avoid core saturation.

Fig. 6-17. Relationship between magnetic field and magnetization

From the circuit analysis point of view, the inductance L is given by:

$$L = \frac{V}{di/dt} \quad (5.29)$$

From equations (5.23), (5.24) and (5.29), it leads to:

$$L = n^2 A \frac{dB/dH}{L_e + \frac{L_{air}}{\mu_0} \frac{dB}{dH}} = n^2 A \frac{\mu_0 (dM/dH + 1)}{L_e + L_{air} (dM/dH + 1)} \quad (5.30)$$

Conventional power converter simulation assumes that the transformer inductance is invariant with its operating current. In reality, the transformer inductance varies with its operating current [141]. Fig. 6-18 (a) and (b) show the drain current with and without J-A model included. Fig. 6-18 (c) shows the comparison of the close-up drain current waveforms with and without the J-A model. It is clearly observed that the change of current with time is not constant when the nonlinearity of the transformer is included with the J-A model. Fig. 6-19 shows the simulated inductance against operating current in accordance with equations (5.29) and (5.30). Due to the reduction in inductance with

increasing current, the amplitude of the current ripple becomes larger and therefore the conducted EMI is expected to be higher than that without nonlinearity included in the model.

Fig. 6-18. (a) Drain current without J-A model, (b) drain current with J-A model, (c) close-up comparison of the current waveforms with and without J-A model

Fig. 6-19. Inductance versus operating current

6.5 Simulation and Measurement Results Comparison

6.5.1 Noise Spectrum Comparison

A flyback converter is designed and assembled based on the specifications given in Table 6-3. Based on the EMI measurement as given in Fig. 6-11, the experimental setup is shown in Fig. 6-20. The power is supplied to the PFC converter through a line impedance stabilization network (LISN). The purpose of LISN is to ensure mains impedance is well defined for measurement repeatability. The output power of the PFC converter can be adjusted in accordance with the loading condition. The time-domain conducted noise is measured by a TDS7054 digital signal oscilloscope (DSO) at the RF output of the LISN. The frequency-domain conducted emission results are then obtained with the use of FFT function. Based on the prediction and measurement results, an EMI filter is designed and inserted between the power source and PFC converter to suppress the EMI noise level to meet specific EMI standard.

As the flyback converter is powered by two-wire, it is found that the DM emission is the dominant component. Hence, only the DM emission results will be presented. Fig. 6-21 shows the time-domain conducted EMI waveform, which is obtained and calculated using the DSO math function as shown in Fig. 6-20.

Fig. 6-20. Experiment setup in the laboratory

Fig. 6-21. Time-domain conducted emission

Fig. 6-22 (a) shows the measured conducted emission in frequency-domain using FFT function. Fig. 6-22 (b) and (c) show the simulated conducted emissions without and with the transformer nonlinearity included, respectively. It is clearly observed that the simulated result with nonlinear effect included closely agrees with the measured result.

Fig. 6-22. Frequency-domain conducted emissions: (a) measured results (b) simulated results without J-A model (c) simulated results with J-A model

For ease of comparison, conducted emission at the fundamental switching frequency and several harmonics of are summarized in Table 6-4. It shows that without the nonlinear behavior of the transformer included, the simulated result under-estimates the conducted emission, which can lead to an under-designed EMI filter and fails the EMC compliance testing at the final product qualification stage. With the nonlinearity of the transformer properly simulated by the J-A model, the predicted conducted emission resembles well

with the measured conducted emission, therefore proper EMI filter could be designed to avoid EMI failure. Comparing to simulation model without consideration of nonlinearity, the proposed prediction model has improved the accuracy significantly. Some discrepancies between simulation and experiment are expected due to some assumptions made for the switching device. More accurate model of the switching device can be developed and modeled for even better prediction accuracy.

Table 6-4 Comparison of measured and simulated conducted DM emission

Frequency [kHz]	Experiment spectrum [dB μ V]	Simulation without J-A model [dB μ V]	Simulation with J-A model [dB μ V]
61	107	100	105
122	104	92	101
183	93.5	85	93
244	83	78	84
305	77.8	75.5	78
366	78	70	77
427	73.8	68.7	74
488	67	63.8	68
549	69	63.9	68
610	67.2	59	66

6.5.2 EMI Filter Design

Based on the accurate prediction, the emission has exceeded the EMI limit and an EMI filter is a necessary to bring the emission below the limit. Fig. 6-22 (b) shows the simulated conducted emission ignoring transformer nonlinearity. The highest emission frequency is 183 kHz (3rd harmonic) and the predicted emission level at 183 kHz is 85dB μ V. By comparing it with the limit and adding a 6 dB margin, the EMI filter attenuation required is:

$$A_{req,DM} = V_{DM} - V_{limit} + 6dB = 40 \log \frac{f_{DM}}{f_{DM,cor}} \quad (5.31)$$

$$f_{DM,cor} = \frac{1}{2\pi\sqrt{L_{DM}C_{DM}}} = \frac{1}{2\pi\sqrt{2L_{D1}C_{X1}}} \quad (5.32)$$

where $A_{req,DM}$ is the desired attenuation of the DM emission at f_{DM} (183 kHz) = $85 - 65 + 6 = 26 \text{ dB}$, and the calculated $f_{DM,cor}$ will be 41kHz based on equation (5.31). If the filter capacitor $C_{X1} = C_{X2} = 220\text{nF}$ is chosen and with the corner frequency calculated earlier, the required filter inductor will be $34\mu\text{H}$ based on equation (5.32). The inductor value is chosen to be $35\mu\text{H}$ for the EMI filter. With the given filter, the measured conducted emissions without and with the filter are shown in Fig. 6-23 (a) and (b), respectively. It is found that the EMI filter is under-designed and conducted emission at lower frequency still exceeds the limit.

Fig. 6-23. DM noise spectrum before and after EMI filter installation, (a) without EMI filter, (b) with EMI filter

When the nonlinear behavior of the transformer is taken into consideration, the simulated conducted emission level at 183 kHz is 93 dB μ V. The desired attenuation is 34 dB and $f_{DM,cor}$ is 26 kHz. With the same filter capacitor, the required filter inductor is 85 μ H. Choosing 85 μ H for the EMI filter, the measured conducted emissions without and with the filter are shown in Fig. 6-24 (a) and (b), respectively. Now, the emission level is well below the limit.

Fig. 6-24. DM noise spectrum before and after EMI filter installation, (a) without EMI filter, (b) with EMI filter

6.6 Summary

Flyback PFC converter is a common topology due to its obvious advantage but its large leakage inductance also causes significant high frequency conducted emission and it is

neccenary to develop an accurate EMI prediction model for the flyaback PFC converter. In this chapter, an EMI prediction model including the transformer hysteresis is proposed and a more accurate EMI model has been established. Compare with the traditional EMI prediction model, the improved model comprises the transformer hsysteresis is described by the J-A model is more accurate. The improvement in the accuracy of the prediction model avoids underestimation of conducted EM and therefore an appropriate EMI filter can be designed.

Chapter 7 Conclusions and Recommended Future Work

7.1 Conclusion

This thesis has developed the conducted EMI prediction model for the PFC converters including interleaved and flyback topologies. The major faced difficulties, accomplishments and contribution in the whole PhD research are summarized as follows.

Comparing with passive PFC converter, active PFC solution is a better option to achieve nearly unity power factor and sinusoidal input current waveform with extremely low THD but high frequency switching can produce conducted EMI noise and impairs sensitive electronic equipment in the same power grid. Thus conducted EMI noise prediction and suppression for the PFC converter is an important aspect in PFC converter design process.

To accurately model a power convertor circuit for the purpose of EMI prediction and analysis, proper modeling of circuit components are critical. The ability to construct an accurate EMI noise prediction model before realization of a prototype is necessary so that early EMI problems can be identified and rectified to meet the required EMC specifications.

Inductive components, such as inductors and transformers with magnetic cores are common sights in power convertor circuits. Accurate modeling of their behaviors plays a key role for accurate EMI prediction. Therefore, there is a need to develop a model that can handle the non-linear hysteresis of magnetic core so that the non-linearity of the inductive components can be included in power convertor circuit design.

Looking for a suitable method to model the magnetic nonlinearity of the inductor or transformer is rather challenging in a circuit simulation platform. There are many attempts to construct mathematical models that handle nonlinearity of magnetic components but only several of these have been proven to be feasible. These models are the Preisach model, the Stoner-Wohlfarth model, the Globus model, and the Jiles-

Atherton (J-A) model. The merits and constraints of each of these models need to be understood, as each of them is only capable of giving an approximate description for the magnetic materials under certain conditions. In addition, the chosen model must be able to interface with the circuit-based simulation tool for ease of analysis,

To enable the implementation of nonlinearity of an inductive component in a circuit-based simulation tool, it is necessary to express an inductive component with magnetic core in terms of its current and voltage. The J-A model has proven such an ability to handle this relationship. It consists of a group of first-order ordinary differential equations, which can be solved numerically to provide the relationship between the magnetic core's magnetization and magnetic field so that its nonlinear characteristics and magnetic hysteresis can be properly modeled. In the simulation part, it is difficult to integrate the nonlinear magnetization model with the power converter due to the complex hysteresis description and controller loop. Choosing a small simulation time step can increase the stability of the whole system but at the expense of simulation time. Hence, as a compromise between system stability and computational efficiency, the simulation time step is kept below one tenth but higher than one hundredth of the period of the switching frequency.

In order to obtain accurate EMI model, the usual parasitic components of inductor, capacitor and resistor in the PFC converter and EMI filter need to be considered. The passive components modeling which are applied in the EMI prediction model are introduced in section 3.3. Parasitic elements of the EMI filter must be considered to predict filter's performance correctly and to facilitate the proper design process. Based on the measured impedance response, parasitic components of the inductors or capacitors of the EMI filter are extracted and included in the simulation model to reflect the true EMI filter performance.

In terms of EMI prediction, the model developed in this thesis predicts conducted EMI noise more accurately than the traditional model. As a result, correct EMI filter can be designed with confidence.

7.2 Recommended Future Work

The following aspects that are worth to be explored as future work:

- Improvement on the non-linear model

Although the J-A model has been developed in this thesis to model the magnetic material phenomenon with reasonably accuracy, some other models such as Preisach model or self-learning algorithm based on Gaussian distribution are also worth exploring to investigate the non-linearity of the magnetic material. The Preisach model of hysteresis generalizes hysteresis loops as the parallel connection of independent relay hysterons. It has become a widely accepted model of hysteresis. The Preisach model is especially accurate in the field of ferromagnetism, as the ferromagnetic material can be described as a collection of small domains. For example, iron may have randomly distributed magnetic domains, resulting in a net magnetic field of zero. Based on the proposed self-learning algorithm, accuracy of the estimated inductor current can be improved to enhance the overall model accuracy. Also, other more efficient J-A model parameters extraction methods can also be investigated.

- Improvement on numerical analysis

Differential evolution (DE) method based on the experiment data or secant method based on the datasheet is used to extract the hysteresis model parameters in this thesis. In the aspect of numerical analysis, the secant method is a root-finding algorithm that uses a succession of roots of secant lines to better approximate a root of a function. The secant method can be thought of as a finite difference approximation of Newton's method. Hybrid technique can be proposed to determine the parameters of the hysteresis model. The secant method based on the material datasheet can be used to find the initial parameters, then the DE method can converge more quickly to obtain the optimal parameters with the initial values. During the optimization process, the difference between the mutation vectors and the base vector becomes smaller, which means the members of the population are getting closer to the optimal solution

- High power converter applications

For high power applications, the interleaved multi-channel PFC is becoming more and more popular. It can effectively reduce the input ripple current due to the ripple cancellation effect. It is generally believed that by properly choosing the interleaving channel number and the switching frequency, the EMI noise can be reduced and the EMI filter size and cost can be effectively saved. Further more, PFC converter with asymmetrical interleaving strategy reduces the current ripple in the output capacitor. Smaller current ripple can prolong the electrolytic capacitor life time. The hysteresis model can be applied to model more than two boost inductors at the same time.

- EMI filter design

The pursuit of obtaining higher power density AC/DC and DC/DC converters leads to increasing switching frequencies in order to reduce the size of the energy storage elements. These energy storage elements usually influence the overall size of a converter significantly. This has resulted in the development of electromagnetically integrated structures, which combine several functions in one passive component. The passive EMI filter has the advantage of simple topology and low cost, but it also has disadvantage of large size and narrow compensation bandwidth as compared to the active EMI filter (AEF) or integrated EMI filter in the power electronic systems. The AEF has a similar compensation principle with the active power filter (APF), it generates a current or voltage ripples to cancel the EMI noise. Thus, hybrid filter with a combined passive and active EMI filter can be investigated. In order to reduce the filter volume and there is a trend to integrate discrete EMI filter components together. Usually there are two types of integration methods for passive components. Firstly, the integrated passives with planar printed circuit board (PCB) and secondly, the integrated passive with flexible multilayer (FML) foil. The active EMI filters use feedback or feed forward methods in the control topologies to suppress EMI noise. Comparing to the passive EMI filters, active EMI filters have the advantage of wide switching frequency applications and more efficient to eliminate the noise, of course, at the expense of higher cost and circuit complexity. Three

phase AC or DC rectifier is another major EMI source in the power electronic system, EMI noise prediction and suppression for this type of rectifier becomes necessary. Comparing with the single phase system, there are more high frequency switchers and higher current/voltage in the three phase system, thus it has more complex EMI paths and require a more comprehensive model. Due to the higher current and voltage in the three phase systems, choosing the appropriate filter components is an important issue not to be neglected.

Author's Publications

This research work has led to the following publications.

Journal Papers

- [1] Yitao. Liu and D. Zhang, "Comprehensive modeling of resonant inverter for driving fluorescent lamp with the consideration of nonlinear magnetization of inductor," *IEEE Transactions on Magnetics*, vol. 47, pp. 4318-4321, 2011.
- [2] Yitao. Liu and D. Zhang, "Effect of nonlinear inductor behavior on the performance of interleaved power factor correction," *IEEE Transactions on Magnetics*, vol. 48, pp. 4143-4147, 2012.
- [3] Yitao. Liu, K. Y. See, and K. J. Tseng, "Conducted EMI Prediction of the PFC Converter Including Nonlinear Behavior of Boost Inductor," *IEEE Transactions on Electromagnetic Compatibility*, vol. 55, pp. 1107-1114, 2014.
- [4] Yitao. Liu and K. Y. See, "Effect of Nonlinear Magnetization of Transformer on Conducted EMI of Flyback PFC Converter," *IET Power Electronics*, 2014, under review.

Conference Papers

- [1] Yitao. Liu and D. Zhang, "Modeling the dynamics of a circuit containing saturable magnetic cores based on J-A model," in *IPEC, 2010 Conference Proceedings*, 2010, pp. 46-51.
- [2] Yitao. Liu and D. Zhang, "Study of effect of the nonlinear inductor behavior on the performance of resonant tank for driving fluorescent lamp," in *Power Electronics and Drive Systems (PEDS), 2011 IEEE Ninth International Conference on*, 2011, pp. 1153-1157.
- [3] Yitao. Liu and D. Zhang, "Comparative THD Analysis of Different Resonant Tanks for Driving Fluorescent Lamp," in *Power and Energy Engineering Conference (APPEEC), 2011 Asia-Pacific*, 2011, pp. 1-5.

Bibliographies

- [1] F. Costa and D. Magnon, "Graphical analysis of the spectra of EMI sources in power electronics," *IEEE Trans. Power Electron.*, vol. 20, pp. 1491-1498, 2005.
- [2] I. Stevanovic, S. Skibin, M. Masti, and M. Laitinen, "Behavioral Modeling of Chokes for EMI Simulations in Power Electronics," *Power Electronics, IEEE Transactions on*, vol. 28, pp. 695-705, 2013.
- [3] S. Hayashi and M. Yamada, "EMI-noise analysis under ASIC design environment," *IEEE Trans. Comput. Aided Des. Integr. Circuits Syst.*, vol. 19, pp. 1337-1346, 2000.
- [4] Z. Li, L. Qingmin, W. Wei, and S. Wah Hoon, "Electromagnetic Interference Analysis in HV Substation Due to a Static Var Compensator Device," *IEEE Trans. Power Delivery*, vol. 27, pp. 147-155, 2012.
- [5] L. Jih-Sheng, H. Xudong, E. Pepa, C. Shaotang, and T. W. Nehl, "Inverter EMI modeling and simulation methodologies," *IEEE Trans. Ind. Electron.*, vol. 53, pp. 736-744, 2006.
- [6] J. Meng and W. Ma, "A new technique for modeling and analysis of mixed-mode conducted EMI noise," *IEEE Trans. Power Electron.*, vol. 19, pp. 1679-1687, 2004.
- [7] J. S. Scott and C. Van Zyl, *Introduction to EMC*: Butterworth-Heinemann, 1997.
- [8] J. Colotti, "EMC design fundamentals," in *IEEE Systems, Applications and Technology Conference, LISAT 2006*, 2006, pp. 5-5.
- [9] L. Tihanyi, *EMC in power electronics*: Newnes, 1995.

- [10] Q. Z. C. Henglin, "State of Art of Electromagnetic Compatibility Research on Power Electronic Equipment [J]," *Transactions of China Electrotechnical Society*, vol. 7, p. 000, 2007.
- [11] J. Meng and W. Ma, "Power Converter EMI Analysis Including IGBT Nonlinear Switching Transient Model," *IEEE Trans. Ind. Electron.*, vol. 53, pp. 1577-1583, 2006.
- [12] M. Jin, M. Weiming, P. Qijun, Z. Lei, and Z. Zhihua, "Multiple Slope Switching Waveform Approximation to Improve Conducted EMI Spectral Analysis of Power Converters," *IEEE Trans. Electromagn. Compat.*, vol. 48, pp. 742-751, 2006.
- [13] M. Jin, M. Weiming, P. Qijun, Z. Zhihua, and Z. Lei, "Noise Source Lumped Circuit Modeling and Identification for Power Converters," *IEEE Trans. Ind. Electron.*, vol. 53, pp. 1853-1861, 2006.
- [14] M. M. Hernando, A. Fernandez, M. Arias, M. Rodriguez, Y. Alvarez, and F. Las-Heras, "EMI Radiated Noise Measurement System Using the Source Reconstruction Technique," *IEEE Trans. Ind. Electron.*, vol. 55, pp. 3258-3265, 2008.
- [15] V. Tarateeraseth, "EMI filter design: Part II: Measurement of noise source impedances," *Electromagnetic Compatibility Magazine, IEEE*, vol. 1, pp. 42-49, 2012.
- [16] A. R. Rivera-Ramos and M. Jimenez, "Analytical models for estimating parasitic components in power electronics PCBs," in *Circuits and Systems, 2005. 48th Midwest Symposium on*, 2005, pp. 1235-1238 Vol. 2.
- [17] L. Jih-Sheng, H. Xudong, C. Shaotang, and T. W. Nehl, "EMI characterization and simulation with parasitic models for a low-voltage high-current AC motor drive," *IEEE Trans. Ind. Appl.*, vol. 40, pp. 178-185, 2004.

- [18] H. Xudong, E. Pepa, L. Jih-Sheng, A. R. Hefner, Jr., D. W. Berning, C. Shaotang, *et al.*, "EMI characterization with parasitic modeling for a permanent magnet motor drive," in *Industry Applications Conference, 2003. 38th IAS Annual Meeting. Conference Record of the*, 2003, pp. 416-423 vol.1.
- [19] W. Shuo, F. C. Lee, and J. D. Van Wyk, "A Study of Integration of Parasitic Cancellation Techniques for EMI Filter Design With Discrete Components," *IEEE Trans. Power Electron.*, vol. 23, pp. 3094-3102, 2008.
- [20] J. C. Crebier and J. Ferrieux, "PFC full bridge rectifiers EMI modeling and analysis-common mode disturbance reduction," *IEEE Trans. Power Electron.*, vol. 19, pp. 378-387, 2004.
- [21] G. Xun, Josifovic, x, I., and J. A. Ferreira, "Modeling and Reduction of Conducted EMI of Inverters With SiC JFETs on Insulated Metal Substrate," *IEEE Trans. Power Electron.*, vol. 28, pp. 3138-3146, 2013.
- [22] J. Meng, W. Ma, Q. Pan, J. Kang, L. Zhang, and Z. Zhao, "Identification of Essential Coupling Path Models for Conducted EMI Prediction in Switching Power Converters," *IEEE Trans. Power Electron.*, vol. 21, pp. 1795-1803, 2006.
- [23] M. Moreau, N. Idir, and P. Le Moigne, "Modeling of Conducted EMI in Adjustable Speed Drives," *IEEE Trans. Electromagn. Compat.*, vol. 51, pp. 665-672, 2009.
- [24] D. Jiles and D. Atherton, "Ferromagnetic hysteresis," *IEEE Trans. Magn.*, vol. 19, pp. 2183-2185, 1983.
- [25] D. C. Jiles, "A self consistent generalized model for the calculation of minor loop excursions in the theory of hysteresis," *IEEE Trans. Magn.*, vol. 28, pp. 2602-2604, 1992.

- [26] D. C. Jiles, J. B. Thoele, and M. K. Devine, "Numerical determination of hysteresis parameters for the modeling of magnetic properties using the theory of ferromagnetic hysteresis," *IEEE Trans. Magn.*, vol. 28, pp. 27-35, 1992.
- [27] P. Kis and A. Iványi, "Jiles-Atherton Model Implementation to Edge Finite Element Method," PhD thesis, Budapest University of Technology and Economics, 2007.
- [28] J. Cale, S. D. Sudhoff, and R. R. Chan, "A Field-Extrema Hysteresis Loss Model for High-Frequency Ferrimagnetic Materials," *IEEE Trans. Magn.*, vol. 44, pp. 1728-1736, 2008.
- [29] D. C. Jiles and D. L. Atherton, "Theory of ferromagnetic hysteresis (invited)," *Journal of Applied Physics*, vol. 55, pp. 2115-2120, 1984.
- [30] A. Zuccato and L. Rossetto, "Understanding and complying with CISPR and IEC 1000 standards on EMC," in *Tut. Eur. Power Electron. Drives Conf., Trondheim, Norway*, 1997.
- [31] P. Andersen, "An overview of automotive EMC standards," in *Electromagnetic Compatibility, 2006. EMC 2006. 2006 IEEE International Symposium on*, 2006, pp. 812-816.
- [32] G. L. Skibinski, R. J. Kerkman, D. Schlegel, and R. Automation, "EMI EMISSIONS," *IEEE Industry Applications Magazine*, vol. 5, pp. 47-81, 1999.
- [33] "American National Standard Dictionary for Technologies of Electromagnetic Compatibility (EMC), Electromagnetic Pulse (EMP), and Electrostatic Discharge (ESD) (Dictionary of EMC/EMP/ESD Terms and Definitions)," *ANSI C63.14-1992*, p. 1, 1992.

- [34] C. A. Marshman and A. C. Marvin, "The new emc directive 2004/108/EC," in *Electromagnetic Compatibility, 2006. EMC 2006. 2006 IEEE International Symposium on*, 2006, pp. 482-485.
- [35] Overview of EMI/EMC [Online]. Available: <http://www.sameercal.org/docs/emi-emc-pdf.pdf>
- [36] D. Morgan, *A handbook for EMC testing and measurement* vol. 8: Iet, 1994.
- [37] Z. Hanzelka, A. Kempski, and R. Smoleński, "Quality Problems in Smart Networks," in *Power Electronics in Smart Electrical Energy Networks*, ed: Springer, 2008, pp. 107-146.
- [38] Available: www.iec.ch
- [39] R. Redl, "Power electronics and electromagnetic compatibility," in *Power Electronics Specialists Conference, 1996. PESC '96 Record., 27th Annual IEEE*, 1996, pp. 15-21 vol.1.
- [40] D. A. Weston, *Electromagnetic Compatibility: Principles and Applications* 1991.
- [41] C. R. Paul, *Introduction to electromagnetic compatibility* vol. 184: John Wiley & Sons, 2006.
- [42] F. Mihalic and D. Kos, "Reduced conductive EMI in switched-mode DC–DC power converters without EMI filters: PWM versus randomized PWM," *IEEE Trans. Power Electron.*, vol. 21, pp. 1783-1794, 2006.
- [43] S. Ogasawara and H. Akagi, "Modeling and damping of high-frequency leakage currents in PWM inverter-fed AC motor drive systems," in *Industry Applications Conference, 1995. Thirtieth IAS Annual Meeting, IAS '95., Conference Record of the 1995 IEEE*, 1995, pp. 29-36 vol.1.

- [44] EN 60598-1: 2004, MEASUREMENT AND TEST REPORT For INTELSTEER PTE LTD [Online]. Available: <http://www.ekolite.com/intelight/images/certificates/lvdill2w.pdf>
- [45] G. L. Skibinski, R. J. Kerkman, and D. Schlegel, "EMI emissions of modern PWM AC drives," *Industry Applications Magazine, IEEE*, vol. 5, pp. 47-80, 1999.
- [46] G. Grandi, D. Casadei, and U. Reggiani, "Common- and differential-mode HF current components in AC motors supplied by voltage source inverters," *IEEE Trans. Power Electron.*, vol. 19, pp. 16-24, 2004.
- [47] W. Shuo, F. C. Lee, and W. G. Odendaal, "Characterization, evaluation, and design of noise Separator for conducted EMI noise diagnosis," *IEEE Trans. Power Electron.*, vol. 20, pp. 974-982, 2005.
- [48] D. a. S. Circuits, IEE Proceedings -Kumar, M. and V. Agarwal, "Power line filter design for conducted electromagnetic interference using time-domain measurements," *IEEE Trans. Electromagn. Compat.*, vol. 48, pp. 178-186, 2006.
- [49] W. Shen, F. Wang, D. Boroyevich, and Y. Liu, "Definition and acquisition of CM and DM EMI noise for general-purpose adjustable speed motor drives," in *Power Electronics Specialists Conference, 2004. PESC 04. 2004 IEEE 35th Annual*, 2004, pp. 1028-1033 Vol.2.
- [50] C. Po-Shen and L. Yen-Shin, "Effective EMI Filter Design Method for Three-Phase Inverter Based Upon Software Noise Separation," *IEEE Trans. Power Electron.*, vol. 25, pp. 2797-2806, 2010.
- [51] W. Shuo, "Modeling and Design of EMI Noise Separators for Multiphase Power Electronics Systems," *IEEE Trans. Power Electron.*, vol. 26, pp. 3163-3173, 2011.

- [52] W. Shuo, L. Fang, and F. C. Lee, "Characterization and Design of Three-Phase EMI Noise Separators for Three-Phase Power Electronics Systems," *Power Electronics, IEEE Transactions on*, vol. 26, pp. 2426-2438, 2011.
- [53] Y. Suzuki, T. Teshima, I. Sugawara, and A. Takeuchi, "Experimental studies on active and passive PFC circuits," in *Telecommunications Energy Conference, 1997. INTELEC 97., 19th International*, 1997, pp. 571-578.
- [54] J. Lam and P. K. Jain, "A simple single switch electronic ballast for compact fluorescent lamps with passive power factor correction (PFC) and soft switching capability," in *IECON 2012-38th Annual Conference on IEEE Industrial Electronics Society*, 2012, pp. 4503-4508.
- [55] F. Zare, "EMI issues in modern power electronic systems," *The IEEE EMC Society Newsletters*, pp. 53-58, 2009.
- [56] M. C. Caponet, F. Profumo, and A. Tenconi, "EMI filters design for power electronics," in *Power Electronics Specialists Conference, 2002. pesc 02. 2002 IEEE 33rd Annual*, 2002, pp. 2027-2032.
- [57] R. W. Erickson and D. Maksimovic, *Fundamentals of Power Electronics*: Springer, 2001.
- [58] M. Xiaolin, R. Ayyanar, and H. K. Krishnamurthy, "Optimal Variable Switching Frequency Scheme for Reducing Switching Loss in Single-Phase Inverters Based on Time-Domain Ripple Analysis," *IEEE Trans. Power Electron.*, vol. 24, pp. 991-1001, 2009.
- [59] M. Vilathgamuwa, J. Deng, and K. J. Tseng, "EMI suppression with switching frequency modulated DC-DC converters," *Industry Applications Magazine, IEEE*, vol. 5, pp. 27-33, 1999.
- [60] E. B. Joffe and K. S. Lock, *Grounding on Printed Circuit Boards (PCBs)*, 2010.

- [61] K. Gulez, N. Mutoh, M. Ogata, F. Harashima, and K. Ohnishi, "An approximation to EMI noise problem to design an appropriate EMI filter including double-LISN," in *Industrial Electronics Society, 2001. IECON '01. The 27th Annual Conference of the IEEE*, 2001, pp. 1237-1242 vol.2.
- [62] J. Stahl, D. Kuebrich, and T. Duerbaum, "Characterisation of an effective EMI noise separation including a standard LISN," in *Electromagnetic Theory (EMTS), 2010 URSI International Symposium on*, 2010, pp. 13-16.
- [63] J. Stahl, D. Kuebrich, and T. Duerbaum, "Modification and characterization of a standard LISN for effective EMI noise separation," in *Electromagnetics in Advanced Applications (ICEAA), 2010 International Conference on*, 2010, pp. 39-42.
- [64] S. Braun, T. Donauer, and P. Russer, "A Real-Time Time-Domain EMI Measurement System for Full-Compliance Measurements According to CISPR 16-1-1," *IEEE Trans. Electromagn. Compat.*, vol. 50, pp. 259-267, 2008.
- [65] Z. Xiangming, Z. Zhihua, G. Fei, M. Jin, L. Wenlu, and H. Anqi, "Frequency spectrum analysis method for short-time and frequency conversion signals based on simulation EMI receiver," in *Electromagnetic Compatibility (APEMC), 2010 Asia-Pacific Symposium on*, 2010, pp. 1606-1609.
- [66] W. Schaefer, "Measurement of impulsive signals with a spectrum analyzer or EMI receiver," in *Electromagnetic Compatibility, 2005. EMC 2005. 2005 International Symposium on*, 2005, pp. 267-271 Vol. 1.
- [67] A. W. Scott and R. Frobenius, *RF measurements for cellular phones and wireless data systems*, 1 ed.: Wiley. com, 2011.
- [68] C. R. Paul and K. B. Hardin, "Diagnosis and reduction of conducted noise emissions," *IEEE Trans. Electromagn. Compat.*, vol. 30, pp. 553-560, 1988.

- [69] G. W. Chang, Y. Chia-Ming, and C. Wei-Cheng, "Meeting IEEE-519 current harmonics and power factor constraints with a three-phase three-wire active power filter under distorted source voltages," *IEEE Trans. Power Delivery*, vol. 21, pp. 1648-1654, 2006.
- [70] D. Xiong, Z. Luowei, L. Hao, and T. Heng-Ming, "DC Link Active Power Filter for Three-Phase Diode Rectifier," *IEEE Trans. Ind. Electron.*, vol. 59, pp. 1430-1442, 2012.
- [71] M. Youssef, G. Antonini, E. Clavel, J. Roudet, S. Cristina, and A. Orlandi, "Conducted and radiated EMI characterisation of power electronics converter," in *Industrial Electronics, 1997. ISIE '97., Proceedings of the IEEE International Symposium on*, 1997, pp. SS207-SS211 vol.1.
- [72] K. Y. See and E. M. Freeman, "Modelling finite-size PCBs for radiated EMI analysis," *Electronics Letters*, vol. 33, pp. 570-571, 1997.
- [73] C. Jettanasen, "Passive common-mode EMI filter adapted to an adjustable-speed AC motor drive," in *IPEC, 2010 Conference Proceedings*, 2010, pp. 1025-1030.
- [74] I. Kovacevic, A. Muesing, T. Friedli, and J. W. Kolar, "Electromagnetic Modeling of EMI Input Filters," in *Integrated Power Electronics Systems (CIPS), 2012 7th International Conference on*, 2012, pp. 1-9.
- [75] D. Hamza, Q. Mei, and P. K. Jain, "Application and Stability Analysis of a Novel Digital Active EMI Filter Used in a Grid-Tied PV Microinverter Module," *IEEE Trans. Power Electron.*, vol. 28, pp. 2867-2874, 2013.
- [76] C. Wenjie, Y. Xu, and W. Zhaoan, "An experimental study and comparison of common mode and differential mode noise compensation characteristic for active EMI filter," in *Power Electronics Specialists Conference, 2008. PESC 2008. IEEE*, 2008, pp. 4399-4404.

- [77] W. Shuo, C. Rengang, J. D. Van Wyk, F. C. Lee, and W. G. Odendaal, "Developing parasitic cancellation technologies to improve EMI filter performance for switching mode power supplies," *IEEE Trans. Electromagn. Compat.*, vol. 47, pp. 921-929, 2005.
- [78] E. Gubia, P. Sanchis, A. Ursua, J. Lopez, and L. Marroyo, "Frequency domain model of conducted EMI in electrical drives," *Power Electronics Letters, IEEE*, vol. 3, pp. 45-49, 2005.
- [79] P. Russer, J. Russer, C. Hoffmann, and H. H. Slim, "Real-time EM spectrum and EM interference measurement techniques," in *Microwave Conference (EuMC), 2013 European*, 2013, pp. 1115-1118.
- [80] Z. Huibin, A. R. Hefner, and L. Jih-Sheng, "Characterization of power electronics system interconnect parasitics using time domain reflectometry," *IEEE Trans. Power Electron.*, vol. 14, pp. 622-628, 1999.
- [81] C. Henglin and Q. Zhaoming, "Modeling and Characterization of Parasitic Inductive Coupling Effects on Differential-Mode EMI Performance of a Boost Converter," *IEEE Trans. Electromagn. Compat.*, vol. 53, pp. 1072-1080, 2011.
- [82] W. Fu and L. Junwei, "High frequency planar transformer with helical winding structure," *IEEE Trans. Magn.*, vol. 36, pp. 3524-3526, 2000.
- [83] I. W. Hofsjager, K. Fricker, and M. O. Crosato, "Reduction of EMI in planar transformers by charge balancing," *Electronics Letters*, vol. 38, pp. 1712-1714, 2002.
- [84] F. Bai, Z.-x. Niu, Y.-J. Shi, and D.-f. Zhou, "EMI Modeling and Simulation of High Voltage Planar Transformer," in *Power Electronics, Drives and Energy Systems, 2006. PEDES '06. International Conference on*, 2006, pp. 1-4.

- [85] C. Ming, S. Xudong, and H. Lipei, "A method to model the conducted EMI propagation path for PFC," in *Power Electronics and Motion Control Conference, 2009. IPEMC '09. IEEE 6th International*, 2009, pp. 674-677.
- [86] F. B. M. Van Horck, A. P. J. Van Deursen, and P. C. T. Van Der Laan, "Common-mode currents generated by circuits on a PCB-measurements and transmission-line calculations," *IEEE Trans. Electromagn. Compat.*, vol. 43, pp. 608-617, 2001.
- [87] J. Yu, Z. Zou, Y. Li, Q. Wang, X. Li, and Y. Zhang, "Numerical Simulation on EMI of Wire," in *Microwave, Antenna, Propagation and EMC Technologies for Wireless Communications, 2007 International Symposium on*, 2007, pp. 1326-1329.
- [88] P. Wilson, *The circuit designer's companion*: Elsevier, 2012.
- [89] N. DasGupta and A. DasGupta, "A new SPICE MOSFET Level 3-like model of HEMT's for circuit simulation," *Electron Devices, IEEE Transactions on*, vol. 45, pp. 1494-1500, 1998.
- [90] A. Karvonen and T. Thiringer, "MOSFET modeling adapted for switched applications using a state-space approach and internal capacitance characterization," in *Power Electronics and Drive Systems, 2009. PEDS 2009. International Conference on*, 2009, pp. 1107-1112.
- [91] D. A. Grant and J. Gowar, *Power MOSFETs: theory and applications*: Wiley New York etc., 1989.
- [92] K. R. A. Britto, R. Dhanasekaran, R. Vimala, and B. Saranya, "EMI analysis and evaluation of an improved flyback converter," in *Computer Communication and Informatics (ICCCI), 2012 International Conference on*, 2012, pp. 1-7.

- [93] M. Miloudi, A. Bendaoud, H. Miloudi, S. Nemmich, and H. Slimani, "Analysis and reduction of common-mode and differential-mode EMI noise in a Flyback switch-mode power supply (SMPS)," in *Telecommunications Forum (TELFOR)*, 2012 20th, 2012, pp. 1080-1083.
- [94] C. McLyman, *Transformer and inductor design handbook*: Marcel Dekker, 2004.
- [95] T. M. Liakopoulos and C. H. Ahn, "3-D microfabricated toroidal planar inductors with different magnetic core schemes for MEMS and power electronic applications," *IEEE Trans. Magn.*, vol. 35, pp. 3679-3681, 1999.
- [96] G. M. Popov, "Quasistatic and dynamic electric and magnetic fields in a coil-ferromagnetic core system," *Measurement Techniques*, vol. 42, pp. 162-168, 1999.
- [97] A. A. Huzayyin, "Utilizing the nonlinearity of a magnetic core inductor as a source of variable reactive power compensation in electric power systems," in *2008 Annual IEEE Student Paper Conference*, 2008, pp. 1-4.
- [98] G. Bertotti, V. Basso, and M. Pasquale, "Application of the Preisach model to the calculation of magnetization curves and power losses in ferromagnetic materials," *IEEE Trans. Magn.*, vol. 30, pp. 1052-1057, 1994.
- [99] E. C. Stoner and E. P. Wohlfarth, "A mechanism of magnetic hysteresis in heterogeneous alloys," *IEEE Trans. Magn.*, vol. 27, pp. 3475-3518, 1991.
- [100] J. H. Blase, "Hysteresis and Catastrophe Theory - Empirical Identification in Transportation Modeling," *Environ Plann A*, vol. 11, pp. 675-688, 1979.
- [101] J. Ostiguy, x, F, and P. P. Silvester, "Magnetic field computation in hysteretic media by the finite element method, using the Preisach model," *Journal of Applied Physics*, vol. 57, pp. 3853-3855, 1985.
- [102] B. E. Kashevskii, S. B. Kashevskii, I. V. Prokhorov, and V. E. Ivanov, "Study of energy dissipation in a suspension of particles with magnetic hysteresis in

- alternating supercritical-frequency field," *Colloid Journal*, vol. 68, pp. 54-58, 2006.
- [103] P. M. B. Silva Girao and J. F. Borges da Silva, "Analog circuit simulation of magnetic dipole behavior according to the Langevin-Weiss theory of ferromagnetism," *Education, IEEE Transactions on*, vol. 35, pp. 153-158, 1992.
- [104] D. C. Jiles and D. L. Atherton, "Theory of ferromagnetic hysteresis," *Journal of Magnetism and Magnetic Materials*, vol. 61, pp. 48-60, 1986.
- [105] I. Klik, C. Chang, x, Ray, and J. Lee, "Master equation approach to anhysteresis of noninteracting particles," *Journal of Applied Physics*, vol. 75, pp. 5487-5489, 1994.
- [106] D. C. Jiles and Y. Melikhov, "Modelling of Nonlinear Behaviour and Hysteresis in Magnetic Materials," in *Handbook of Magnetism and Advanced Magnetic Materials*, ed: John Wiley & Sons, Ltd, 2007.
- [107] G.-y. Li and M.-g. Liu, "The summary of differential evolution algorithm and its improvements," in *Advanced Computer Theory and Engineering (ICACTE), 2010 3rd International Conference on*, 2010, pp. V3-153-V3-156.
- [108] M. Fatih Tasgetiren, O. Bulut, P. Quan-ke, and P. N. Suganthan, "A differential evolution algorithm for the median cycle problem," in *Differential Evolution (SDE), 2011 IEEE Symposium on*, 2011, pp. 1-7.
- [109] S. Chengfu, Z. Haiyan, and C. Liqing, "Improved differential evolution algorithms," in *Computer Science and Automation Engineering (CSAE), 2012 IEEE International Conference on*, 2012, pp. 142-145.
- [110] C. A. Baguley, U. K. Madawala, and B. Carsten, "The Influence of Remanence on Magnetostrictive Vibration and Hysteresis in Mn-Zn Ferrite Cores," *IEEE Trans. Magn.*, vol. 48, pp. 1844-1850, 2012.

- [111] R. C. Smith and Z. Ounaies, "A domain wall model for hysteresis in piezoelectric materials," *Journal of Intelligent Material Systems and Structures*, vol. 11, pp. 62-79, 2000.
- [112] P. D éz, "A note on the convergence of the secant method for simple and multiple roots," *Applied Mathematics Letters*, vol. 16, pp. 1211-1215, 2003.
- [113] R. K. Bawa, "Numerical Methods and Applications," 2009.
- [114] T. Kohama, Y. Mukuno, and H. Shimamori, "Magnetic saturation due to fast dynamic response of isolated DC-DC converter," in *Industrial Technology, 2009. ICIT 2009. IEEE International Conference on*, 2009, pp. 1-5.
- [115] Y. Wei, E. Tam, and S. Y. Hui, "A Semi-Theoretical Fluorescent Lamp Model for Time-Domain Transient and Steady-State Simulations," *IEEE Trans. Power Electron.*, vol. 22, pp. 2106-2115, 2007.
- [116] M. S. Perdigao, E. S. Saraiva, J. M. Alonso, and M. A. Dalla Costa, "Comparative Analysis and Experiments of Resonant Tanks for Magnetically-Controlled Electronic Ballasts," in *Industrial Electronics, 2007. ISIE 2007. IEEE International Symposium on*, 2007, pp. 3041-3046.
- [117] A. J. Holloway, R. C. Tozer, and D. A. Stone, "A Physically Based Fluorescent Lamp Model for a SPICE or a Simulink Environment," *IEEE Trans. Power Electron.*, vol. 24, pp. 2101-2110, 2009.
- [118] Y. Wei and S. Y. R. Hui, "A universal PSpice model for HID lamps," in *Industry Applications Conference, 2002. 37th IAS Annual Meeting. Conference Record of the*, 2002, pp. 1475-1482 vol.2.
- [119] C. F. H. W. L. G. Lama, Thomas J.; Walsh, Peter J., "Analytical model for low-pressure gas discharges: application to the Hg+Ar discharge," *Appl. Opt.*, vol. 21, pp. 1801-1811, May 15, 1982.

- [120] W. Yan, S. Y. R. Hui, H. Chung, and X. H. Cao, "Genetic algorithm optimised high-intensity-discharge lamp model," *Electronics Letters*, vol. 38, pp. 110-112, 2002.
- [121] L. Rossetto, S. Buso, and G. Spiazzi, "Conducted EMI issues in a 600-W single-phase boost PFC design," *IEEE Trans. Ind. Appl.*, vol. 36, pp. 578-585, 2000.
- [122] L. Ray-Lee, L. Hung-Yi, and S. Hsu-Ming, "AC-Side CCM CS-CP-PFC Electronic Ballast," *IEEE Trans. Power Electron.*, vol. 22, pp. 789-796, 2007.
- [123] L. Shu Fan and A. M. Khambadkone, "A simple digital DCM control scheme for boost PFC operating in both CCM and DCM," *IEEE Trans. Ind. Appl.*, vol. 47, pp. 1802-1812, 2011.
- [124] K. Yao, X. Ruan, X. Mao, and Z. Ye, "DCM boost PFC converter with high input PF," in *Applied Power Electronics Conference and Exposition (APEC), 2010 Twenty-Fifth Annual IEEE*, 2010, pp. 1405-1412.
- [125] T. Grote, H. Figge, N. Frohlike, J. Bocker, and F. Schafmeister, "Digital control strategy for multi-phase interleaved boundary mode and DCM boost PFC converters," in *Energy Conversion Congress and Exposition (ECCE), 2011 IEEE*, 2011, pp. 3186-3192.
- [126] C. Hangseok and L. Balogh, "A cross-coupled master-slave interleaving method for boundary conduction mode (BCM) PFC converters," *IEEE Trans. Power Electron.*, vol. 27, pp. 4202-4211, 2012.
- [127] Y. Sung-Pei, C. Shin-Ju, and H. Chao-Ming, "Small-signal modeling and controller design of BCM boost PFC converters," in *Industrial Electronics and Applications (ICIEA), 2012 7th IEEE Conference on*, 2012, pp. 1096-1101.
- [128] H. S. Athab and D. D. C. Lu, "Simple controller for single-phase power factor correction rectifier," *Power Electronics, IET*, vol. 3, pp. 590-600, 2010.

- [129] D. Xiong, Z. Luowei, and T. Heng-Ming, "Average current control of a series-type single-phase PFC with hybrid modulation," *IEEE Trans. Power Electron.*, vol. 26, pp. 2381-2385, 2011.
- [130] J. Luo, M. K. Jeoh, and H. C. Huang, "A new continuous conduction mode PFC IC with average current mode control," in *Power Electronics and Drive Systems, 2003. PEDS 2003. The Fifth International Conference on*, 2003, pp. 1110-1114 Vol.2.
- [131] X. Xiaojun, L. Wei, and A. Q. Huang, "Two-phase interleaved critical mode PFC boost converter with closed loop interleaving strategy," *IEEE Trans. Power Electron.*, vol. 24, pp. 3003-3013, 2009.
- [132] K. Raggl, T. Nussbaumer, G. Doerig, J. Biela, and J. W. Kolar, "Comprehensive design and optimization of a high-power-density single-phase boost PFC," *IEEE Trans. Ind. Electron.*, vol. 56, pp. 2574-2587, 2009.
- [133] J. Balcells, A. Santolaria, A. Orlandi, D. Gonzalez, and J. Gago, "EMI reduction in switched power converters using frequency Modulation techniques," *IEEE Trans. Electromagn. Compat.*, vol. 47, pp. 569-576, 2005.
- [134] R. Morrison and D. Power, "The Effect of Switching Frequency Modulation on the Differential-Mode Conducted Interference of the Boost Power-Factor Correction Converter," *IEEE Trans. Electromagn. Compat.*, vol. 49, pp. 526-536, 2007.
- [135] S. Y. R. Hui and J. Zhu, "Numerical modelling and simulation of hysteresis effects in magnetic cores using transmission-line modelling and the Preisach theory," *Electric Power Applications, IEE Proceedings* -, vol. 142, pp. 57-62, 1995.

- [136] M. L. Heldwein, J. Biela, H. Ertl, T. Nussbaumer, and J. W. Kolar, "Novel three-phase CM/DM conducted emission separator," *IEEE Trans. Ind. Electron.*, vol. 56, pp. 3693-3703, 2009.
- [137] T. Nussbaumer, K. Raggl, and J. W. Kolar, "Design guidelines for interleaved single-phase boost PFC circuits," *IEEE Trans. Ind. Electron.*, vol. 56, pp. 2559-2573, 2009.
- [138] S. Fu-Yuan, D. Y. Chen, W. Yan-Pei, and C. Yie-Tone, "A procedure for designing EMI filters for AC line applications," *IEEE Trans. Power Electron.*, vol. 11, pp. 170-181, 1996.
- [139] N. P. Papanikolaou and E. C. Tatakis, "Minimisation of power losses in PFC flyback converters operating in the continuous conduction mode," *Electric Power Applications, IEE Proceedings* -, vol. 149, pp. 283-291, 2002.
- [140] STP5NK80ZFP Datasheet (PDF) - STMicroelectronics - N-channel 800V - 1.9 Ω - 4.3A - TO-220/TO-220FP Zener-protected SuperMESH™ Power MOSFET [Online]. Available: <http://pdf1.alldatasheet.com/datasheet-pdf/view/208563/STMICROELECTRONICS/STP5NK80ZFP.html>
- [141] W. Laili, P. Yunqing, Y. Xu, Q. Yang, and W. Zhaoan, "Improving light and intermediate load efficiencies of buck converters with planar nonlinear inductors and variable on time control," *IEEE Trans. Power Electron.*, vol. 27, pp. 342-353, 2012.