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**NANYANG
TECHNOLOGICAL
UNIVERSITY**

**MODELING AND ANALYSIS OF
ELECTROMAGNETIC INTERFERENCE RADIATION FOR
BROADBAND POWER LINE COMMUNICATION
NETWORKS**

PANG TENG SENG

**SCHOOL OF ELECTRICAL AND ELECTRONIC ENGINEERING
2011**

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ELECTROMAGNETIC INTERFERENCE RADIATION
FOR BROADBAND POWER LINE COMMUNICATION
NETWORKS**

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SUMMARY

Power Line Communications (PLC) is a term used to identify technologies, equipment, applications, and services aiming at providing users with communication means over the existing power lines. With the increasing demands for broadband access and home networking, new technologies are being created to tap into this growing broadband communication market. Among them, PLC technology has recently attracted lots of attention worldwide due to its convenience of wide availability and advantage of low cost. Any power socket in the home and office is a potential broadband access point and there is no need to hack the walls to run the communication cables again.

However, like all other technologies, PLC also faces its own set of obstacles and technical challenges. The power line is designed to transmit electrical power at low frequency, and it appears as a harsh environment for the high frequency communication signals. Electromagnetic interference (EMI) radiation from PLC networks is a major concern for the widespread use of broadband PLC technology. As the power line is not shielded, it behaves like an antenna causing high radiation when high frequency PLC signals are transmitted.

Although it is well known that the radiation mechanism from the PLC network is common mode in nature, no comprehensive study has been carried out to characterize and understand the behavior of CM current in the PLC network. Such a study is important as it provides insight of radiation emissions from the PLC network. In this thesis, a complete current propagation model for a broadband PLC network is proposed. A complete current propagation model, including of a differential-mode (DM) current and common-mode (CM) current propagation model for a three-wire power line cable, is developed using the Multi-conductor Transmission Line (MTL) theory. The model is used to study the EMI radiation mechanism from the PLC network. The accuracy of the model is verified through numerical simulations and practical measurements conducted on the actual power line network. The developed model will allow us to predict the DM

and CM currents in the PLC network with reasonable accuracy. With this model, we can extend it to model more complicated PLC networks and it serves as a valuable tool to identify effective ways to suppress EMI radiation from PLC networks.

As such, a new signal injection technique is proposed by utilizing the additional Earth wire that is present in the electrical system to reduce the radiation level for the PLC network. The typical method of signal injection is done by injecting the signal into the Live and Neutral wires. What we propose here is to inject the signal into the Live and Earth wires and another complementary signal into the Neutral and Earth wires. By doing this, the effective CM currents flowing in the power lines are reduced through the cancellation of the electromagnetic fields. The proposed signal injection technique is also compared firstly with the typical signal injection technique that is used in PLC modems nowadays, and secondly with the alternative signal injection technique that is proposed by other authors. From the comparison, it is concluded that the proposed new signal injection technique can produce the best results in terms of mitigation of electromagnetic radiation. The practicality and feasibility of utilizing the Earth wire in the injection of PLC signals is also discussed.

Lastly, we also conduct a series of experiments on commercial PLC modems to identify the possible CM coupling paths. Analysis of the commercial PLC modems shows that parasitic capacitances of the power transformer and isolation transformer provide low impedance paths for the high frequency noise to couple to the powerlines and induce CM currents. Based on the analysis, solutions are implemented on the modems to prevent high frequency noise from coupling to the powerlines and inducing CM currents. In the studies, the power transformer is removed from each modem and a CM choke is added in series with the isolation transformer. Both the solutions lower the CM current and emissions by around 20dB.

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LIST OF ABBREVIATIONS

(Alphabetical Sequence)

AFE:	Analog Front End
CM:	Common Mode
CW:	Continuous Wave
DBPSK:	Differential Binary Phase Shift Keying
DM:	Differential Mode
DQPSK:	Differential Quadrature Phase Shift Keying
DSL:	Digital Subscriber Line
EM:	Electromagnetic
EMA:	Energy Market Authority
EMC:	Electromagnetic Compatibility
EMI:	Electromagnetic Interference
GDP:	Gross Domestic Product
HDB:	Housing Development Board of Singapore
ICT:	Info-Communications Technology
IDA:	Infocomm Development Authority of Singapore
iN2015:	Intelligent Nation 2015
IP:	Internet Protocol
LAN:	Local Area Network
LCL:	Longitudinal Conversion Loss
LISN:	Line Impedance Stabilization Network
LN:	Live and Neutral
LNE:	Live, Neutral and Earth
LV:	Low voltage
LVDN:	Low Voltage Distribution Network
MTL:	Multi-conductor Transmission Line
OFDM:	Orthogonal Frequency Division Multiplexing
PC:	Personal Computers

List of Abbreviations

PLC:	Power Line Communications
PLM:	Power Line Modem
PVC:	Polyvinyl Chloride
RF:	Radio Frequency
R&D:	Research and Development
RegTP:	Regulating Administration for Telecommunications and Posts
Rx End:	Receiving End
TEM:	Transverse Electromagnetic
Tx End:	Transmission End

Chapter 1

INTRODUCTION

1.1 MOTIVATIONS

Over the past decade, due to the advances in technology and infrastructure, computers and Internet access have become an indispensable part of the residential, educational and industrial communities. The Internet has changed how people interact, play, work and learn; it has changed peoples' lives. People go online nowadays to download information, correspond with friends, play games, listen to music, watch videos, purchase stuff, do banking and even make an income out of it. At the year end of 2007, the total world Internet users was estimated to be 1.32 billion (See Figure 1.1) and the average world growth in the Internet users in the world between year 2000 and year 2007 is around 265% [1] (See Figure 1.2). From these data, we can see the importance and the widespread use of the Internet.

Figure 1.1. Internet users in the world for December 2007 [1].

Figure 1.2. Growth in Internet users in the world between year 2000 and year 2007 [1].

Fast Internet access is also paramount for users to remain competitive in this fast-paced environment, especially for a small country like Singapore without any natural resources. Singapore has a high level of adoption of broadband technology. According to the annual surveys conducted on the info-communication usage by Singaporeans by the Infocomm Development Authority (IDA) of Singapore, there is a significant growth in the use of broadband Internet access over the years [2]. As shown in Figure 1.3, these surveys show that the home computer penetration has increased from a mere 11% in the year 1988 to 78% in the year 2006. The percentage of home Internet penetration has also increased from only 9% in the year 1996 to 71% in the year 2006. This means that in the year 2006, nearly three out of four Singaporean households own at least a computer and they have access to the Internet. As for broadband Internet access, there is a rising trend since it came into the scene from 33% in the year 2003 to 77.1% in the year 2007 and the figure is expected to continue rising as shown in Figure 1.4. With the appearance of broadband internet access, there is a downward trend in the narrowband market which shows a decline from 48% in the year 2002 to 23% in the year 2007. This downward trend in the narrowband market is due to the fact that more clients are shifting from narrowband to broadband, thus showing an increasing need for broadband Internet access.

Figure 1.3. Household info-communication penetration in Singapore (Computer and Internet) between 1998 and 2006 [2].

Figure 1.4. Household info-communication penetration in Singapore (Narrowband and Broadband) between 1999 and 2007 [2].

Singapore has a vibrant and dynamic info-communications technology (ICT) industry. The strong growth in this sector is driven by key government initiatives. The ICT sector is regulated by iDA. Despite growing competition, Singapore's ICT industry continues to advance, raking in revenues of S\$45.2 billion in 2007, as shown in Figure 1.5. This figure is a near 20% rise from the year 2006. In 2006, the industry revenues stood at S\$37.9 billion, a 9% increase from S\$34.8 billion in 2005. Today, the ICT industry contributes over 6% of Singapore's Gross Domestic Product (GDP). On average, two out of every three business Research and Development (R&D) are ICT projects [3].

Singapore has developed an advanced and reliable telecommunications infrastructure, and its government is strongly committed to making Singapore one of the most connected

cities in the world. Over the years, the Singapore government has been promoting the usage of broadband Internet access, as part of its Intelligent Nation 2015 (iN2015) master-plan. This iN2015 plans to transform Singapore into an intelligent nation and global city, powered by info-communication, by 2015. The master-plan recommends the way forward for Singapore, into a future where info-communication will bring about changes to the way people live, learn, work and play.

Figure 1.5. Total ICT Industry Revenue 1998 – 2006, Annual [3].

The iN2015 master-plan outlined four key strategies in which the first is to leverage existing network readiness, capabilities for wireless pervasiveness, develop contents and promote info-communication literacy. The second is to develop Singapore as a leading, global and digital distribution as well as a trading centre. The third is to grow new economic activities and create jobs in the info-communication industry through developing innovation capabilities, encouraging technology development and expanding access to overseas markets. The fourth is to help businesses and government agencies use info-communication to re-engineer key business clusters and government services.

In harnessing info-communication technologies for the key economic sectors, some key recommendations include the use of personalised services to enhance healthcare, education, tourism and e-government; seamless delivery of financial services and supply-chain management. The Next Generation National Info-communication Infrastructure will be put in place by 2012, capable of delivering broadband speeds up to 1 Gbps, and offer pervasive connectivity around the country. The infrastructure will also be Internet

Protocol version 6 (IPv6) compliant and will enable an exciting host of new broadband-enabled services and applications. This includes applications such as immersive learning experiences, telemedicine, high definition TV, immersive video conferencing and grid computing. There are many technologies to achieve this, using either wired or wireless methods. Among them, Power Line Communications (PLC) has been identified as a potential candidate.

1.1.1 Broadband Power Line Communication Systems

Power Line Communications (PLC) refers to the technology that uses the existing electrical power supply networks for the communications purposes. The system can transmit voice and data by superimposing modulated analogue signals over the standard 50 or 60 Hz alternating current. PLC technology can be classified into two categories and they are narrowband PLC and broadband PLC. For narrowband PLC, two standards are established and they are the Europe CENELEC Standard EN50065-7 which makes use of the 3 to 148.5 kHz bandwidth and the USA standard which makes use of the 45 to 450 kHz bandwidth [4]. They operate at low data rates and low power and the main use of this technology is in the area of automatic meter reading, lighting control, tariff control of electricity and load management. Broadband PLC, also known as power line Internet, operates at high frequencies in the range of 1 to 30 MHz in which 1 to 10 MHz is the access band and the in-house band is from 10 to 30 MHz [5]. The main use for broadband PLC is to provide high-speed, broadband Internet services to homes and offices. Any computer or communication device would only need to plug a PLC modem into any outlet in an equipped building to gain access to high-speed Internet access. Broadband PLC offers advantages over other regular cable or digital subscriber line (DSL) connections in such a way that “no new wires” are required due to the availability of extensive power line infrastructure. Although wireless technologies allow flexibility of adding or moving workstations and providing connectivity in areas where it is difficult to lay cables, they are greatly affected by surroundings where concrete walls in buildings absorb the microwave resulting in total loss of signals. PLC does not face the above

problem and it is especially useful in the basements of buildings where reception from wireless technologies can be totally blocked off. Also, when it is impractical to have wireless technologies in rural areas where the population density is low and houses are very far apart, PLC serves as good alternative. As long as users have access to power supply, they will be accessible to the Internet as well. This is the most attractive benefit of PLC. PLC will definitely be the best candidate for applications in future smart homes or buildings where all the home electronic devices can be remotely controlled by a single command without having to worry about communication links that cannot be established due to the virtue of their positions as in the case of wireless technologies

1.1.2 History and Development of PLC

PLC has a history of around 170 years. It dates back to 1838 where the first remote electricity supply metering was proposed by Edward Davy [6] for the purpose of checking the voltage levels of batteries at unmanned sites on the London - Liverpool telegraph system. In 1897, Joseph Routin and C. E. L. Brown of Zurich, Switzerland took out British Patent No. 24833 on their power line signalling electricity meter [7]. In 1905, Chester Thoradson of Chicago patented his system, devised in 1902 for the remote reading of electricity meters [8]. The first commercial production of electromechanical repeaters took place in 1913 [9]. The applications then were of narrow band and they involved only energy meter reading, and home automation. By late 1980s, relatively sophisticated error correction techniques within the hardware of PLC modems were proposed [9]. And more applications of PLC for home networking only increase in the recent years with the development of the HomePlug Powerline Alliance [10].

The HomePlug Powerline Alliance is an industry-led initiative that brings necessary capabilities and a financial commitment to the successful launch and implementation of the technology. It includes many famous companies like Motorola, Sharp, Texas Instruments, Intel, Intellon, etc. Together, these companies evaluate the technology and create a specification that allows the use of the power outlets and electrical wires to

connect devices to each other and the Internet. Before the formation of the alliance, there was no recognized industrial standard. It was only in the June of 2001 that the HomePlug 1.0 specification was released and since then, it is the only globally recognized standard for high-speed powerline networking with millions of products sold worldwide. Currently, HomePlug Alliance is working on a new specification, the HomePlug AV (Audio/Video), which is the next generation of powerline technology. It specifies data rates up to 200 Mbps and can support entertainment applications such as High Definition Television (HDTV) and Home Theatre.

1.1.3 Electromagnetic Compatibility Issues in Broadband PLC Systems

PLC technology is an attractive technology as it has the largest existing infrastructure in the world as the medium for transmission of data is through the existing power line cables available. Wherever there is power available, access to PLC technology is possible. Broadband Internet connection can be accessed in every socket in every room with a simple plug-and-play installation. There is no need for any changes in the current business or household wirings and this means less cost for installation. The advantages of this technology make it an attractive alternative to conventional broadband technologies and it is gaining lots of attention in Europe, Asia and the United States as utilities and telcos are seeking cost-effective way to bring broadband to consumers. Singapore is one of the countries at the forefront of implementing broadband PLC technologies. A one-year field trial from November 2000 to October 2001 was conducted by SP Telecom, the telecommunications department of Singapore Power Ltd, within the Singapore Polytechnic campus [11], [12]. The field trials had demonstrated that PLC was capable to function effectively and efficiently without compromising the power supply and telecommunications service standards.

Although PLC has the advantage of “no new wires” and being cost-effective in its implementation, the power line network is designed for the transmission of power, and not for communication means. The power line network is characterized as unbalanced

signal lines with mismatches due to stubs, switches, junctions, outlets, etc. Due to the unbalance nature of the power line network, a high level of electromagnetic radiation is generated when high frequency broadband PLC signals are transmitted through the network. This high level of electromagnetic radiation is a concern for many regulatory bodies. Many field trials in the United States, Europe and other Asian countries, in the frequency range between 1-30 MHz have shown significant level of radiated emissions from the PLC networks [13] – [19]. In Singapore, the Infocomm Development Authority (IDA) and the Energy Market Authority (EMA) are bothered by the potential risk of interference to other radio communications and wireless services. With the electromagnetic radiation issue unresolved, the full-scale commercial deployment of broadband PLC would be delayed [20].

1.1.4 EMC Standards and Regulations for PLC Systems

The power line cables used for PLC are designed and installed for the transmission of power and not for communication purposes. These cables are usually untwisted and unshielded and they behave like long antennas radiating huge amount of electromagnetic radiation when data are transmitted at high frequencies. Being unshielded, the PLC systems are also susceptible to interference from other electromagnetic interference (EMI) sources. Electromagnetic compatibility (EMC) of wireless services and PLC is a bi-directional problem where electromagnetic radiation from one network affects the other. In order to achieve compliance in EMC between PLC and other radio services, there must be limitation of radiated fields from PLC systems.

Many international standards have been developed to control radiated emissions below 30 MHz in the short-wave user spectrum. There are different standards being proposed and different countries adopt different standards as EMC regulations developed for one country may not be suitable for the other. Figure 1.6 shows the electric field limits adopted by different countries. The Regulating Administration for Telecommunications and Posts (RegTP) in Germany adopted the NB30 [21] for the radiation of telecommunications services in the frequency range of 9 to 3 GHz, while the Radio

Communications Agency in the United Kingdom has defined MPT 1570 [22] as its current standards for telecommunication systems operating in the frequency range of 9 to 30 MHz. It is observed that the limits set by UK's MPT 1570 are about 20 dB lower than that of the German RegTP's recommendation. Such a strict regulation posed a difficulty for the widespread use of the PLC technology in the UK. United States adopted FCC Part 15 [23] regulations for its PLC systems. With a more relaxed electric field limits regulation in the United States, it is not surprising to see that the development of high-speed PLC equipment is more progressive in the United States as compared to anywhere else in Europe. This is the reason why the HomePlug Powerline Alliance is adopting the FCC Part 15 as their EMI regulatory standards.

Figure 1.6. Electrical field level limits for radiation for wire-bound communication for three different countries [20], [21].

1.2 OBJECTIVES

It is well known that the radiation mechanism from the PLC network is common mode (CM) in nature, however, no comprehensive study has been carried out to characterize and understand the behaviour of CM current on the PLC network. Such a study is important as it provides insight of radiation emissions from the PLC network. This research aims to develop an EMI model for broadband PLC networks so that the EMI radiation mechanism can be properly studied and analyzed. A complete radio frequency current propagation model for a three-wire power line cable using the Multi-conductor Transmission Line (MTL) theory, using a bottom-up approach, is proposed. The complete model consists of differential mode (DM) and common mode (CM) current propagation models which allow us to predict the DM and CM currents in the power lines. These models will serve as valuable tools in the future to suppress EMI radiation from PLC networks.

With better understanding of the radiation mechanism, we explore new signal injection techniques so as to achieve lower radiation levels. We also study the power line modems (PLMs) in greater detail so as to find new methods to mitigate the EMI radiation level.

1.3 MAJOR CONTRIBUTIONS OF THIS THESIS

The major contributions of the research can be summarized as follows:

1. The idea of using MTL theory to model the power lines in broadband PLC networks is proposed. The DM and CM current propagation models for a three-wire power line cable in broadband PLC networks are derived based on a bottom-up approach. The accuracies of the proposed models are verified through numerical simulations and actual measurements. With these models, we can predict the levels of DM or CM voltages and currents at the load end once we know the levels of the DM or CM voltages and currents at the source end and vice versa for different lengths of the power line cable.

2. A new signal injection technique that utilizes the Earth wire is proposed to achieve the cancellation of electromagnetic fields so as to reduce the EMI radiation from broadband PLC networks. As compared to the alternative injection technique (which produced a maximum of 20 dB reduction in emission level) proposed by other authors and the normal injection technique, we can achieve great reduction (up to a maximum of 40 dB) in the radiation for the balanced case and slight reduction in the radiation for the unbalanced case. Although our new injection technique and the alternative injection technique can both achieve reduction in the level of radiation, our injection technique results in a lower level of CM current in the Earth wire. With a lower CM current in the Earth wire, the resultant radiation emitted from the actual system may be lower. Also, our new injection technique proved to be a safer technique as we are using the Earth wire as the reference conductor.
3. The PLMs are studied in detail and different CM coupling paths are identified. The parasitic capacitances between the primary and secondary windings of the power transformer and isolation transformer of the PLM make CM coupling paths possible. It is found that the CM impedances for the power transformer and isolation transformer are around 100 k Ω and 1000 k Ω respectively. With lower CM impedance, the power transformer allows more CM current to flow through and hence resulting in higher radiation level. We manage to bring about an average reduction of 10 dB for conducted emission and 8 dB for radiated emission by implementing a CM choke in series with the power transformer.

1.4 ORGANISATION OF THIS THESIS

There are 7 chapters in this thesis. In Chapter 1, the introduction of the research work, including the motivations, objectives and contributions of this research work, as well as the organisation of this thesis, is presented. Chapter 2 is the literature review, where previous works of other researchers are presented.

Chapter 3 presents the theoretical background on the Multi-conductor Transmission Line (MTL) theory where the complete current propagation model in a broadband PLC network is based. Chapter 4 is devoted to the complete modelling of the broadband PLC network. Here, a complete current propagation model, inclusive of DM current and CM current propagation models for a three-wire power line cable, is developed using the MTL theory. The model is used to study the EMI radiation mechanism from the PLC network. The accuracy of the model is verified through numerical simulations and practical measurements conducted on the actual power line network. The developed model will allow us to predict the DM and CM currents in the power line with reasonable accuracy. With this model, we can extend it to model more complicated PLC networks and it serves as a valuable tool to identify effective ways to suppress EMI radiation from PLC networks.

In Chapter 5, we propose a new signal injection technique by utilizing the Earth wire that is present in the electrical system to reduce the radiation level for the PLMs. The typical method of signal injection is done by injecting the signal into the Live and Neutral wires. What we proposed here is to inject the signal into the Live and Earth wires and another complementary signal into the Neutral and Earth wires. By doing this, we are trying to lower the effective CM current flowing in the power line cable through the cancellation of the electromagnetic fields. We also compare our method of signal injection technique with the typical signal injection technique that is used in PLMs these days. Also, we compare our signal injection technique with the alternative signal injection technique that was proposed by other authors. Upon comparison, we conclude that the new signal injection technique proposed by us produces the best results in terms of mitigation of electromagnetic radiation. We also discuss the practicality and feasibility of utilizing the Earth wire in the injection of PLC signals.

In Chapter 6, the generation of CM current over the power line network due to coupling from source is investigated in detail by means of experiments. Analytical studies conducted on the commercial PLMs reveal that high frequency noise generated at the source can propagate to the power lines especially at high frequency via low impedance

parasitic capacitances and induce significant CM currents which in turn give rise to high levels of radiated emissions. They also show that parasitic capacitances of the power transformer and isolation transformer provide low impedance paths for the high frequency noise to couple to the power lines and induce CM currents. Based upon the analysis, solutions are implemented on the modems to prevent high frequency noise from coupling to the power lines and inducing CM currents. In the studies, power transformer is removed from each modem and a CM choke is added in series with the isolation transformer. Both the solutions lower the CM currents and emissions by around 15 dB.

Finally, conclusions of the research work and recommendations for future work are presented in Chapter 7.

Chapter 2

LITERATURE REVIEW

The risk of EMI to radio communications and wireless services [13], [14] caused by the generation of electromagnetic fields from the power lines is a major concern that prevents the widespread use of broadband PLC systems. There have been many field trials with different broadband PLC systems running worldwide in the recent years to study the feasibility of using the low-voltage power distribution networks for broadband Internet access as well as to quantify the interference potential in the frequency range of 1 to 30 MHz. The measurement data collected in [15], [16] have shown significant levels of radiated emissions from PLC networks which are beyond that of the accepted level [24].

2.1 SOURCES OF EMI

Electromagnetic disturbances can appear in the form of “differential mode” (DM) (also called “symmetrical mode”) and “common mode” (CM) (also called “asymmetrical mode”) voltages and currents. The definitions of the DM and CM are shown in Figure 2.1. The components of these modes are defined by the voltages and currents, measured on the mains terminals, and are expressed as follows [24]:

$$DM \text{ voltage}, V_d = V_1 - V_2 \quad (2.1)$$

$$DM \text{ current}, I_d = \frac{I_1 - I_2}{2} \quad (2.2)$$

$$CM \text{ voltage}, V_c = \frac{V_1 + V_2}{2} \quad (2.3)$$

$$CM \text{ current}, I_c = \frac{I_1 + I_2}{2} \quad (2.4)$$

Figure 2.1. Model of a typical EMI source and its DM and CM voltages and currents.

The general model of an EMI source is illustrated in Figure 2.1. The EMI source injects two types of currents into the mains network, one of which is the DM current I_d and the other of which is the CM current I_c . Whenever we inject a current signal in a wire, it reacts as an antenna radiating an electromagnetic (EM) field into the environment. The source generates a DM current into the network which results in the first EM field, and the other DM current with the same magnitude as the first one but in the opposite direction. This second DM current also generates an EM field with the same magnitude as the first one but in the opposite direction. As a result of symmetry, the generated EM fields cancel out each other and so no EM disturbance from the DM current can propagate in the environment. Opposite to the differential mode, the current signal in the common mode flows in the same direction. Therefore the resulting EM fields are propagating in an asymmetrical mode, and the total field radiated in the environment is the superposition of these two fields. For this reason, the main cause of the EM field radiation from the PLC network is due to the presence of the CM currents.

2.2 COUPLING FACTOR

Many researchers have tried to characterize the radiated emissions associated with PLC networks, ranging from theoretical characterization to experimental assessment. Tests have been carried out to investigate the relationship between the levels of data signals injected into the PLC network and the resulting levels of radiated emissions [16]-[18]. A

coupling factor [25], [26], which is a transfer function, has been proposed to define such a relationship.

Figure 2.2 shows the setup for PLC radiated field measurement using a magnetic loop. The coupling factor is a function of the magnetic field $H(f)$ and the injected voltage $U_{inj}(f)$ given by

$$K_H (dB) = 20 \log \left(\frac{H(f)}{U_{inj}(f)} \right) \quad (2.5)$$

Figure 2.2. A setup for PLC radiated field measurement using a magnetic loop.

Given the voltage injected during the transmission over a PLC network, the associated magnetic field radiated by the network can be calculated by the coupling factor. Then the radiated electric field can be easily found from the magnetic field H and the free space impedance ($Z=377\Omega$) as follows:

$$E_{PLC} (V/m) = Z \cdot H \quad (2.6)$$

Measurements of the magnetic field were conducted in different areas in order to define the coupling factor for various configurations, in-house and outdoor. The results were

very similar for some installations but very different for the other ones [27]. Therefore, it is not possible to use a unique coupling factor in the standardization. Different characterizing coupling factors could be defined, according to the network configurations, the environment locations, the power line parameters and so on. This method of the characterization of electromagnetic fields could only provide an estimation of the level of EMI radiation emitted by an arbitrary input signal in the broadband PLC network and was not able to provide further useful information. Also, direct measurements of electromagnetic fields could only determine or confirm the existence of EMI radiation emitted from the broadband PLC network but could not provide insight into the radiation mechanism of the broadband PLC network.

2.3 LONGITUDINAL CONVERSION LOSS (LCL)

As the radiated fields from the PLC networks are caused by the asymmetrical part (or the CM) of the signals transmitted over the power lines, other investigations were proposed to directly measure this asymmetrical part of the signals and to deduce from it the strength of the radiated fields [28]. In other words, it is important to determine the amount of the DM signal (transverse signal) and what is converted into a CM signal (longitudinal signal). This property, expressed by the Longitudinal Conversion Loss (LCL) value, can be measured as the ratio between a symmetrical (DM) signal injected into a network and a portion of this signal being converted to an asymmetrical (CM) signal, and it describes how balanced or unbalanced a device or a network is. The LCL is a logarithmic ratio between the asymmetrical voltage E_L and the symmetric voltage V_T according to the following relation [29]:

$$LCL(dB) = 20 \log \left(\frac{E_L}{V_T} \right) \quad (2.7)$$

A special probe, called the Macfarlane probe [29], has been designed and proposed for measuring the LCL value of existing in-house low-voltage lines. This probe has been

used widely [30], [31] for characterization of in-house power lines for future EMI regulatory and PLC modem design considerations.

However, due to the fact that the “transmission lines” of the low voltage network consist of individual wires not bundled together to form a cable, the transverse impedance of this transmission line may vary significantly along the line. The equipment connected to the low voltage distribution network will not be matched to the average characteristic impedance of the low voltage distribution network. In addition, both the CM impedance of the power line and the CM impedance of the probe can show significant variations. Both the variations in transverse impedance and in CM impedance have significant influence on the measured LCL value. It is therefore concluded that the LCL method is not suited to determine the amount of mode conversion on the low voltage distribution network.

Given the above mentioned difficulties in determining the CM current indirectly by means of LCL, it seems recommendable to determine the CM current by means of a current probe.

2.4 POWER LINE NETWORK IN SINGAPORE

Figure 2.3 shows a typical distribution system in Singapore where the outdoor low voltage distribution network (LVDN) system supplies about 5 to 6 high-rise buildings, each having around 100 households, in a ring configuration. If PLC system would be implemented in Singapore, it could be implemented at the low-voltage (LV) output of the distribution transformer in the LVDN, thus transforming the LVDN into a large-scale local area network (LAN). In such system, the local distribution transformer substation would be connected to the telecommunication network or the Internet Protocol (IP) backbone via conventional telecommunications infrastructure as shown in Figure 2.4. A PLC master modem in the distribution transformer substation would provide the high-speed connection from the telecommunication backbone onto the LVDN. Filtering devices could be implemented at the substation to prevent the 50 Hz AC current from

leaking into the station communication system, and to provide isolation to ensure that distant faults do not affect the PLC system. PLC system could also be implemented within the intra-building scope as a 'last mile' solution where the PLC range would be limited to the indoor LVDN from the main distribution board of any residential or industrial building as shown in Figure 2.5.

Figure 2.3. Simplified structure of the Low Voltage Distribution Network in Singapore.

Figure 2.4. PLC system for the Low Voltage Distribution Network.

Figure 2.5. Intra-building PLC system for the indoor Low Voltage Distribution Network.

Figure 2.6. A four-room HDB residential unit in Singapore.

Figure 2.6 illustrates the layout of a four-room residential unit developed by the Housing Development Board of Singapore (HDB), which is the typical residential units in Singapore, along with the location of the power outlets. The structure of the network is single-phased and is divided into rings or loops configuration. There are basically three active loops within the network, which are defined as Loop A, Loop B and Loop C and each loop is connected to different electrical appliances through the power outlets that are located within each loop. Loops A and B have almost similar wiring perimeters with the estimated values of 22 m and 23 m, respectively, except for Loop C, which is smaller in wiring perimeter with an estimated value of 16 m. The three loop networks are connected to the external 230V AC power supply through an AC distribution board and an electricity meter, which effectively filters off the high frequency signals and noises coming from the neighbouring residential units and thus isolating the network from the effects of the varying external electrical loads from the neighbouring units.

2.5 CHANNEL CHARACTERIZATION AND MODELING APPROACH

There have been numerous studies on the characterization of propagation characteristics and performance of the LV networks for the use of PLC technology [32]-[42]. Impedance, attenuation and noise are identified as the three critical channel parameters that are of major importance for the design of wire-bounded transmission systems [33]. There are two main effects of the propagation of the PLC signals along the power cables [24]:

- Low pass characteristic:
Cables have a strong low pass characteristic depending on the type of cable, the length of the cable and the frequency of the signal. The low pass behaviour is essentially caused by dielectric losses in the insulation of the cable and is not present in overhead lines. The low pass characteristic of access networks limits the maximum distance over which PLC signals can be transmitted.
- Frequency selective fading:
Due to branches and multiple reflection points, the signal not only propagates on the direct connection between transmitter and receiver, but additional propagation paths have to be considered. Those usually have longer lengths and cause time-delayed echoes. This results in multi-path signal propagation with frequency selective fading [24]. In some extreme cases, a set of frequencies may be wiped out completely. Another result of multi-path signal propagation is an extension of the impulse response of the channel, which might impair the functionality of PLC systems with high data rates, as generally the duration of the impulse response of the channel is greater than the duration of the transmitted signal.

In power networks, there exist different types of noise which additively and independently overlay each other [43]:

- Colored background noise
- Narrow-band noise

- Periodic impulsive noise asynchronous to the mains frequency
- Periodic impulsive noise synchronous to the mains frequency
- Asynchronous impulsive noise

Theoretical models for propagation patterns and noise are extensively discussed in [24] and [43]. Low voltage electrical cabling does not represent a perfect data transmission medium. Indeed, being designed for frequency of 50 or 60 Hz, its impedance and attenuation characteristics in the MHz region used for broadband PLC display high variability, not only when networks from different countries are considered, but also when two different, even neighboring networks are considered. Besides a relatively high attenuation, multiple reflections from different branches make the modeling of the channel characteristics especially difficult [32]. Perhaps the most difficult problem that remains to be solved is the fact that the network characteristics can vary strongly as a function of time, since new appliances can be plugged in and out or turned on and off at any time and they present, in the general case, nonlinear impedances to the network.

When it comes to modeling, the factors that determine the reliability and accuracy of the model are the model parameters and the modeling algorithms. There are generally two types of approaches to obtain the model parameters namely the top-down approach and the bottom-up approach [32]. When the model parameters are obtained from measurements [33]-[38], it is called the empirical approach or the top-down approach. Usually this form of approach requires little computation and is easy to implement. However, since the parameters depend mainly on the measurement results, the model is prone to measurement errors. On the other hand, the analytical approach is based on the theoretical derivation of the model parameters [39]-[41], and it is called the bottom-up approach. Using this approach, the parameters are derived according to the physical properties of the power line, including the properties of its conducting material, its insulating dielectric material and its orientation. Although this approach requires more computational efforts compared to the empirical approach, it can describe clearly the relationship between the network behavior and the model parameters. Moreover, this modeling approach is more versatile and flexible since all the parameters are formulated,

making it easy to predict the changes in the model should there be any change in the system configurations.

For modeling algorithms, they can be classified into time-domain modeling and frequency-domain modeling. In the time-domain modeling, the power line is regarded as a multipath environment and an echo model is developed to represent this physical characteristic [32], [33], [34], [39]. This modeling is simple to implement in the top-down approach [32], [33], [34], but in the bottom-up approach, it is based on approximation that the backward reflections from impedance discontinuities are negligible [39]. In addition, the echo model also becomes complex when multiple branches are connected at a common joint. In the frequency-domain modeling, the network is regarded as a composition of many cascade distributed portions. Then the whole network behavior can be described based on the transmission matrices or the scattering matrices of the cascaded portions [36], [37], [40], [42]. The main advantage of the frequency-domain modeling is its ability to consider all the signals reflected from the impedance discontinuities regardless of the complexity of the network.

Currently, there are a few frequency-domain models that are based on the complete bottom-up approach [40], [42]. The model parameters can be derived from the eigen analysis of multiconductor network matrices [42] or from the lump-circuit transmission line model [40]. The typical modeling algorithms used are either the transmission matrix [40] or the scattering matrix [42] as shown in Figure 2.7.

The transmission matrix gives the relationship of the voltage V and the current I at the two terminals while the scattering matrix gives the relationship of the incident a and reflected b waves.

(a) Transmission matrix

(b) Scattering matrix

Figure 2.7. Comparison of (a) transmission matrix with (b) scattering matrix of a 2-port network.

Multi-conductor Transmission Line (MTL) theory is used here for the modeling of the DM and CM propagation paths of the power line cable. This approach allows us to determine a circuit model that well characterizes the underlying physics of signal propagation over power line cable. We will show how the proposed approach allows one to compute the DM and CM current of any power line link by using two-port transmission matrices, as commonly done for telephone channel modeling. The model parameters are obtained through a bottom-up approach based on the transmission matrices. The choice of the transmission matrix over the scattering matrix is due to the fact that we are interested in the level of DM and CM current propagating in the network.

2.6 MITIGATION TECHNIQUES

The capacity of a transmission channel (in bits per second) depends on the bandwidth available, the power of the signal used for the transmission of the data and the power of the noise in the channel. Once all of the available bandwidths are used, the transmission speed can only be increased either by reducing the noise or by increasing the transmitted signal power. Although it is possible to reduce some of the noise in a transmission

channel, one has, in general, little or no control over the noise floor and over some of the noise sources. Increasing the transmitted power is therefore the only alternative if higher bit rates are required. EMC considerations, however, place limits on the maximum power that can be transmitted. Indeed, in general, the higher the transmitted power, the higher the emission level. This EMC-based limitation in the maximum transmitted power could represent a major obstacle to the success of PLC technology.

A solution to this problem is to devise techniques to mitigate emissions. The expression “mitigation techniques” in the PLC domain refers to any method designed to lower emissions without significantly affecting the attainable transmission throughput. Examples of mitigation techniques include shielding of the low voltage cables in the home or shielding of the victim equipment to improve the immunity of the victim equipment. Others include the reduction of the injected common mode through the use of improved couplers, the notching of sensitive frequencies, common mode suppressors and active signal injection for the cancellation of field components through deliberate induction of destructive interference [24], [43]-[46]. In Chapter 5, we propose a new injection technique that makes use of the cancellation of field components through induction of destructive interference to mitigate the radiation emission levels.

Chapter 3

REVIEW OF THE THEORY OF MULTI-CONDUCTOR TRANSMISSION LINE

The theory used in the modeling of three conductors in a straight line is discussed in this chapter. The low voltage power line is analyzed as a Multi-conductor Transmission Line (MTL) and the corresponding transmission line equations and parameters are derived.

3.1 MTL APPROXIMATION OF POWER LINE

The power line under study in this research is the typical single-phase house wirings commonly found in Singapore. The three wires (live, neutral and earth) are usually laid together inside the wall as shown in Figure 3.1. These wires are made up of stranded copper conductors with PVC (Polyvinyl Chloride) insulation.

Figure 3.1. Cross sectional view of the house service power line cable.

This structure of the power line can be approximated as MTL when it is used to transmit the high frequency communication signals. The term MTL refers to a set of $(n+1)$ parallel conductors that serve to transmit electrical signals between sources and loads, where n is the number of conductors excluding the reference conductor. The dominant mode of propagation in a MTL is the Transverse Electromagnetic (TEM) mode where the electric and magnetic fields surrounding the conductors lie solely in the transverse plane orthogonal to the line axis. This structure is capable of guiding waves

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whose frequencies range from dc to where the line cross-sectional dimension becomes a significant fraction of the wavelength.

The assumption for the MTL approximation of the power line cable is that the wave propagation on the power line cable follows the TEM mode. This requires the line structure to have electrically small cross-sectional dimensions, meaning that the distance between the conductors is small in comparison with the operating wavelength.

The other assumption made is that the power line cable should be in a uniform structure, which means that, the conductors in the structure should have uniform cross sections along their length and are parallel to each other. In the actual installation, the power line cables are actually laid close together inside a plastic tray or a metal conduit which has a small cross-sectional area. This limits the variation in the separation between the cables and we approximate it to be a uniform structure.

Based on the above considerations, the transmission line equations and the chain parameter matrix are derived based on [47].

3.2 DERIVATION OF TRANSMISSION LINE EQUATIONS

The transmission line equations are very important for deriving the line parameters which are later used as the model parameters in network modeling. They are usually derived from a representation of the line as lumped circuit elements distributed along the line. The Maxwell equations will be used to derive the MTL equations.

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Figure 3.2. Definition of contour for derivation of the first MTL equation.

Figure 3.2 shows the general $(n+1)$ conductor lines to be considered. It consists of n conductors and a reference conductor (earth conductor), which is denoted as the 0^{th} conductor in the figure. Recall the Faraday's law in integral form:

$$\oint_c \vec{E} \cdot d\vec{l} = -\mu \frac{d}{dt} \int_s \vec{H} \cdot d\vec{s} \quad (3.1)$$

where $\vec{E}$: Electrical field intensity vector (V/m)

$\vec{H}$: Magnetic field intensity vector (A/m)

c : Contour of field area

s : Surface of field area

μ : Permeability of conductor (H/m)

l : Length along the z -axis

Applying this to the conductor c_i , which encloses surface s_i as shown between the reference conductor and the i -th conductor, and encircling it in the clockwise direction gives [47]

$$\int_a^{a'} \vec{E}_t \cdot d\vec{l} + \int_{a'}^{b'} \vec{E}_z \cdot d\vec{l} + \int_{b'}^b \vec{E}_t \cdot d\vec{l} + \int_b^a \vec{E}_z \cdot d\vec{l} = \mu \frac{d}{dt} \int_{s_i} \vec{H}_t \cdot d\vec{s} \quad (3.2)$$

where $\vec{E}_t$ denotes the transverse electric field in the x-y cross-sectional plane and $\vec{E}_z$ denotes the electric field in the z direction. Because of the choice of the direction of the contour, the direction $\vec{H}_t$, and the right-hand rule, the minus sign on the right-hand side of (3.1) is absent in (3.2).

Under the assumption of TEM structure, the voltage between the i -th conductor and the reference conductor can be defined as

$$V_i(z, t) = - \int_a^{a'} \vec{E}_t(x, y, z, t) \cdot d\vec{l} \quad (3.3a)$$

$$V_i(z + \Delta z, t) = - \int_b^{b'} \vec{E}_t(x, y, z + \Delta z, t) \cdot d\vec{l} \quad (3.3b)$$

The integrals along the z direction are zero if the conductors are considered to be perfect conductors. It is pointed out that the TEM mode cannot exist if the conductors are not perfect conductors. This is because a component of electric field will be directed in the z direction due to the voltage drops along the conductors. However, small losses can be accommodated as an approximation under the quasi-TEM mode assumption. To allow for imperfect conductors, the per-unit-length conductor resistance R (Ω/m) can be defined. Thus,

$$\int_{a'}^{b'} \vec{E}_z \cdot d\vec{l} = R_i \cdot \Delta z \cdot I_i(z, t) \quad (3.4a)$$

$$\int_b^a \vec{E}_z \cdot d\vec{l} = R_0 \cdot \Delta z \cdot \sum_{k=1}^n I_k(z, t) \quad (3.4b)$$

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Substituting (3.3) and (3.4) into (3.2) yields

$$-V_i(z, t) + R_i \cdot \Delta z \cdot I_i(z, t) + V_i(z + \Delta z, t) + R_0 \cdot \Delta z \cdot \sum_{k=1}^n I_k(z, t) = \mu \frac{d}{dt} \int_{s_i} \vec{H}_t \cdot \vec{ds} \quad (3.5)$$

Dividing both sides by Δz , and rearranging gives

$$\frac{V_i(z + \Delta z, t) - V_i(z, t)}{\Delta z} = -R_0 I_1 - R_0 I_2 - \dots - (R_0 + R_i) I_i - \dots - R_0 I_n + \mu \frac{1}{\Delta z} \frac{d}{dt} \int_{s_i} \vec{H}_t \cdot \vec{ds} \quad (3.6)$$

Clearly, the total magnetic fluxes penetrating the surface s_i in Figure 3.2 will be a linear combination of the fluxes due to the currents in the conductors. This total magnetic flux can be written as

$$\begin{aligned} \Psi_i &= -\mu \lim_{\Delta z \rightarrow 0} \frac{1}{\Delta z} \int_{s_i} \vec{H}_t \cdot \vec{ds} \\ &= L_{i1} I_1 + L_{i2} I_2 + \dots + L_{ii} I_i + \dots + L_{in} I_n \end{aligned} \quad (3.7)$$

L_{i1} , L_{i2} , ..., L_{ii} and L_{in} are the per-unit-length inductances in (H/m). Substituting (3.7) into (3.5) and taking the limit as $\Delta z \rightarrow 0$ give

$$\begin{aligned} \frac{\partial V_i(z, t)}{\partial z} &= -R_0 I_1 - R_0 I_2 - \dots - (R_0 + R_i) I_i - \dots - R_0 I_n \\ &\quad - L_{i1} \frac{\partial I_1(z, t)}{\partial t} - L_{i2} \frac{\partial I_2(z, t)}{\partial t} - \dots - L_{ii} \frac{\partial I_i(z, t)}{\partial t} - \dots - L_{in} \frac{\partial I_n(z, t)}{\partial t} \end{aligned} \quad (3.8)$$

The first MTL equation can be written in a compact form using matrix notation as

$$\frac{\partial}{\partial z} V(z, t) = -R \cdot I(z, t) - L \frac{\partial}{\partial t} I(z, t) \quad (3.9)$$

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where the voltage and current vectors are defined as

$$V(z, t) = [V_1(z, t) \quad \dots \quad V_i(z, t) \quad \dots \quad V_n(z, t)]^T \quad (3.10a)$$

$$I(z, t) = [I_1(z, t) \quad \dots \quad I_i(z, t) \quad \dots \quad I_n(z, t)]^T \quad (3.10b)$$

The per-unit-length resistance matrix can be defined as

$$R = \begin{bmatrix} R_1 + R_0 & R_0 & \Lambda & R_0 \\ R_0 & R_2 + R_0 & \Lambda & R_0 \\ M & M & O & M \\ R_0 & R_0 & \Lambda & R_n + R_0 \end{bmatrix} \quad (3.11)$$

The per-unit-length inductance matrix contains the individual per-unit-length self-inductance L_{ii} of the conductor and the per-unit-length mutual inductance L_{ij} between the conductors.

$$L = \begin{bmatrix} L_{11} & L_{12} & \Lambda & L_{1n} \\ L_{21} & L_{22} & \Lambda & L_{2n} \\ M & M & O & M \\ L_{n1} & L_{n2} & \Lambda & L_{nn} \end{bmatrix} \quad (3.12)$$

The first MTL equation is a differential equation for the voltage on the line. The second equation will be for the current. It can also be derived according to the Maxwell's equations and by considering a closed surface $\oint$ around the i -th conductor as shown in Figure 3.3.

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Figure 3.3. Definition of the surface for derivation of the second MTL equation.

The portion of the surface end caps is denoted as $\hat{S}_e$, while the portion over the sides is denoted as $\hat{S}_o$. Recalling the continuity equation of conservation of charge [48],

$$\oiint_{\hat{S}} \vec{J} \cdot d\hat{S} = -\frac{\partial}{\partial t} Q \quad (3.13)$$

where $\vec{J}$ is the conduction current density vector in (A/m²), while Q is the charge enclosed by the closed surface $\hat{S}$ quantity in Coulombs.

Over the end caps (surface $\hat{S}_e$), it can be easily written that [47]

$$\iint_{\hat{S}_e} \vec{J} \cdot d\hat{S} = I_i(z + \Delta z, t) - I_i(z, t) \quad (3.14)$$

Over the sides of the surface (surface $\hat{S}_o$), there are two currents: conduction current, $\vec{J}_c = \sigma \vec{E}_t$, and the displacement current, $\vec{J}_d = \epsilon \frac{\partial \vec{E}_t}{\partial t}$, where the surrounding medium is characterized by conductivity σ (S/m) and permittivity ϵ (F/m). From Figure 3.3, the transverse conduction current flowing between the conductors is

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$$\iint_{\hat{s}_0} \vec{J}_c \cdot d\hat{s} = \sigma \iint_{\hat{s}_0} \vec{E}_t \cdot d\hat{s} \quad (3.15)$$

This can be considered by defining per-unit-length conductance G_{ij} (S/m) between each pair of conductors as the ratio of conduction current flowing between the two conductors in the transverse plane to the voltage difference between the two conductors. Therefore,

$$\begin{aligned} \sigma \lim_{\Delta z \rightarrow 0} \iint_{\hat{s}_0} \vec{E}_t \cdot d\hat{s} &= G_{i1}(V_i - V_1) + K + G_{ii}V_i + K + G_{in}(V_i - V_n) \\ &= -G_{i1}V_1(z, t) - G_{i2}V_2(z, t) - K + \sum_{k=1}^n G_{ik}V_k(z, t) - K - G_{in}V_n(z, t) \end{aligned} \quad (3.16)$$

Similarly, by Gauss's Law, the charge enclosed by the surface (residing on the conductor surface) is

$$Q = \varepsilon \iint_{\hat{s}_0} \vec{E}_t \cdot d\hat{s} \quad (3.17)$$

The charge per unit of line length can be defined in terms of the per-unit-length capacitances C_{ij} (F/m) between each pair of conductors as

$$\begin{aligned} \varepsilon \lim_{\Delta z \rightarrow 0} \frac{1}{\Delta z} \iint_{\hat{s}_0} \vec{E}_t \cdot d\hat{s} &= C_{i1}(V_i - V_1) + K + C_{ii}V_i + K + C_{in}(V_i - V_n) \\ &= -C_{i1}V_1(z, t) - C_{i2}V_2(z, t) - K + \sum_{k=1}^n C_{ik}V_k(z, t) - K - C_{in}V_n(z, t) \end{aligned} \quad (3.18)$$

Substituting (3.14), (3.15) and (3.17) into (3.13), and dividing both sides by Δz gives

$$\frac{I_i(z + \Delta z, t) - I_i(z, t)}{\Delta z} + \sigma \frac{I}{\Delta z} \iint_{\hat{s}_0} \vec{E}_t \cdot d\hat{s} = -\frac{\partial}{\partial t} \left[\varepsilon \frac{I}{\Delta z} \iint_{\hat{s}_0} \vec{E}_t \cdot d\hat{s} \right] \quad (3.19)$$

Substituting (3.16) and (3.18) into (3.19), and taking the limit as $\Delta z \rightarrow 0$ yields

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$$\begin{aligned} \frac{\partial I_i(z, t)}{\partial z} = & G_{i1} V_1(z, t) + \Lambda - \sum_{k=1}^n G_{ik} V_k(z, t) + K + G_{in} V_n(z, t) \\ & + \frac{\partial}{\partial t} \left\{ C_{i1} V_1(z, t) + K - \sum_{k=1}^n C_{ik} V_k(z, t) + K + C_{in} V_n(z, t) \right\} \end{aligned} \quad (3.20)$$

Equation (3.20) can be written into the compact form with matrix notation, which forms the second MTL equation.

$$\frac{\partial}{\partial z} \mathbf{I}(z, t) = -\mathbf{G} \cdot \mathbf{V}(z, t) - \mathbf{C} \frac{\partial}{\partial t} \mathbf{V}(z, t) \quad (3.21)$$

where $\mathbf{V}$ and $\mathbf{I}$ are defined by (3.10). The per-unit-length conductance matrix, which represents the conduction current flowing between the conductors in the transverse plane, is defined as

$$\mathbf{G} = \begin{bmatrix} \sum_{k=1}^n G_{1k} & -G_{12} & \Lambda & -G_{1n} \\ -G_{21} & \sum_{k=1}^n G_{2k} & \Lambda & -G_{2n} \\ \mathbf{M} & \mathbf{M} & \mathbf{O} & \mathbf{M} \\ -G_{n1} & -G_{n2} & \Lambda & \sum_{k=1}^n G_{nk} \end{bmatrix} \quad (3.22)$$

The per-unit-length capacitance matrix, which represents the displacement current flowing between the conductors in the transverse plane, is defined as

$$\mathbf{C} = \begin{bmatrix} \sum_{k=1}^n C_{1k} & -C_{12} & \Lambda & -C_{1n} \\ -C_{21} & \sum_{k=1}^n C_{2k} & \Lambda & -C_{2n} \\ \mathbf{M} & \mathbf{M} & \mathbf{O} & \mathbf{M} \\ -C_{n1} & -C_{n2} & \Lambda & \sum_{k=1}^n C_{nk} \end{bmatrix} \quad (3.23)$$

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In summary, the distributed equivalent circuit for the MTL with $(n+1)$ conductors can be illustrated in Figure 3.4.

Figure 3.4. Distributed equivalent circuit for MTL.

3.3 DERIVATION OF LINE PARAMETERS

Equations (3.9) and (3.21) are the two MTL equations which describe the complete relationship between the line voltage $V(z,t)$ and current $I(z,t)$ and the per-unit-length line parameters. In this section the characteristic impedance, which describes the behavior of the wave propagation on the line, is further derived.

The line parameters can be derived through the frequency domain analysis of the MTL. In this analysis, it is assumed that the excitation sources to the MTLs are sinusoids which have been applied for a sufficient length of time so that the line voltage and current are in steady state. Therefore, the line voltage and current are also sinusoidal having a magnitude and phase angle. They can be written into the phasor form as

$$V_i(z,t) = \text{Re} \left\{ \hat{V}_i(z) e^{j\omega t} \right\} \quad (3.24)$$

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$$I_i(z, t) = \text{Re} \left\{ \hat{I}_i(z) e^{j\omega t} \right\} \quad (3.25)$$

where $\text{Re} \{ \}$ denotes the real part of the enclosed complex quantity and the phasor voltage and current have a magnitude and phase angle as

$$\hat{V}_i(z, t) = V_i(z) \angle \theta_i(z) = V_i(z) e^{j\theta_i(z)} \quad (3.26)$$

$$\hat{I}_i(z, t) = I_i(z) \angle \phi_i(z) = I_i(z) e^{j\phi_i(z)} \quad (3.27)$$

Substituting the phasor form of line voltage and current into MTL equations (3.9) and (3.21) gives

$$\frac{\partial}{\partial z} \hat{V}(z) = -Z \cdot \hat{I}(z) \quad (3.28a)$$

$$\frac{\partial}{\partial z} \hat{I}(z) = -Y \cdot \hat{V}(z) \quad (3.28b)$$

where the per-unit-length impedance matrix $\mathbf{Z}$, and the admittance matrix $\mathbf{Y}$, are given by

$$Z = R + j\omega L \quad (3.29a)$$

$$Y = G + j\omega C \quad (3.29b)$$

Combining (3.28a) and (3.28b) into a matrix form produces

$$\frac{\partial}{\partial z} \hat{X}(z) = A \cdot \hat{X}(z) \quad (3.30)$$

where

$$\hat{X}(z) = \begin{bmatrix} \hat{V}(z) \\ \hat{I}(z) \end{bmatrix}$$

$$A = \begin{bmatrix} 0 & -Z \\ -Y & 0 \end{bmatrix}$$

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The solution of (3.30) will give a full picture of the voltage and current propagation on the line. However, it is not easily solved since it is coupled. To overcome this problem, similarity transformation is used [47]. Firstly, the mode voltage $\hat{V}_m(z)$ and current $\hat{I}_m(z)$ can be defined as

$$\hat{V}(z) = T_v \cdot \hat{V}_m(z) \quad (3.31a)$$

$$\hat{I}(z) = T_l \cdot \hat{I}_m(z) \quad (3.31b)$$

The $n \times n$ complex matrix $\mathbf{T}_v$ and $\mathbf{T}_l$ are the similarity transformations between the actual line voltage and current and the mode voltage and current. In order for this to be valid, $\mathbf{T}_v$ and $\mathbf{T}_l$ must be non-singular, i.e. $\mathbf{T}_v^{-1}$ and $\mathbf{T}_l^{-1}$ must exist. Substituting the mode voltage and current into (3.30)

$$\frac{d}{dz} \begin{bmatrix} \hat{V}_m(z) \\ \hat{I}_m(z) \end{bmatrix} = \begin{bmatrix} 0 & -T_v^{-1} Z T_l \\ -T_l^{-1} Y T_v & 0 \end{bmatrix} \begin{bmatrix} \hat{V}_m(z) \\ \hat{I}_m(z) \end{bmatrix} \quad (3.32)$$

If $\mathbf{T}_v$ and $\mathbf{T}_l$ can be obtained such that $T_v^{-1} Z T_l$ and $T_l^{-1} Y T_v$ are diagonal as follows:

$$T_v^{-1} Z T_l = Z' = \begin{bmatrix} Z'_1 & 0 & \Lambda & 0 \\ 0 & Z'_2 & 0 & M \\ M & 0 & 0 & 0 \\ 0 & \Lambda & 0 & Z'_n \end{bmatrix} \quad (3.33a)$$

$$T_l^{-1} Y T_v = Y' = \begin{bmatrix} Y'_1 & 0 & \Lambda & 0 \\ 0 & Y'_2 & 0 & M \\ M & 0 & 0 & 0 \\ 0 & \Lambda & 0 & Y'_n \end{bmatrix} \quad (3.33b)$$

then the MTL equations can be decoupled as

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$$\begin{cases} \frac{d}{dz} V_{m1}(z) = -Z'_{1 \cdot m1}(z) & \frac{d}{dz} I_{m1}(z) = -Y'_{1 \cdot m1}(z) \\ \quad \quad \quad \mathbf{M} & \quad \quad \quad \mathbf{M} \\ \frac{d}{dz} V_{mn}(z) = -Z'_{1 \cdot mn}(z) & \frac{d}{dz} I_{mn}(z) = -Y'_{1 \cdot mn}(z) \end{cases}$$

To find the similarity transformation matrices of $\mathbf{T}_v$ and $\mathbf{T}_l$ to make Z' and Y' diagonal, we combine (3.28a) and (3.28b) as follows:

$$\frac{d^2}{dz^2} \hat{V}(z) = ZY\hat{V}(z) \quad (3.34a)$$

$$\frac{d^2}{dz^2} \hat{I}(z) = YZ\hat{I}(z) \quad (3.34b)$$

Substituting in the similarity transformations as given in (3.31), we have

$$\frac{d^2}{dz^2} \hat{V}_m(z) = T_v^{-1} ZYT_v \hat{V}_m(z) = T_v^{-1} ZT_l T_l^{-1} YT_v \hat{V}_m(z) = Z' Y' \hat{V}_m(z) \quad (3.35a)$$

$$\frac{d^2}{dz^2} \hat{I}_m(z) = T_l^{-1} YZT_l \hat{I}_m(z) = T_l^{-1} YT_v T_v^{-1} ZT_l \hat{I}_m(z) = Y' Z' \hat{I}_m(z) \quad (3.35b)$$

Recalling that $\mathbf{Z}$ and $\mathbf{Y}$ are symmetric, i.e. $\mathbf{Z}^T = \mathbf{Z}$ and $\mathbf{Y}^T = \mathbf{Y}$, and since the transpose of the product of row matrices is the product of the transposes of the matrices in reverse order, it can be derived that

$$\begin{aligned} (T_v^{-1} ZYT_v)^T &= (T_v)^T (Y)^T (Z)^T (T_v^{-1})^T \\ &= (T_v)^T YZ(T_v^{-1})^T \\ &= Z' Y' \\ &= Y' Z' \end{aligned} \quad (3.36)$$

Since the required Z' and Y' are diagonal, comparing (3.36) with (3.35b), the following relationship can be obtained:

$$T_L^T = T_v^{-1} = T \quad (3.37)$$

Substituting (3.37) into (3.35b), we have

$$\frac{d^2}{dz^2} \hat{I}_m(z) = T^{-1} Y Z T \hat{I}_m(z) = \gamma^2 \hat{I}_m(z) \quad (3.38)$$

where γ^2 is a diagonal matrix as

$$\gamma^2 = T^{-1} Y Z T = \begin{bmatrix} \gamma_1^2 & 0 & \Lambda & 0 \\ 0 & \gamma_2^2 & 0 & M \\ M & 0 & 0 & 0 \\ 0 & \Lambda & 0 & \gamma_n^2 \end{bmatrix} \quad (3.39)$$

It should be noted that γ is the diagonal matrix that represents the propagation constants between different conductors and the reference conductors. The element inside the matrix γ , which is defined as γ_i , contains both real and imaginary parts ($\gamma_i = \alpha_i + j\beta_i$). α_i is the attenuation constant (in Np/m) and it is related to the impedance matrix Z . β_i is the phase constant (in rad/m) and it is related to the admittance matrix Y .

From (3.38), the general solution for the mode current is

$$\hat{I}_m(z) = e^{-\gamma z} \hat{I}_m^+ - e^{\gamma z} \hat{I}_m^- \quad (3.40)$$

where the plus and minus superscripts denote waves travelling in the $+z$ and $-z$ directions. The actual current can also be obtained by multiplying the mode current by the transformation matrix $\mathbf{T}=\mathbf{T}_1$ to give

$$\hat{I}(z) = T \hat{I}_m(z) = T(e^{-\gamma z} \hat{I}_m^+ - e^{\gamma z} \hat{I}_m^-) \quad (3.41)$$

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Similarly, by substituting (3.37) into (3.35a), the decoupled second-order differential equation for the mode voltage can be obtained as

$$\frac{d^2}{dz^2} \hat{V}_m(z) = T^{-1} Z Y T \hat{V}_m(z) = T^T Z Y (T^T)^{-1} \hat{V}_m(z) = \gamma^2 \hat{V}_m(z) \quad (3.42)$$

The general solution for the mode voltage is

$$\hat{V}_m(z) = e^{-\gamma z} \hat{V}_m^+ - e^{\gamma z} \hat{V}_m^- \quad (3.43)$$

The actual voltage can be obtained by multiplying the mode voltage by the transformation matrix, $\mathbf{T}_v = (\mathbf{T}_l^{-1}) = (\mathbf{T}^{-1})^T$ to give

$$\hat{V}(z) = (T^{-1})(e^{-\gamma z} \hat{V}_m^+ - e^{\gamma z} \hat{V}_m^-) \quad (3.44)$$

The solutions (3.41) and (3.44) are actually related. To determine this relation, (3.41) is substituted into the second MTL equation (3.28b),

$$\begin{aligned} \hat{V}(z) &= -Y^{-1} \frac{d}{dz} \hat{I}(z) \\ &= Y^{-1} T \gamma (e^{-\gamma z} \hat{I}_m^+ + e^{\gamma z} \hat{I}_m^-) \\ &= (Y^{-1} T \gamma T^{-1}) T (e^{-\gamma z} \hat{I}_m^+ + e^{\gamma z} \hat{I}_m^-) \\ &= Z_0 T (e^{-\gamma z} \hat{I}_m^+ + e^{\gamma z} \hat{I}_m^-) \end{aligned} \quad (3.45)$$

where the characteristic impedance matrix can be defined as [47]

$$\begin{aligned} Z_0 &= Y^{-1} T \gamma T^{-1} \\ &= Z T \gamma T^{-1} \end{aligned} \quad (3.46)$$

3.4 CHARACTERIZATION OF LINE AS A 2n PORT WITH CHAIN PARAMETER MATRIX

The phasor voltages and currents at the two ends of the line can be related with the chain parameter matrix as

$$\begin{bmatrix} \hat{V}(L) \\ \hat{I}(L) \end{bmatrix} = \hat{\Phi}(L) \begin{bmatrix} \hat{V}(0) \\ \hat{I}(0) \end{bmatrix} = \begin{bmatrix} \hat{\Phi}_{11}(L) & \hat{\Phi}_{12}(L) \\ \hat{\Phi}_{21}(L) & \hat{\Phi}_{22}(L) \end{bmatrix} \begin{bmatrix} \hat{V}(0) \\ \hat{I}(0) \end{bmatrix} \quad (3.47)$$

where $\hat{\Phi}(L)$ is the chain parameter matrix.

This corresponds to viewing the (n+1)-conductor line as a 2n port network as illustrated in Figure 3.5. The essential task in solving the phasor MTL equations is to determine the entries in the n x n submatrices $\hat{\Phi}_{ij}$. This section is devoted to this task.

The general solutions of the phasor MTL equations are given, via similarity transformations in (3.41) and (3.45) as follows:

$$\hat{V}(z) = Y^{-1} \hat{T} \hat{\gamma} (e^{-\hat{\gamma} z} \hat{I}_m^+ - e^{\hat{\gamma} z} \hat{I}_m^-) \quad (3.48a)$$

$$\hat{I}(z) = \hat{T} (e^{-\hat{\gamma} z} \hat{I}_m^+ - e^{\hat{\gamma} z} \hat{I}_m^-) \quad (3.48b)$$

Figure 3.5. Illustration of viewing an (n+1)-conductor line as a 2n port.

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Evaluating these at $z=0$ and $z=L$ and eliminating $\hat{I}_m^\pm$ give the parameter matrix submatrices as [47]:

$$\hat{\Phi}_{11}(L) = \frac{1}{2} \hat{Y}^{-1} \hat{T} (e^{\hat{\gamma}L} + e^{-\hat{\gamma}L}) \hat{T}^{-1} \hat{Y} \quad (3.49a)$$

$$\begin{aligned} \hat{\Phi}_{12}(L) &= -\frac{1}{2} \hat{Y}^{-1} \hat{T} \hat{\gamma} (e^{\hat{\gamma}L} - e^{-\hat{\gamma}L}) \hat{T}^{-1} \\ &= -\frac{1}{2} \hat{Y}^{-1} \hat{T} \hat{\gamma} \hat{T}^{-1} [\hat{T} (e^{\hat{\gamma}L} - e^{-\hat{\gamma}L}) \hat{T}^{-1}] \\ &= -\frac{1}{2} \hat{Z}_0 [\hat{T} (e^{\hat{\gamma}L} - e^{-\hat{\gamma}L}) \hat{T}^{-1}] \end{aligned} \quad (3.49b)$$

$$\begin{aligned} \hat{\Phi}_{21}(L) &= -\frac{1}{2} \hat{T} (e^{\hat{\gamma}L} - e^{-\hat{\gamma}L}) \hat{\gamma}^{-1} \hat{T}^{-1} \hat{Y} \\ &= -\frac{1}{2} [\hat{T} (e^{\hat{\gamma}L} - e^{-\hat{\gamma}L}) \hat{T}^{-1}] \hat{T} \hat{\gamma}^{-1} \hat{T}^{-1} \hat{Y} \\ &= -\frac{1}{2} [\hat{T} (e^{\hat{\gamma}L} - e^{-\hat{\gamma}L}) \hat{T}^{-1}] \hat{Y}_0 \end{aligned} \quad (3.49c)$$

$$\hat{\Phi}_{22}(L) = \frac{1}{2} \hat{T} (e^{\hat{\gamma}L} + e^{-\hat{\gamma}L}) \hat{T}^{-1} \text{ where } \hat{Y}_0 = \hat{Z}_0^{-1}. \quad (3.49d)$$

Recall the basic diagonalization in (3.39):

$$T^{-1} Y Z T = \gamma^2 \quad (3.50)$$

The square root of a matrix can be defined as a matrix which when multiplied by itself gives the original matrix, i.e. $\sqrt{M} \sqrt{M} = M$. From this we can define the square root of the matrix product as

$$\sqrt{YZ} = T \gamma T^{-1} \quad (3.51)$$

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The matrix exponential can be defined as

$$e^{\sqrt{YZ}L} = I_n + \frac{L}{1!}\sqrt{YZ} + \frac{L^2}{2!}(\sqrt{YZ})^2 + \dots \quad (3.52a)$$

$$e^{\gamma L} = I_n + \frac{L}{1!}\gamma + \frac{L^2}{2!}\gamma^2 + \dots \quad (3.52b)$$

We may define the matrix hyperbolic functions as

$$\begin{aligned} \cosh(\sqrt{YZ}L) &= \frac{1}{2}(e^{\sqrt{YZ}L} + e^{-\sqrt{YZ}L}) \\ &= I_n + \frac{L^2}{2!}(\sqrt{YZ})^2 + \frac{L^4}{4!}(\sqrt{YZ})^4 + \dots \\ &= \frac{1}{2}\hat{T}(e^{\hat{\gamma}L} + e^{-\hat{\gamma}L})\hat{T}^{-1} \end{aligned} \quad (3.53a)$$

$$\begin{aligned} \sinh(\sqrt{YZ}L) &= \frac{1}{2}(e^{\sqrt{YZ}L} - e^{-\sqrt{YZ}L}) \\ &= \frac{L}{1!}(\sqrt{YZ}) + \frac{L^3}{3!}(\sqrt{YZ})^3 + \dots \\ &= \frac{1}{2}\hat{T}(e^{\hat{\gamma}L} - e^{-\hat{\gamma}L})\hat{T}^{-1} \end{aligned} \quad (3.53b)$$

The chain parameter matrix can be re-written as

$$\hat{\Phi}_{11}(L) = \cosh(\sqrt{ZY}L) = \hat{Y}^{-1} \cosh(\sqrt{YZ}L)\hat{Y} \quad (3.54a)$$

$$\hat{\Phi}_{12}(L) = -Z_0 \sinh(\sqrt{YZ}L) = -\sinh(\sqrt{ZY}L)Z_0 \quad (3.54b)$$

$$\hat{\Phi}_{21}(L) = -Z_0^{-1} \sinh(\sqrt{ZY}L) = -\sinh(\sqrt{YZ}L)Z_0^{-1} \quad (3.54c)$$

$$\hat{\Phi}_{22}(L) = \cosh(\sqrt{YZ}L) = \hat{Y} \cosh(\sqrt{ZY}L)\hat{Y}^{-1} \quad (3.54d)$$

3.5 PRACTICAL APPLICATION OF THE CHAIN PARAMETER MATRIX AND THE RELATIONSHIP TO MODAL VOLTAGES

When the chain parameter matrix is determined, the voltage and current ($\hat{V}(L)$ and $\hat{I}(L)$) at the load end can be determined when the voltage and current ($\hat{V}(0)$ and $\hat{I}(0)$) at the source end are known and vice-versa.

Figure 3.6. Application of chain parameter matrix.

Let us assume that $Z_s = 0$ and let $\hat{V}_s$ and $\hat{V}_L$ denote the voltages at the source and load ends respectively. These give

$$\hat{V}_s = \hat{V}(0) \quad (3.55)$$

$$\hat{V}_L = Z_L \hat{I}(L) \quad (3.56)$$

From (3.47), (3.55) and (3.56), we get

$$\hat{V}_L = \hat{\Phi}_{11}(L) \hat{V}_s + \hat{\Phi}_{12}(L) \hat{I}(0) \quad (3.57)$$

$$\hat{I}(L) = \hat{\Phi}_{21}(L) \hat{V}_s + \hat{\Phi}_{22}(L) \hat{I}(0) \quad (3.58)$$

Using (3.56) and multiplying Z_L by (3.58), we get

$$\hat{V}_L = Z_L \hat{\Phi}_{21}(L) \hat{V}_S + Z_L \hat{\Phi}_{22}(L) \hat{I}(0) \quad (3.59)$$

Equating (3.57) and (3.59) and solving them, we get

$$\hat{I}(0) = [\hat{\Phi}_{12}(L) - Z_L \hat{\Phi}_{22}(L)]^{-1} [Z_L \hat{\Phi}_{21}(L) - \hat{\Phi}_{11}(L)] \cdot \hat{V}_S \quad (3.60)$$

Substituting (3.60) into (3.57), we get

$$\begin{aligned} \hat{V}_L &= \hat{\Phi}_{11}(L) \hat{V}_S + \hat{\Phi}_{12}(L) [\hat{\Phi}_{12}(L) - Z_L \hat{\Phi}_{22}(L)]^{-1} [Z_L \hat{\Phi}_{21}(L) - \hat{\Phi}_{11}(L)] \cdot \hat{V}_S \\ &= \left(\hat{\Phi}_{11}(L) + \hat{\Phi}_{12}(L) [\hat{\Phi}_{12}(L) - Z_L \hat{\Phi}_{22}(L)]^{-1} [Z_L \hat{\Phi}_{21}(L) - \hat{\Phi}_{11}(L)] \right) \cdot \hat{V}_S \end{aligned} \quad (3.61)$$

Equation (3.61) shows that the power line cable effects on voltage $\hat{V}_L$ are to be ascribed to submatrices $\hat{\Phi}_{11}(L)$, $\hat{\Phi}_{12}(L)$, $\hat{\Phi}_{21}(L)$ and $\hat{\Phi}_{22}(L)$ of the chain parameter matrix $\hat{\Phi}(L)$. In order to describe qualitatively such effects, the power line cable cross-section properties and length should be specified and (3.61) should be solved explicitly. This will be done in the following chapter.

Figure 3.7. Circuit model of the EMI source.
(a) Total voltage sources. (b) Modal voltage sources.

Figure 3.7 shows the circuit model of the electromagnetic interference (EMI) sources. The EMI voltage sources are made up of the modal voltage sources namely differential-mode (DM) and common-mode (CM). The total voltages seen at nodes 1 and 2 are $\hat{V}_{S1}$ and $\hat{V}_{S2}$ respectively. The modal voltages, DM and CM voltages, are $\hat{V}_{Sdm}$ and $\hat{V}_{Scm}$ respectively.

The relationships between the total and modal voltages at the EMI source and load ends are expressed as

$$\hat{V}_S = \begin{bmatrix} \hat{V}_{S1} \\ \hat{V}_{S2} \end{bmatrix} = \begin{bmatrix} \hat{V}_{Scm} - 0.5\hat{V}_{Sdm} \\ \hat{V}_{Scm} + 0.5\hat{V}_{Sdm} \end{bmatrix} \quad (3.62)$$

$$\hat{V}_{Sm} = \begin{bmatrix} \hat{V}_{Scm} \\ \hat{V}_{Sdm} \end{bmatrix} = \begin{bmatrix} 0.5(\hat{V}_{S2} + \hat{V}_{S1}) \\ \hat{V}_{S2} - \hat{V}_{S1} \end{bmatrix} \quad (3.63)$$

$$\hat{V}_L = \begin{bmatrix} \hat{V}_{L1} \\ \hat{V}_{L2} \end{bmatrix} = \begin{bmatrix} \hat{V}_{Lcm} - 0.5\hat{V}_{Ldm} \\ \hat{V}_{Lcm} + 0.5\hat{V}_{Ldm} \end{bmatrix} \quad (3.64)$$

$$\hat{V}_{Lm} = \begin{bmatrix} \hat{V}_{Lcm} \\ \hat{V}_{Ldm} \end{bmatrix} = \begin{bmatrix} 0.5(\hat{V}_{L2} + \hat{V}_{L1}) \\ \hat{V}_{L2} - \hat{V}_{L1} \end{bmatrix} \quad (3.65)$$

where subscript m is used to indicate modal voltage and subscripts S and L are used to indicate source and load ends respectively.

Matrix transfer functions relating the EMI source voltages $\hat{V}_S, \hat{V}_{Sm}$ to the voltages at the load end $\hat{V}_L, \hat{V}_{Lm}$ are defined by the following:

$$\hat{V}_L = H\hat{V}_S \quad (3.66)$$

$$\hat{V}_{Lm} = H_m\hat{V}_{Sm} \quad (3.67)$$

$$\hat{V}_L = H_{Sm}\hat{V}_{Sm} \quad (3.68)$$

where

$$H = \left(\hat{\Phi}_{11}(L) + \hat{\Phi}_{12}(L) \left[\hat{\Phi}_{12}(L) - Z_L \hat{\Phi}_{22}(L) \right]^{-1} \left[Z_L \hat{\Phi}_{21}(L) - \hat{\Phi}_{11}(L) \right] \right) \quad (3.69)$$

$$H_m = T^{-1} H T \quad (3.70)$$

$$H_{sm} = H T = T H_m \quad (3.71)$$

Equation (3.69) is obtained from equation (3.61). Referring back to equation (3.41), the same analogy applies where the actual voltages and the modal voltages are related by the transformation matrix $\mathbf{T}$, which allows us to arrive at (3.70) and (3.71).

In general, H is a full matrix that does not show any symmetrical structure, i.e., non-zero elements with different diagonal and off-diagonal entries. As a consequence, the same is true for matrix H_m . This means that in general, mode conversion takes place, and structural asymmetry influences such a phenomenon. Here, structural asymmetry refers to the asymmetry in the line parameters of the power line cables. Besides this, imbalance of the source impedance and load impedance can also contribute to the mode conversion. Impedance mismatch which occurs at line junctions can also contribute to mode conversion.

In the next chapter, we will be using the derived line parameters and using this chain parameter matrix to model the DM and CM currents in the PLC network.

Chapter 4

THE COMPLETE CURRENT PROPAGATION MODEL

In this chapter, a complete current propagation model for a three-wire power line cable is derived based on the Multi-conductor Transmission Line (MTL) theory. This complete model consists of a differential-mode (DM) current propagation model and a common-mode (CM) current propagation model. The frequency of interest is from 1 to 30 MHz. The complete current propagation model can be used to characterize and understand the behavior of DM and CM in the PLC network.

4.1 THE DM AND CM CURRENT PROPAGATION MODEL

Figure 4.1. A simple PLC system.

Figure 4.1 shows a simple in-house PLC system consisting of a power line cable and two PLC modems. The signals transmitted from the PLC Modem 1 to PLC Modem 2 are in the form of DM signals. They propagate on the Live and Neutral wires of the power line cable. As a result of the imbalance of the PLC Modem 1, the pair of wires and PLC Modem 2, a portion of the injected DM signals is converted into unwanted CM signals [49]. The paths of the DM and CM currents in the PLC system are shown in Figure 4.2.

Figure 4.2. The complete current propagation paths in a PLC system.

Figure 4.2 shows the complete current propagation paths in a PLC system with three wires when a PLC signal is injected by the PLC Modem 1 into the Live and Neutral wires. The wanted PLC signal is the DM current I_{dm} and the CM current I_{cm} is the unwanted by-product. Z_{S1} and Z_{S2} are the DM source impedances and Z_{L1} and Z_{L2} are the DM load impedances. The CM source and load impedances are Z_{S3} and Z_{L3} respectively. The currents flowing in the Live, Neutral and Earth wires are given by

$$I_L = I_{dm} + I_{cm} \quad (4.1)$$

$$I_N = -I_{dm} + I_{cm} \quad (4.2)$$

$$I_E = -2I_{cm} \quad (4.3)$$

For the DM current I_{dm} , it flows into the Live wire from PLC Modem 1 to PLC Modem 2 and returns via the Neutral wire. PLC Modem 1 is the DM signal source and PLC Modem 2 is the load. The Neutral wire is taken as the reference conductor for the DM current propagation model as shown in Figure 4.3. The DM signal source consists of the DM source voltage V_{dm} and DM source impedance Z_{Sdm} . PLC modem 2 is represented by its own internal DM impedance Z_{Ldm} which acts as the DM terminating load. R_{dm} is the effective DM resistance per unit length (in Ω/m); G_{dm} is the effective DM conductance

per unit length (in S/m); L_{dm} is the effective DM inductance per unit length (in H/m); C_{dm} is the effective DM capacitance per unit length (in F/m). All the above parameters can be derived from the properties of the cable.

Figure 4.3. The DM current propagation model of a power line cable.

Figure 4.4. The CM current propagation model of a power line cable.

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The CM currents I_{cm} flow from PLC Modem 1 to PLC Modem 2 in the Live and Neutral wires and return to PLC Modem 1 via the Earth wire. PLC Modem 1 is taken as the CM noise source and PLC Modem 2 is taken as the load. The Earth wire is taken as the reference conductor for the CM current propagation model as shown in Figure 4.4. The CM noise source consists of the CM source voltage V_{cm} and CM source impedances Z_{Scm-L} , Z_{Scm-N} and Z_{Scm-E} . PLC modem 2 is represented by its own internal CM impedances Z_{Lcm-L} , Z_{Lcm-N} and Z_{Lcm-E} , which act as the CM terminating load. R_0 , R_l and R_2 are the effective CM resistances per unit length (in Ω/m); G_{11} , G_{12} and G_{22} are the effective CM conductances per unit length (in S/m); L_{11} , L_{12} and L_{22} are the effective CM inductances per unit length (in H/m); C_{11} , C_{12} and C_{22} are the effective CM capacitances per unit length (in F/m). All the above parameters can be derived from the properties of the cable.

The effective per-unit-length resistance matrices, conductance matrices, inductance matrices and capacitance matrices for the DM and CM current propagation models are as follows:

$$R_{DM} = [R_{dm}] \quad (4.4)$$

$$G_{DM} = [G_{dm}] \quad (4.5)$$

$$L_{DM} = [L_{dm}] \quad (4.6)$$

$$C_{DM} = [C_{dm}] \quad (4.7)$$

$$R_{CM} = \begin{bmatrix} R_0 + R_l & R_0 \\ R_0 & R_0 + R_2 \end{bmatrix} \quad (4.8)$$

$$G_{CM} = \begin{bmatrix} G_{11} + G_{12} & -G_{12} \\ -G_{12} & G_{12} + G_{22} \end{bmatrix} \quad (4.9)$$

$$L_{CM} = \begin{bmatrix} L_{11} & L_{12} \\ L_{12} & L_{22} \end{bmatrix} \quad (4.10)$$

$$C_{CM} = \begin{bmatrix} C_{11} + C_{12} & -C_{12} \\ -C_{12} & C_{12} + C_{22} \end{bmatrix} \quad (4.11)$$

These matrices allow for the computation of the Z and Y matrices which are the per-unit-length impedance and admittance matrices, respectively

$$Z = R + j\omega L \quad (4.12)$$

$$Y = G + j\omega C \quad (4.13)$$

where $R=R_{DM}$, $L=L_{DM}$, $G=G_{DM}$ and $C=C_{DM}$ for the DM current propagation model and $R=R_{CM}$, $L=L_{CM}$, $G=G_{CM}$ and $C=C_{CM}$ for the CM current propagation model.

Figure 4.5. Reduction of the three-wire transmission line model to a two-wire transmission line model.

Figure 4.5 shows the reduction of the partial three-wire transmission line model to a two-wire transmission line model. Using the idea of Figure 4.5 and by comparing Figures 4.3 and 4.4, we can see the relationship between the parameters in Figures 4.3 and 4.4 as below:

$$R_{dm} = R_1 + R_2 \quad (4.14)$$

$$L_{dm} = L_{11} + L_{22} - 2L_{12} \quad (4.15)$$

$$G_{dm} = G_{12} \quad (4.16)$$

$$C_{dm} = C_{12} \quad (4.17)$$

$$Z_{Sdm} = Z_{Scm-L} + Z_{Scm-N} \quad (4.18)$$

$$Z_{Ldm} = Z_{Lcm-L} + Z_{Lcm-N} \quad (4.19)$$

The chain parameter matrices can be computed as (from equation (3.54))

$$\Phi_{11}(L) = \cosh(\sqrt{ZY}L) = Y^{-1} \cosh(\sqrt{YZ}L)Y \quad (4.20)$$

$$\Phi_{12}(L) = -Z_0 \sinh(\sqrt{YZ}L) = -\sinh(\sqrt{ZY}L)Z_0 \quad (4.21)$$

$$\Phi_{21}(L) = -Z_0^{-1} \sinh(\sqrt{ZY}L) = -\sinh(\sqrt{YZ}L)Z_0^{-1} \quad (4.22)$$

$$\Phi_{22}(L) = \cosh(\sqrt{YZ}L) = Y \cosh(\sqrt{ZY}L)Y^{-1} \quad (4.23)$$

The DM and CM current propagation models of the power line cable in Figures 4.3 and 4.4 can be represented by their chain parameter matrices as shown in Figures 4.6 and 4.7.

Figure 4.6. DM current propagation model using the chain parameter matrix.

Figure 4.7. CM current propagation model using the chain parameter matrix.

The relationship between the injected voltages and the currents at the source and at any point on the transmission line is given by the chain parameter matrix as follows (from (3.47)):

$$\begin{bmatrix} V(L) \\ I(L) \end{bmatrix} = \Phi(L) \begin{bmatrix} V(0) \\ I(0) \end{bmatrix} \quad (4.24)$$

where L is the length of the power line cable.

The generalized Thevenin equivalent representations at the terminals relating the voltages and the currents are as follows:

$$V(0) = V_s - Z_s I(0) \quad (4.25)$$

$$V(L) = V_L + Z_L I(L) \quad (4.26)$$

For DM current propagation model:

$$V(0) = [V_{dm1}(0)] \quad (4.27)$$

$$V_s = [V_{dm}] \quad (4.28)$$

$$Z_s = [Z_{Sdm}] \quad (4.29)$$

$$I(0) = [I_{dm}(0)] \quad (4.30)$$

$$V(L) = [V_{dm1}(L)] \quad (4.31)$$

$$V_L = [I_{dm}(L) \cdot Z_{Ldm}] \quad (4.32)$$

$$Z_L = [Z_{Ldm}] \quad (4.33)$$

$$I(L) = [I_{dm}(L)] \quad (4.34)$$

Figures 4.8 and 4.9 show the modeling of the source and load terminations of the power line cable for the CM current propagation model respectively.

Figure 4.8. Modeling of the source termination for the CM current propagation model.

Figure 4.9. Modeling of the load termination for the CM current propagation model.

For CM current propagation model:

$$V(0) = \begin{bmatrix} V_{cm1}(0) \\ V_{cm2}(0) \end{bmatrix} \quad (4.35)$$

$$V_S = \begin{bmatrix} V_{cm1} \\ V_{cm2} \end{bmatrix} = \begin{bmatrix} V_{cm} \\ V_{cm} \end{bmatrix} \quad (4.36)$$

$$Z_S = \begin{bmatrix} 2Z_{Scm-E} + Z_{Scm-L} & 0 \\ 0 & 2Z_{Scm-E} + Z_{Scm-N} \end{bmatrix} \quad (4.37)$$

$$I(0) = \begin{bmatrix} I_{cm1}(0) \\ I_{cm2}(0) \end{bmatrix} = \begin{bmatrix} I_{cm}(0) \\ I_{cm}(0) \end{bmatrix} \quad (4.38)$$

$$V(L) = \begin{bmatrix} V_{cm1}(L) \\ V_{cm2}(L) \end{bmatrix} \quad (4.39)$$

$$V_L = \begin{bmatrix} V_{L1} \\ V_{L2} \end{bmatrix} = \begin{bmatrix} I_{cm1}(L) \cdot 2Z_{Lcm-E} \\ I_{cm2}(L) \cdot 2Z_{Lcm-E} \end{bmatrix} \quad (4.40)$$

$$Z_L = \begin{bmatrix} Z_{Lcm-L} & 0 \\ 0 & Z_{Lcm-N} \end{bmatrix} \quad (4.41)$$

$$I(L) = \begin{bmatrix} I_{cm1}(L) \\ I_{cm2}(L) \end{bmatrix} = \begin{bmatrix} I_{cm}(L) \\ I_{cm}(L) \end{bmatrix} \quad (4.42)$$

4.2 DETERMINATION OF LINE PARAMETERS

The per-unit-length matrices of the power line cable need to be derived first in order to obtain the chain parameter matrix.

4.2.1 Resistance

When a high frequency current flows in the power line cable, we need to consider the skin effect when computing the per-unit-length resistance. Here, we assume that the wires are widely separated such that proximity effect is not prominent. This wide-separation approximation is valid as the Live, Neutral and Earth wires are usually not tied together. We assume that the current flows only within the skin depth of the wire. The skin depth δ is given by [47]

$$\delta = \frac{1}{\sqrt{\pi f \mu \sigma}} \quad (4.43)$$

where μ is the permeability of the metal wire and σ is its conductivity. The high frequency per-unit-length resistance for a wire of radius r_w in a homogeneous medium can be approximated by

$$R_w = \frac{1}{2\pi r_w \sigma \delta} \quad (4.44)$$

The above equation is valid only for a complete circular wire with $r_{cable} > 2\delta$. At the frequency of 1 MHz, copper with $\mu = 1.257 \times 10^{-6} \text{ H}\cdot\text{A}^{-2}$ and $\sigma = 5.813 \times 10^7 \text{ S}\cdot\text{m}^{-1}$ has a skin depth δ of $2.087 \times 10^{-4} \text{ m}$. For frequencies above 1 MHz, any cable with radius $r_{cable} > 4.174 \times 10^{-4} \text{ m}$ will fulfil this condition of $r_{cable} > 2\delta$. In the case for stranded conductor cables, corrections need to be accounted for the gaps at the circumference of the conductor strands. This can be shown in Figure 4.10 in which δ is the skin depth, and the shaded area indicates the area of current flow. To compute the resistance, the ratio of the

effective current flow area of the whole cable and the stranded conductor needs to be determined.

Figure 4.10. The area of current flow on the skin depth of the wire.

The current flow area for the whole wire is given by

$$\begin{aligned} A_w &= \pi r_w^2 - \pi (r_w - \delta)^2 \\ &= \pi (2r_w \delta - \delta^2) \end{aligned} \quad (4.45)$$

In order to calculate the total shaded area, we need to consider the area of current flow on one stranded conductor. The area of the shaded area A_{shaded} in Figure 4.11 can be obtained by subtracting the area of triangle OIG, A_{OIG} and the area of the crescent shape IHGJ, A_{IHGJ} from the area of the circular sector OI DG, A_{OIDG} .

Figure 4.11. The area of current flow on a single stranded conductor.

Figure 4.11 shows the area of current flow on one stranded wire. Using the Law of Cosines, we can get

$$OI^2 = AI^2 + AO^2 - 2AI \cdot AO \cdot \cos(\angle IAO) \quad (4.46)$$

From this,

$$\begin{aligned} \cos(\angle IAO) &= \frac{AI^2 + AO^2 - OI^2}{2 \cdot AI \cdot AO} \\ &= \frac{(r_w - \delta)^2 + (r_w - r_{cond})^2 - r_{cond}^2}{2 \cdot (r_w - \delta) \cdot (r_w - r_{cond})} \\ &= \frac{2r_w^2 - 2r_w \cdot \delta + \delta^2 - 2r_w \cdot r_{cond}}{2 \cdot (r_w - \delta) \cdot (r_w - r_{cond})} \end{aligned} \quad (4.47)$$

$$\angle IAO = \cos^{-1} \left[\frac{2r_w^2 - 2r_w \cdot \delta - 2r_w \cdot r_{cond} + \delta^2}{2 \cdot (r_w - \delta) \cdot (r_w - r_{cond})} \right]$$

$$\angle IAG = 2 \cdot \angle IAO \quad (4.48)$$

Using (4.47) and (4.48), the area of triangle AIG ,

$$\begin{aligned} A_{AIG} &= \frac{1}{2} AI \cdot AG \cdot \sin(\angle IAG) \\ &= \frac{1}{2} (r_w - \delta)^2 \cdot \sin(\angle IAG) \\ &= \frac{1}{2} (r_w - \delta)^2 \cdot \sin \left\{ 2 \cos^{-1} \left[\frac{2r_w^2 - 2r_w \cdot \delta - 2r_w \cdot r_{cond} + \delta^2}{2 \cdot (r_w - \delta) \cdot (r_w - r_{cond})} \right] \right\} \end{aligned} \quad (4.49)$$

To calculate the area of sector AIHG,

$$\begin{aligned}
 A_{AIHG} &= \frac{1}{2} AI^2 \cdot \angle IAG \\
 &= \frac{1}{2} (r_w - \delta)^2 \cdot \left\{ 2 \cos^{-1} \left[\frac{2r_w^2 - 2r_w \cdot \delta - 2r_w \cdot r_{cond} + \delta^2}{2 \cdot (r_w - \delta) \cdot (r_w - r_{cond})} \right] \right\} \\
 &= (r_w - \delta)^2 \cdot \cos^{-1} \left[\frac{2r_w^2 - 2r_w \cdot \delta - 2r_w \cdot r_{cond} + \delta^2}{2 \cdot (r_w - \delta) \cdot (r_w - r_{cond})} \right]
 \end{aligned} \tag{4.50}$$

Using (4.49) and (4.50), the area of crescent shape IHGJ, is obtained by

$$\begin{aligned}
 A_{IHGJ} &= A_{AIHG} - A_{AIG} \\
 &= (r_w - \delta)^2 \cdot \cos^{-1} \left[\frac{2r_w^2 - 2r_w \cdot \delta - 2r_w \cdot r_{cond} + \delta^2}{2 \cdot (r_w - \delta) \cdot (r_w - r_{cond})} \right] \\
 &\quad - \frac{1}{2} (r_w - \delta)^2 \cdot \sin \left\{ 2 \cos^{-1} \left[\frac{2r_w^2 - 2r_w \cdot \delta - 2r_w \cdot r_{cond} + \delta^2}{2 \cdot (r_w - \delta) \cdot (r_w - r_{cond})} \right] \right\}
 \end{aligned} \tag{4.51}$$

Using trigonometry,

$$\begin{aligned}
 IJ &= AI \cdot \sin(\angle IAJ) \\
 &= (r_w - \delta) \cdot \sin \left\{ \cos^{-1} \left[\frac{2r_w^2 - 2r_w \cdot \delta - 2r_w \cdot r_{cond} + \delta^2}{2 \cdot (r_w - \delta) \cdot (r_w - r_{cond})} \right] \right\}
 \end{aligned} \tag{4.52}$$

$$\begin{aligned}
 \angle IOJ &= \sin^{-1} \left(\frac{IJ}{IO} \right) \\
 &= \sin^{-1} \left\{ \frac{(r_w - \delta) \cdot \sin \left\{ \cos^{-1} \left[\frac{2r_w^2 - 2r_w \cdot \delta - 2r_w \cdot r_{cond} + \delta^2}{2 \cdot (r_w - \delta) \cdot (r_w - r_{cond})} \right] \right\}}{r_{cond}} \right\}
 \end{aligned} \tag{4.53}$$

$$\angle IOG = 2 \cdot \angle IOJ \tag{4.54}$$

To calculate the area of the sector OIDG,

$$\begin{aligned}
 A_{OIDG} &= \frac{1}{2} OI^2 \cdot \angle IOG \\
 &= \frac{1}{2} r_{cond}^2 \cdot \left\{ 2 \cdot \sin^{-1} \left\{ \frac{(r_w - \delta) \cdot \sin \left\{ \cos^{-1} \left[\frac{2r_w^2 - 2r_w \cdot \delta - 2r_w \cdot r_{cond} + \delta^2}{2 \cdot (r_w - \delta) \cdot (r_w - r_{cond})} \right] \right\}}{r_{cond}} \right\} \right\} \\
 &= r_{cond}^2 \cdot \sin^{-1} \left\{ \frac{(r_w - \delta) \cdot \sin \left\{ \cos^{-1} \left[\frac{2r_w^2 - 2r_w \cdot \delta - 2r_w \cdot r_{cond} + \delta^2}{2 \cdot (r_w - \delta) \cdot (r_w - r_{cond})} \right] \right\}}{r_{cond}} \right\}
 \end{aligned} \tag{4.55}$$

To calculate the area of triangle OIG,

$$\begin{aligned}
 A_{OIG} &= \frac{1}{2} OI \cdot OG \cdot \sin(\angle IOG) \\
 &= \frac{1}{2} r_{cond}^2 \cdot \sin \left\{ 2 \cdot \sin^{-1} \left\{ \frac{(r_w - \delta) \cdot \sin \left\{ \cos^{-1} \left[\frac{2r_w^2 - 2r_w \cdot \delta - 2r_w \cdot r_{cond} + \delta^2}{2 \cdot (r_w - \delta) \cdot (r_w - r_{cond})} \right] \right\}}{r_{cond}} \right\} \right\}
 \end{aligned} \tag{4.56}$$

Chapter 4 The Complete Current Propagation Model

Using (4.51), (4.55) and (4.56), the area of shaded area,

$$\begin{aligned}
 A_{shaded} &= A_{OIDG} - A_{OIG} - A_{IHGJ} \\
 &= r_{cond}^2 \cdot \sin^{-1} \left\{ \frac{(r_w - \delta) \cdot \sin \left\{ \cos^{-1} \left[\frac{2r_w^2 - 2r_w \cdot \delta - 2r_w \cdot r_{cond} + \delta^2}{2 \cdot (r_w - \delta) \cdot (r_w - r_{cond})} \right] \right\}}{r_{cond}} \right\} \\
 &\quad - \frac{1}{2} r_{cond}^2 \cdot \sin \left\{ 2 \cdot \sin^{-1} \left\{ \frac{(r_w - \delta) \cdot \sin \left\{ \cos^{-1} \left[\frac{2r_w^2 - 2r_w \cdot \delta - 2r_w \cdot r_{cond} + \delta^2}{2 \cdot (r_w - \delta) \cdot (r_w - r_{cond})} \right] \right\}}{r_{cond}} \right\} \right\} \\
 &\quad - (r_w - \delta)^2 \cdot \cos^{-1} \left[\frac{2r_w^2 - 2r_w \cdot \delta - 2r_w \cdot r_{cond} + \delta^2}{2 \cdot (r_w - \delta) \cdot (r_w - r_{cond})} \right] \\
 &\quad + \frac{1}{2} (r_w - \delta)^2 \cdot \sin \left\{ 2 \cos^{-1} \left[\frac{2r_w^2 - 2r_w \cdot \delta - 2r_w \cdot r_{cond} + \delta^2}{2 \cdot (r_w - \delta) \cdot (r_w - r_{cond})} \right] \right\}
 \end{aligned} \tag{4.57}$$

The total area current flow on the outer surface of the stranded conductors is given by

$$A_{stranded} = N A_{shaded} \tag{4.58}$$

where N is the number of the surface wires of the stranded conductors. The correction factor is given by

$$X_{corr} = \frac{A_{stranded}}{A_w} \tag{4.59}$$

With this correction factor, the final resistance for the wire is

$$R_{w-corr} = X_{corr}^{-1} R_w \tag{4.60}$$

Since the wires are all identical, the per-unit-length DM and CM resistance matrices are obtained as

$$R_{DM} = [2R_{w-corr}] \quad (4.61)$$

$$R_{CM} = \begin{bmatrix} 2R_{w-corr} & R_{w-corr} \\ R_{w-corr} & 2R_{w-corr} \end{bmatrix} \quad (4.62)$$

4.2.2 Inductance

To derive the per-unit-length inductance and capacitance matrices of the power line cable, knowledge of the static electromagnetic analysis needs to be applied. Basically, the determination of inductance is based on magnetostatics, and requires consideration of the magnetic flux ψ_i that penetrates a surface that is parallel to the cable when a current flows [47].

Before deriving the inductances for the power line cable, let us consider the magnetic flux that penetrates a surface S that is parallel to the cable with current I as shown in Figure 4.12.

Figure 4.12. Illustration of the calculation of magnetic flux through a surface.

The cylindrical-coordinate system is used in this case. The edges of the surface are at distances R_1 and R_2 from the cable. $\vec{B}_t$ is the transverse magnetic flux density vector and $\vec{H}_t$ is the transverse magnetic field intensity vector. Thus the total magnetic flux through the surface S is

$$\begin{aligned}\psi_m &= \oiint_S \vec{B}_t \cdot d\vec{s} = \Delta z \int_{R_1}^{R_2} \frac{\mu I}{2\pi r} dr \\ &= \Delta z \frac{\mu I}{2\pi} \ln\left(\frac{R_2}{R_1}\right)\end{aligned}\quad (4.63)$$

so that the per-unit-length magnetic flux is

$$\psi = \frac{\psi_m}{\Delta z} = \frac{\mu I}{2\pi} \ln\left(\frac{R_2}{R_1}\right) \text{ (Wb/m)} \quad (4.64)$$

Figure 4.13. A visual description of the calculation of the per-unit-length self-inductance.

Consider the case of three wires in a homogeneous medium. The level of self-inductance is calculated in the manner depicted in Figure 4.13 by considering the total magnetic flux ψ_i that penetrates the surface parallel to Conductor i and Conductor 0 – when there is no current flowing in Conductor j and a current I_i flows in Conductor i and returns through Conductor 0. This gives us self-inductance L_{ii} [47] as

$$\begin{aligned}
 L_{ii} &= \frac{\psi_i}{I_i} \Big|_{I_j=0} \\
 &= \frac{\mu}{2\pi} \ln \left(\frac{d_{i0}}{r_{wi}} \right) + \frac{\mu}{2\pi} \ln \left(\frac{d_{i0}}{r_{w0}} \right) \\
 &= \frac{\mu}{2\pi} \ln \left(\frac{d_{i0}^2}{r_{wi} r_{w0}} \right)
 \end{aligned} \tag{4.65}$$

Mutual inductance is obtained, as depicted in Figure 4.14, by considering the total magnetic flux ψ_i that penetrates the surface parallel to Conductor i and Conductor 0 – when there is no current flowing in Conductor i and a current I_j flows in Conductor j and returns in Conductor 0.

Figure 4.14. A visual description of the calculation of the per-unit-length mutual inductance.

This gives us the mutual inductance L_{ij} [47] as

$$\begin{aligned}
 L_{ij} &= \frac{\psi_i}{I_j} \Big|_{I_i=0} \\
 &= \frac{\mu}{2\pi} \ln \left(\frac{d_{j0}}{d_{ij}} \right) + \frac{\mu}{2\pi} \ln \left(\frac{d_{i0}}{r_{w0}} \right) \\
 &= \frac{\mu}{2\pi} \ln \left(\frac{d_{i0} d_{j0}}{d_{ij} r_{w0}} \right)
 \end{aligned} \tag{4.66}$$

Again, the skin effects need to be considered. Not only do they cause an increase in wire resistance, they also bring about a decrease in self-inductance. According to [50], the self-inductance for conductor i can be corrected by multiplying it by a correction factor of

$$X_i = \frac{1}{0.315 \times 0.53 \times \frac{r_{wi}}{\delta}} \tag{4.67}$$

where r_{wi} is the radius of conductor i .

The corrected self-inductance is finally obtained as

$$L_{ii\text{corr}} = X_i \cdot \frac{\mu}{2\pi} \ln \left(\frac{d_{i0}}{r_{wi}} \right) + X_0 \cdot \frac{\mu}{2\pi} \ln \left(\frac{d_{i0}}{r_{w0}} \right) \tag{4.68}$$

where X_i is the correction factor for the Conductor i and X_0 is the correction factor for the Conductor 0.

The mutual inductance also needs to be corrected in order to take into account the skin effects. The corrected mutual inductance is finally obtained as

$$L_{ij\text{ corr}} = X_j \cdot \frac{\mu}{2\pi} \ln\left(\frac{d_{j0}}{d_{ij}}\right) + X_0 \cdot \frac{\mu}{2\pi} \ln\left(\frac{d_{i0}}{r_{w0}}\right) \quad (4.69)$$

where X_j is the correction factor for the Conductor j .

The per-unit-length DM and CM inductance matrices are finally written as

$$L_{DM} = [L_{ii\text{ corr}} + L_{jj\text{ corr}} - 2L_{ij\text{ corr}}] \quad (4.70)$$

$$L_{CM} = \begin{bmatrix} L_{ii\text{ corr}} & L_{ij\text{ corr}} \\ L_{ji\text{ corr}} & L_{jj\text{ corr}} \end{bmatrix}_{i=1, j=2} \quad (4.71)$$

4.2.3 Capacitance

The per-unit-length CM capacitance matrices for the power line cable are derived based on the principle of electrostatics. The equivalent coupling effects occurring inside the power line cable are shown in Figure 4.15. C_{pair} is the per-unit-length CM capacitance between any two parallel wires.

Figure 4.15. Cross-sectional view of the power line cable and the per-unit-length capacitance between any two wires.

Figure 4.16. Cross section of two parallel wires and the corresponding equivalent line charges.

Referring to Figure 4.16, the equipotential surfaces of the two wires can be considered to be generated by a pair of line charges $+q$ and $-q$ separated by a distance $(D-2d_i) = d - d_i$ [48]. The potential difference between the two wires is that between any two points on the respective wires.

Let 1 and 2 denote the wires surrounding the equivalent line charges $+q$ and $-q$ respectively. By using path integration, the potential differences between wires 1 and 2 due to charges carried by wires 1 and 2 respectively are

$$V_1 = -\frac{q}{2\pi\epsilon} \ln \frac{a}{d} \quad (4.72)$$

$$V_2 = \frac{q}{2\pi\epsilon} \ln \frac{a}{d} \quad (4.73)$$

Note that V_1 is a positive quantity, whereas V_2 is negative as $a < d$. The per-unit-length capacitance is

$$C = \frac{q}{V_1 - V_2} = \frac{\pi\epsilon}{\ln \frac{d}{a}} \quad (4.74)$$

According to cable geometry, d can be determined by

$$d = \frac{I}{2} \left(D + \sqrt{D^2 - 4a^2} \right) \quad (4.75)$$

From (4.74) and (4.75), we get

$$C_{pair} = \frac{\pi\epsilon}{\ln \left[(D/2a) + \sqrt{(D/2a)^2 - 1} \right]} \text{ (F/m)} \quad (4.76)$$

where ϵ is the permittivity of the dielectric material between the cable conductors.

where D is the distance between any two wires and a is the radius of the wire.

The per-unit-length DM and CM capacitance matrices are given by

$$C_{DM} = [C_{pair}] \quad (4.77)$$

$$C_{CM} = \begin{bmatrix} 2C_{pair} & -C_{pair} \\ -C_{pair} & 2C_{pair} \end{bmatrix} \quad (4.78)$$

4.2.4 Conductance

From [48], if the surrounding medium is homogenous, we can obtain

$$\frac{C}{G} = \frac{\epsilon}{\sigma_d} \Rightarrow G = \frac{\sigma_d C}{\epsilon} \quad (4.79)$$

where σ_d and ϵ are the conductivity and permittivity of the dielectric material respectively and G is the conductance of the wire.

Since the wires of the household power cable shown in Figure 4.15 sit in close proximity to each other, we assume that the dielectric material is homogenous and is made up of a mixture of insulation material and air. Based on this assumption, and from (4.79), we get

$$G_{DM} = \frac{\sigma_d}{\varepsilon} [C_{pair}] \quad (4.80)$$

$$G_{CM} = \frac{\sigma_d}{\varepsilon} C = \frac{\sigma_d}{\varepsilon} \begin{bmatrix} 2C_{pair} & -C_{pair} \\ -C_{pair} & 2C_{pair} \end{bmatrix} \quad (4.81)$$

4.3 EXPERIMENTAL VERIFICATION OF THE CM CURRENT PROPAGATION MODEL

Figure 4.17. Setup for the verification of the CM noise propagation model.

The experimental setup for the verification of the CM noise propagation model is shown in Figure 4.17. The CM source comprises a signal generator, an isolation transformer and two 22 nF ‘Y’ class coupling capacitors. The power line cable used is a single-core PVC insulated, non-sheathed cable with a voltage rating of 450/750 V. Its specifications match those of cables commonly used in the wiring of homes. The power line cable is 3 m in length, and each wire is made up of copper conductors that are stranded and insulated with PVC. The wire has a cross-sectional area of 2.5 mm², and the diameter of a single

copper strand is 0.67 mm (see details in Appendix A.2). The power line cable is terminated with a line impedance stabilization network (LISN) (see details in Appendix A.3), which acts as the balanced load. The frequency range of interest is from 1 to 30 MHz. We limit the scope of the current experiments to scenarios in which the loads are balanced, and the cables are equidistant as they run parallel to each other.

With the properties of the power line cable being defined as such, it becomes possible to calculate the chain parameter matrix. The CM current values for the Live, Neutral and Earth wires at the source end are measured using the radio frequency (RF) current probe which is connected to the spectrum analyzer (see details in Appendices A.4 and A.5 respectively). Given knowledge of the CM source, the CM current values for the Live, Neutral and Earth wires at the load end can be derived and compared to those measured using the RF current probe. This comparison serves as a verification of the CM noise propagation model.

The impedances of the source and the load are measured using a two-current-probe measurement approach. The advantage of using this approach is that it allows measurement of the impedances in the power line network at high voltage levels without damaging the network analyzer. More information on the theory behind the two-current-probe measurement methodology can be found in [51], [52] and Appendix A.1.

Figure 4.18. Configuration of components for measuring impedances using two current probes.

The configuration of the components required to measure impedances using the two-current-probe method is shown in Figure 4.18. The impedances Z_{LN} , Z_{LE} and Z_{NE} are measured when the terminals of the two current probes are connected across the Live and Neutral wires, Live and Earth wires, and Neutral and Earth wires respectively. From the above, we obtain

$$Z_{LN} = Z_1 + Z_2 \quad (4.82)$$

$$Z_{LE} = Z_1 + Z_3 \quad (4.83)$$

$$Z_{NE} = Z_2 + Z_3 \quad (4.84)$$

The relationship between the measured impedances and the network parameters Z_1 , Z_2 and Z_3 is given by

$$Z_1 = \frac{Z_{LN} + Z_{LE} - Z_{NE}}{2} \quad (4.85)$$

$$Z_2 = \frac{Z_{LN} + Z_{NE} - Z_{LE}}{2} \quad (4.86)$$

$$Z_3 = \frac{Z_{LE} + Z_{NE} - Z_{LN}}{2} \quad (4.87)$$

Figure 4.19. Measured impedances of the source.

Figure 4.20. Measured impedances of the load.

Figure 4.19 shows the impedance values of the source which comprises of a signal generator, an isolation transformer and two 22 nF capacitors as shown in Figure 4.17. Z_1 , Z_2 and Z_3 in Figure 4.19 correspond to the values of the impedances Z_{Scm-L} , Z_{Scm-N} , Z_{Scm-E} illustrated in Figure 4.8. From 0 to 6 MHz, the impedances of Z_1 and Z_2 fall from a maximum value of 100 Ω to a minimum of around 0.1 Ω , before rising gradually after the 6 MHz mark. This is a property of a series resonant circuit with its CM capacitive impedance below the resonant frequency and its CM inductive impedance above it.

Figure 4.20 shows the measured impedances of the load, with the load end being terminated with LISN. As the AC power mains vary in their impedance levels over time, the use of LISN as the terminating load serves to stabilize the impedance and the repeatability of measurement results. In Figure 4.20, Z_1 , Z_2 and Z_3 correspond to the values of the impedances of Z_{Lcm-L} , Z_{Lcm-N} , Z_{Lcm-E} of Figure 4.9 respectively. When the graphs in Figures 4.19 and 4.20 are compared, it is obvious that Z_1 and Z_2 are almost equivalent. When Z_1 is not equal to Z_2 , a portion of the injected signal is converted into CM signals [49]. However, given that Z_1 is equal and balanced with Z_2 at the source and load, no additional CM signals will be generated when we inject currents into the Live and Neutral wires. The propagation path of the signals that we inject into the network can thus be monitored. In addition, the value of Z_3 will affect the level of CM current but not the level of DM current as only CM current will flow through it. This will not affect the

symmetry of the network. The implication of this is that if we can increase Z_3 without affecting Z_1 and Z_2 , we can effectively decrease the CM current flowing through the PLC network.

Figure 4.21. Comparison of the measured CM currents in the Live and Neutral wires at the source end

Figure 4.22. Comparison of the measured CM currents in the Live and Neutral wires at the load end

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Since the impedances Z_1 and Z_2 at the source and load ends are equal, we can expect the magnitudes of the current in the Live and Neutral wires to be equal. These are the CM current propagating through the power line cable that we are studying.

In Figure 4.21, the dotted line shows the magnitude of the measured CM current in the Live wire while the solid line shows the magnitude of the measured CM current in the Neutral wire at the source end. In Figure 4.22, the dotted line shows the magnitude of the measured CM current in the Live wire while the solid line shows the magnitude of the measured CM current in the Neutral wire at the load end.

From the graphs in Figures 4.21 and 4.22, we can see that the graphs of the measured CM currents in the Live and Neutral wires are almost overlapping each other at both the source and load ends. This tells us that we have successfully injected CM currents into the Live and Neutral wires at the source end which propagate to the load end.

In Figures 4.23 and 4.24, the sum of the CM currents measured in the Live and Neutral wires are compared with the CM current measured in the Earth wire at both the source and load ends. The comparison shows close agreement between the two measured CM currents. This verifies the CM current propagation model that we proposed in Figure 4.2, even at high frequencies, and demonstrates that the Earth wire is the dominant return path of the CM current in a three-wire cable.

Figure 4.23. Comparison of the sum of the CM currents measured in the Live and Neutral wires with the CM current measured in the Earth wire at the source end of a 3-m cable.

Figure 4.24. Comparison of the sum of the CM currents measured in the Live and Neutral wires with the CM current measured in the Earth wire at the load end of a 3-m cable.

In Figures 4.25 and 4.26, the CM currents measured in the Live and Neutral wires at the source end of a 3 m power line cable are compared with those derived through simulations. The comparison is made based on a CM voltage of $V_{cm}=100$ mV that is injected using a signal generator. The CM currents $I_{cm1}(0)$, $I_{cm2}(0)$, $I_{cm1}(L)$ and $I_{cm2}(L)$ are measured using a current probe (see Figure 4.17). With the impedance of the load measured in the manner depicted in Figure 4.20, it becomes possible to calculate the CM voltage at the load end $V_{cm1}(L)$ and $V_{cm2}(L)$. The CM currents $I_{cm1}(0)$ and $I_{cm2}(0)$ at the source end can be derived using (4.24) with the known values of $V_{cm1}(L)$, $V_{cm2}(L)$, $I_{cm1}(L)$ and $I_{cm2}(L)$ at the load end. The calculated values are then compared with the measured values. The similarity between the simulated and measured values of the CM currents in the Live and Neutral wires is apparent from the graphs in Figures 4.25 and 4.26.

Figure 4.25. Comparison of the measured and simulated CM currents in the Live wire at the source end of a 3-m cable.

Figure 4.26. Comparison of the measured and simulated CM currents in the Neutral wire at the source end of a 3-m cable.

In Figures 4.27 and 4.28, the CM currents measured in the Live and Neutral wires at the load end are compared with the simulated results. Again, a known CM voltage of $V_{cm}=100$ mV is injected using a signal generator and the CM currents $I_{cm1}(0)$, $I_{cm2}(0)$, $I_{cm1}(L)$ and $I_{cm2}(L)$ are measured using a current probe (see Figure 4.17). With the impedance measured at the source (as depicted in Figure 4.19), the CM voltage V_{cm} injected and the CM currents $I_{cm1}(0)$ and $I_{cm2}(0)$ measured, the CM voltages at the source end $V_{cm1}(0)$ and $V_{cm2}(0)$ can be calculated using the basic circuit theorem. Knowing $V_{cm1}(0)$, $V_{cm2}(0)$, $I_{cm1}(0)$ and $I_{cm2}(0)$ at the source end, the CM currents $I_{cm1}(L)$ and $I_{cm2}(L)$ are calculated at the load end using (4.24). The calculated values are then compared with the measured values. The figures show that the graphs of the simulated and measured CM currents match each other very well for the Live and Neutral wires at the load end.

Figure 4.27. Comparison of the measured and simulated CM currents in the Live wire at the load end of a 3-m cable.

Figure 4.28. Comparison of the measured and simulated CM currents in the Neutral wire at the load end of a 3-m cable.

Figure 4.29. Comparison of the measured and simulated CM currents in the Live wire at the load end of a 30-m cable.

Figure 4.30. Comparison of the measured and simulated CM currents in the Neutral wire at the load end of a 30-m cable.

The experiments are repeated using a 30-m long power line cable. Figures 4.29 and 4.30 compare the measured and simulated CM currents in the Live and Neutral wires of a 30-m power line cable. From these figures, it can be seen that the graphs of the CM currents measured in the Live and Neutral wires at the load end are close to those of the

simulated results for the longer cable. Points of maxima and minima in these graphs arise as part of the quarter wavelength effect in the transmission line. The quarter wavelength effect causes such maxima or minima to occur when the cable length is a multiple of the quarter wavelength of the operating signal. This quarter wavelength effect is more visible in the longer cable than the shorter cable. Another observation is that there is a general downward trend in the magnitude of the CM current with increasing operating frequency. For the 3-m long power line cable, the magnitude of the CM currents taken at 1 MHz in both the Live and Neutral wires is around 64 dB μ A. This falls to around 53 dB μ A at 27.5 MHz (see Figures 4.27 and 4.28). This downward trend is also observed in the longer 30-m long cable. The CM current measured at 1 MHz is around 63 dB μ A, and falls to about 46 dB μ A at 27.5 MHz (see Figures 4.29 and 4.30). This downward trend clearly demonstrates that the propagated signals attenuate more at higher frequencies - an observation that is clearly in line with the properties and expected behavior of waves propagating through a transmission line.

The above six graphs (Figures 4.25 to 4.30), which compare measured CM currents with simulated CM currents, indicate that the two CM currents are close to one another. From these results, it is shown that the proposed CM noise propagation model can accurately predict the CM currents at the load end for any length of power line cable when the CM currents at the source end are known, and vice versa.

4.4 EXPERIMENTAL VERIFICATION OF THE COMPLETE MODEL

Figure 4.31. Experimental setup.

The experimental setup for verification of the complete model is shown in Figure 4.31. The DM and CM sources comprise of a signal generator, an isolation transformer and a 22 nF ‘Y’ class coupling capacitor. The power line cable used is a single-core PVC insulated, non-sheathed cable with a voltage rating of 450/750 V. Its specifications match those of cables commonly used in the wiring of homes. The power line cable is 3-m in length, and each wire is made up of copper conductors that are stranded and insulated with PVC. The wire has a cross-sectional area of 2.5 mm^2 , and the diameter of a single copper strand is 0.67 mm. The power line cable is terminated with a LISN, which acts as the balanced load. The frequency range of interest is from 1 to 30 MHz. The signal generator is configured to inject a 100 mV into the Live and Neutral wires. We limit the scope of the current experiments to scenarios in which the loads are balanced, and the cables are equidistant as they run parallel to each other.

Figure 4.32. Measured currents at the source end.

Figure 4.33. Measured currents at the load end.

With the properties of the power line cable being defined as such, it becomes possible to calculate the chain parameter matrix. The currents I_L , I_N and I_E , flowing in the Live,

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Neutral and Earth wires at the source and load ends are measured using the RF current probe. The measured currents at the source and load ends are shown in Figures 4.32 and 4.33 respectively. Given knowledge of the DM and CM sources, the DM and CM current values for the Live, Neutral and Earth wires at the load end can be derived and compared to those measured using the RF current probe. This comparison serves as a verification of the DM and CM current propagation model.

We know that the DM input voltage at the source end to be 100 mV. After measuring the currents I_L , I_N and I_E at the source and load ends (see Figures 4.32 and 4.33) with the RF current probe and using (4.1), (4.2) and (4.3), we can obtain the DM and CM currents at the source and load ends. This gives us $V_{dm}(0)=100$ mV, $I_{dm}(0)$, $I_{dm}(L)$, $I_{cm1}(0)$, $I_{cm2}(0)$, $I_{cm1}(L)$ and $I_{cm2}(L)$.

Using the two-current-probe measurement methodology, we can find Z_{Ldm} , Z_{Lcm-L} , Z_{Lcm-N} , Z_{Lcm-E} in which $Z_{Ldm}=Z_{LN}$, $Z_{Lcm-L}=Z_1$, $Z_{Lcm-N}=Z_2$ and $Z_{Lcm-E}=Z_3$. With the knowledge of these currents and impedances, we can find the voltages $V_{dm}(L)$, $V_{cm1}(L)$ and $V_{cm2}(L)$ at the load end.

With the properties of the cable known, we can calculate the chain parameter matrices for the DM and CM current propagation models. Knowing $V_{dm}(0)=100$ mV, $I_{dm}(0)$ and $I_{dm}(L)$, we can verify the DM current propagation model by comparing the currents at the load end. Figure 4.34 shows the comparison of the measured and simulated currents at the load end. The two graphs almost close to each other serves as a verification for the DM current model.

Figure 4.34. Comparison of the measured and simulated DM currents I_{dm} in the Live wire at the load end.

For the CM current propagation model, we have $I_{cm1}(0)$, $I_{cm2}(0)$, $I_{cm1}(L)$ and $I_{cm2}(L)$. After the impedances of the LISN are known, we can find $V_{cm1}(L)$ and $V_{cm2}(L)$ in which $V_{cm1}(L) = (I_{cm1}(L) + I_{cm2}(L)) * Z_3 + I_{cm1}(L) * Z_1$ and $V_{cm2}(L) = (I_{cm1}(L) + I_{cm2}(L)) * Z_3 + I_{cm2}(L) * Z_2$ (see Figure 4.20). The unknowns are $V_{cm1}(0)$ and $V_{cm2}(0)$. To verify the CM current propagation model, we compare the CM currents at the source end. Since the LISN is balanced, $V_{cm1}(L)$ and $V_{cm2}(L)$ are almost equal and $I_{cm1}(L)$ and $I_{cm2}(L)$ are almost equal, so we just need to compare the simulated and measured currents for $I_{cm1}(L)$. Figure 4.35 shows the comparison of the measured and simulated CM currents in the Live wire at the source end. The two graphs almost matching each other serves as a good verification for the CM current propagation model. Once we have verified the model, we can use this model to find the CM voltages at the source end by using the chain parameter matrix, $I_{cm1}(L)$, $I_{cm2}(L)$, $V_{cm1}(L)$ and $V_{cm2}(L)$. The CM voltages at the source end are computed and shown in Figure 4.36.

Figure 4.35. Comparison of the measured and simulated CM currents in the Live wire at the source end.

Figure 4.36. Simulated CM voltage at the source end.

From these results, it is shown that the proposed DM and CM current propagation model can accurately predict the DM and CM currents at any position of the power line cable when the DM and CM voltages and currents at the source end or load end are known.

Figure 4.37. DM and CM currents in the Live and Neutral wires in a PLC system.

Figure 4.37 shows the DM and CM currents in the Live and Neutral wires in a PLC system. As long as we can calculate the DM and CM currents at any position of the power line cable, we can determine the maximum values of the DM and CM currents. Using these values, we can determine the worst case electric field intensity at a distance from the power line cable using the following formulae [48]:

$$E_{DM-max} = 1.316 \times 10^{-14} \frac{I_{DM-max} f^2 L d}{R} \text{ (V/m)} \quad (4.88)$$

$$E_{CM-max} = 1.257 \times 10^{-6} \frac{I_{CM-max} L f}{R} \text{ (V/m)} \quad (4.89)$$

where f is the frequency of the current, L is the length of the power line cable, d is the distance between the Live and Neutral wires, and R is the distance of the point of measuring from the power line cable.

These two equations show the relative contributions of the DM and CM currents to radiated emissions. From the two formulae, we can see that radiated emissions due to CM currents in PLC networks can well exceed those due to DM currents. From (4.88) and (4.89), it can be seen that

$$\frac{E_{CM-max}}{E_{DM-max}} = 9.552 \times 10^6 \frac{I_{CM-max}}{I_{DM-max}} \frac{1}{f} \frac{1}{d} \quad (4.90)$$

Usually the conduits in in-house wirings are not greater than 25 mm. The wires are usually just pulled through the conduits and are not tied together tightly and hence the separations between wires are not uniform. This means that the spacing between wires (Live and Neutral) would range from 0, where the wires are touching, to a maximum of 25 mm where the wires are at a maximum distance. With this non-uniformity in the separations between the wires, complete cancellation of the electromagnetic fields is not possible. From (4.90) it can be seen that the contribution to radiated emissions due to DM current increases with increasing frequency.

Figure 4.38. Comparison of the field strengths generated by DM and CM currents.

Figure 4.38 shows the electric fields generated by DM and CM currents at a distance of 3 m from the power line cable. From this figure, it is obvious that the main contribution for the EMI radiation is from the CM current and that of the DM current is negligible in comparison. It is noted that the radiation due to CM current is much greater than that of those contributed by DM current in the frequency range of 10 MHz to 15 MHz, where the peak occurs at 12 MHz. From Figure 4.34, it is observed that the DM current at 10 MHz decreases from 60 dB μ A to around 55 dB μ A at 15 MHz. And from Figure 4.35, it is observed that the CM current at 10 MHz increases from 20 dB μ A to a maximum of 40 dB μ A at 12 MHz and decreases back to around 20 dB μ A from 15 MHz onwards. The observations from Figures 4.34 and 4.35 together with the mathematical relationship of (4.90) explain for the shape of the graph in Figure 4.38.

Chapter 5

INJECTION TECHNIQUES

In an attempt to reduce the level of radiation emitted from the PLC network, we study the possibility of achieving it by looking at how the power line modem (PLM) works in greater detail. In this chapter, we are exploring the possibility of using new signal injection techniques that require the presence of the Earth wire in order to mitigate the level of EMI radiation.

5.1 DIFFERENT INJECTION TECHNIQUES

This section shows the different kinds of available signal injection techniques. The normal injection technique is the one used in all PLMs currently. The alternative injection technique was proposed by N. Korovkin, E. Marthe, F. Rachidi and E. Selina in which they managed to achieve a reduction of up to 20 dB in the radiation level [46]. The last new injection technique is proposed by us. The comparison of the three different injection techniques is presented in the next section.

5.1.1 Normal injection technique

Currently, typical modems have 3.3 V_{pp} transmission output signal which is injected into the Live and Neutral wires with driving current of 0.5 App [53]. Figure 5.1 shows the components of the DM and CM currents flowing in the wires when a PLC signal of S_I is injected into the Live and Neutral wires. The time-domain mathematical expression of S_I is given by

$$S_1 = S(t)e^{j2\pi ft + j\theta} \quad (5.1)$$

where $S(t)$ is the magnitude, f is the frequency and θ is the phase of the injected signal.

Since DM currents flow in the opposite direction, the electromagnetic fields generated by the currents cancel out each other. However, CM currents flow in the same direction, causing the electromagnetic fields to add up and resulting in significant levels of electromagnetic interference (EMI) radiation, even if only a low level of CM current is flowing.

Figure 5.1. Illustration of DM and CM current components resulting from a typical injection method.

5.1.2 Alternative injection technique

Korovkin in [46] proposed an alternative injection technique that took advantage of the Earth wire. To the best of our knowledge, this is the only method proposed in the literature to mitigate indoor radiation. In their method, they make use of the auxiliary signal (or compensation signal). The wanted PLC signal is injected into the Live and Neutral wires and an auxiliary PLC signal that is of the same amplitude but 180° out-of-phase is injected into the Earth and Neutral wires. Mathematically, if the original PLC signal S_1 that is injected into the Live and Neutral wires is given by (5.1) then the signal S_2 injected into the Earth and Neutral wires will be below

$$S_2 = S(t)e^{j2\pi ft + j\theta + j\pi} \quad (5.2)$$

Figure 5.2 shows the components of the DM and CM currents flowing in the wires when a PLC signal S_1 is injected into the Live and Neutral wires and a PLC signal S_2 is injected

into the Earth and Neutral wires. When a PLC signal S_2 is injected into the Earth and Neutral wires, the effect is the same as injecting a PLC signal $-S_1$ into the Earth and Neutral wires. When a PLC signal S_1 is injected into the Live and Neutral wires, a DM current flows from the Live wire and returns in the Neutral wire. When a PLC signal S_2 of the same amplitude but 180° out-of-phase is injected into the Earth and Neutral wires, a DM current flows from the Neutral wire and returns in the Earth wire. When the three wires run in parallel and in close proximity to one another, the electromagnetic fields generated by DM currents cancel out each other since the currents flow in the opposite direction. The electromagnetic fields generated by CM currents are of similar amplitudes and approximate 180° out-of-phase. This results in partial or total cancellation of the electromagnetic fields.

Figure 5.2. Illustration of DM and CM current components resulting from an alternative injection method.

5.1.3 New injection technique

Using the above theory of the cancellation of CM currents and hence electromagnetic fields, we propose a new injection technique. Our new injection technique also requires the use of the Earth wire. In our new injection technique, we inject a PLC signal S_3 into the Live and Earth wires and a PLC signal S_4 that is of the same amplitude but 180° out-of-phase into the Neutral and Earth wires, as shown in Figure 5.3. When injecting a PLC signal S_4 into the Neutral and Earth wires, the effect is the same as injecting a PLC signal

$-S_3$ into the Neutral and Earth wires. In our method, the magnitudes of the signals S_3 and S_4 are only half those of S_1 and S_2 . We have S_3 and S_4 as follows:

$$S_3 = \frac{1}{2} S_1 = \frac{1}{2} S(t) e^{j2\pi ft + j\theta} \quad (5.3)$$

$$S_4 = \frac{1}{2} S_2 = \frac{1}{2} S(t) e^{j2\pi ft + j\theta + j\pi} \quad (5.4)$$

In the normal injection method, if we want a signal S_I across the Live and Neutral wires as shown in Figure 5.1, we need to inject S_I into the Live and Neutral wires so that $V_{LN} = S_I$. This is also so for the alternative injection method proposed by other authors. But in our new injection technique, we only need to inject $\frac{1}{2} S_1$ into the Live and Earth wires and $\frac{1}{2} S_2$ into the Neutral and Earth wires. We can show in Figure 5.3 that $V_{LN} = S_I$ by using the following equations:

$$V_{LN} = V_{LE} + (-V_{NE}) = V_{LE} - V_{NE} \quad (5.5)$$

$$V_{LE} = S_3 = \frac{1}{2} S_1 \quad (5.6)$$

$$V_{NE} = S_4 = \frac{1}{2} S_2 \quad (5.7)$$

$$-S_1 = S_2 \quad (5.8)$$

We can get

$$V_{LN} = \frac{1}{2} S_1 - (-\frac{1}{2} S_1) = S_1 \quad (5.9)$$

In Figure 5.3, the components of the DM and CM currents are flowing in the wires when a PLC signal S_3 is injected into the Live and Earth wires and a PLC signal S_4 of the same amplitude but 180° out-of-phase is injected into the Neutral and Earth wires. Using our proposed method, there will be a reduction in the level of radiation as CM currents will cancel out each other. As compared to the alternative injection method proposed by other

authors, our new injection method is expected to produce a lower level of radiation as the voltages injected into the power lines are only half those of theirs. This also means that the power required is lower leading to saving in power consumption. As the injection voltages are lower, we will expect a lower level of current flowing in the Earth wire.

Figure 5.3. Illustration of DM and CM current components resulting from a new injection technique.

In the next section, we will conduct experiments to compare the performances of the different injection methods in reducing the EMI radiation. From the results of the experiments, it is shown that the proposed new injection method is a possible effective EMI reduction technique to suppress radiated emissions from PLC networks in indoor environment.

5.2 EXPERIMENTAL SETUP

This section shows the three different experimental setups. Experiments are carried out based on the different injection techniques for cases in which the loads are balanced and unbalanced. The obtained results are compared and discussed in the next section.

Figure 5.4. Measurement setup for normal injection technique.

Figure 5.5. Measurement setup for alternative injection technique.

Figure 5.6. Measurement setup for new injection technique.

The experimental setups for our studies are shown in Figures 5.4 to 5.6. In all these experiments, we need a voltage of 3 Vpp across the Live and Neutral wires as this is close to the realistic modem operation voltage. The test bed comprises a 3-m long power line cable consisting of Live, Neutral and Earth wires. Each of these wires is made up of stranded copper conductors with PVC insulation. Each wire has a cross-sectional area of 2.5 mm^2 and the diameter of a single copper strand is 0.67 mm. A dual channel Tektronix AFG 3022 function generator is used to inject signals at the transmission end (Tx End) of the power line cable. The power line cable is terminated with a line impedance stabilization network (LISN) which acts as the load at the receiving end (Rx End). The LISN is made balanced by connecting a 50Ω load across the Live to Earth circuit and Neutral to Earth circuit each. It can be made unbalanced by connecting a 50Ω load across the Live to Earth circuit and not connecting anything to the Neutral to Earth circuit.

A Schaffner Chase EMC CSP-8425-1 RF current probe (100-Hz-to-100-MHz bandwidth) is employed to measure the CM current at the Tx End of the power line cable. This current probe allows measurement of microamp levels of RF current even in the presence of high-level power current.

Figure 5.4 shows the measurement setup for the injection technique used by a typical PLC modem nowadays. In this setup, the magnitude of the injected voltage is 3 Vpp and it is injected into the Live and Neutral wires.

Figures 5.5 and 5.6 show the measurement setups for the alternative injection technique and the new injection technique respectively taking into consideration the Earth wire. In Figure 5.5, a 3 Vpp voltage is injected into the Live and Neutral wires and a voltage of 3 Vpp which is similar to the voltage injected into the Live and Neutral wires but with an opposite phase is injected into the Earth and Neutral wires. In Figure 5.6, a 1.5 Vpp voltage is injected into the Live and Earth wires and a voltage of 1.5 Vpp is injected into Neutral and Earth wires.

The electric field is measured using an active rod E-field antenna model 3301B (30 Hz to 50 MHz) placed 3 m away from the power line cable under test.

5.3 MEASUREMENT RESULTS AND DISCUSSIONS

Figure 5.7. Comparison of the radiated electric fields for the three injection techniques for the balanced case.

Figure 5.8. Comparison of the radiated electric fields for the three injection techniques for the unbalanced case.

Figures 5.7 and 5.8 show the comparison of the radiation levels for the three different injection techniques for the balanced and unbalanced cases. As for the balanced case in Figure 5.7, the alternative injection technique only results in a small reduction in the level of radiation. As for the new injection technique, it allows for a minimum of 20 dB reduction in the level of radiation. As for the unbalanced case in Figure 5.8, the alternative injection technique and the new injection technique only result in a small reduction in the level of radiation. It is also observed that the both methods of injection may even cause higher levels of radiation in the frequency range of 10 to 15 MHz as compared to the normal one.

Figure 5.9. Comparison of the radiated electric fields between the balanced and unbalanced cases for the normal injection technique.

Figures 5.9 to 5.11 show the comparison of the radiation levels for the balanced and unbalanced cases for the three different injection techniques. From Figures 5.9 and 5.10, it can be seen that the level of radiation is not affected even if the loads are not balanced for the normal injection technique and the alternative injection technique. However from Figure 5.11, it is obvious that by having balanced loads, the level of radiation is much lower than that of having unbalanced loads when using the new injection technique. A

maximum of around 35 dB difference can be seen at 5 MHz for the new injection technique when comparing the balanced and unbalanced cases.

Figure 5.10. Comparison of the radiated electric fields between the balanced and unbalanced cases for the alternative injection technique.

Figure 5.11. Comparison of the radiated electric fields between the balanced and unbalanced cases for the new injection technique.

Figure 5.12. Comparison of the CM currents for the alternative injection technique.

Figure 5.13. Comparison of the CM currents for new injection technique.

Figures 5.12 and 5.13 show the comparison of the CM currents for the balanced and unbalanced cases for the alternative injection technique and the new injection technique. From Figure 5.12, it can be seen that the CM current flowing in the Live and Neutral (LN) wires is equal to that flowing in the Earth wire for the balanced case (the upper two graphs). This is also true for the unbalanced case (the upper two graphs). This means that

by using this injection technique, the CM current flowing in LN wires returns completely in the Earth wire. It also means that whether the load is balanced or not, there is not much difference in the level of the CM current flowing in the LN and Earth wires. The total amount of the CM current in the Live, Neutral and Earth (LNE) wires for the balanced case is also equal to that for the unbalanced case (the lower two graphs).

From Figure 5.13, it can be seen that the CM current in the LN wires is equal to that in the Earth wire for the unbalanced case. The CM current in the LN wire is almost equal to that in the Earth wire for the balanced case. But the CM current in the LN wires is lower for the balanced case than that for the unbalanced case. This means that by using this injection technique, the CM current flowing in LN wires returns completely in the Earth wire. Also, for balanced case, it allows for a lower CM current as compared to the unbalanced case. It can also be seen that the CM current in the LNE wires is lower for the balanced case as compared to the unbalanced case.

Figures 5.14 and 5.15 show the comparison of the CM currents in the LNE wires for the three different injection techniques for the balanced and unbalanced cases. In Figure 5.14, the alternative injection technique only results in a small reduction in the CM current in the LNE wires as compared to the normal injection technique. The reduction in the CM current is much lower for the new injection technique. In Figure 5.15, the alternative injection technique and the new injection technique only result in a small reduction in the CM current in the LNE wires for most of the frequencies. However, in the frequency band of 10 to 15 MHz, it can be seen that the CM current increases when these two techniques are used. This is the same conclusion that we arrived at Figures 5.7 and 5.8 when we compared the radiation levels. This suggests that there is a relationship between the total CM current in the LNE wires and the resulting radiation level.

Figures 5.16 and 5.17 show the comparison of the CM currents in the Earth wire for the three different injection techniques for both the balanced and unbalanced cases. From these two figures, it can be seen that for both the balanced and unbalanced cases, the CM current in the Earth wire is higher when the alternative injection technique is used as

compared to the normal injection technique. However, the new injection technique allows for a lower CM current in the Earth wire in the frequency band of 5 to 20 MHz for the balanced case and in the frequency band of 7.5 to 25 MHz for the unbalanced case. This lower CM current in the Earth wire may result in a lower radiation level in the actual system.

Figure 5.14. Comparison of the CM currents in the Live, Neutral and Earth (LNE) wires for the three injection techniques for the balanced case.

Figure 5.15. Comparison of the CM currents in the Live, Neutral and Earth (LNE) wires for the three injection techniques for the unbalanced case.

Figure 5.16. Comparison of the CM currents in the Earth wire for the three injection techniques for the balanced case.

Figure 5.17. Comparison of the CM currents in the Earth wire for the three injection techniques for the unbalanced case.

In summary, the proposed new injection technique produces significant reduction in the CM currents compared with the two existing methods especially for the balanced case. This is due to the amount of voltage that is required to be injected into the system. As illustrated in the previous section, a voltage of 3 V_{pp} is required to be injected into the

Live and Neutral wires. Using the normal injection technique and the proposed new injection technique, only a voltage of 3 Vpp is required. However, for the alternative injection technique, a total voltage of 6 Vpp is required. This higher amount of voltage injected may be one of the reasons why the CM currents are higher compared with the new injection technique. As for the normal injection technique, though the total amount of voltage injected is the same as the new injection technique, there is no cancellation effect of the electromagnetic fields.

5.4 FEASIBILITY OF UTILIZING THE EARTH WIRE FOR SIGNAL INJECTION

Figure 5.18. Comparison of the impedances for the Live to Earth circuit, the Neutral to Earth circuit and the Live to Neutral circuit in the power mains.

Experimental results obtained from a PLC test bed have shown that both the alternative injection technique and the new injection technique can achieve better reduction in the radiation level, especially for the balanced case. The impedance of the Live to Earth circuit and that of the Neutral to Earth circuit of the power mains are usually balanced. Figure 5.18 shows a typical impedance measurement from the power mains using the two-current-probe measurement approach. It can be seen that the impedance of the Live to Earth circuit and that of the Neutral to Earth circuit are almost equal. The impedance of the Live to Neutral circuit is usually different from that of the Live to Earth or Neutral to Earth circuit. With the power mains impedance almost balanced, both the alternative injection technique and the new injection technique can be good methods to reduce radiation from PLC networks. The new injection technique tends to produce better results in reducing radiation from PLC networks compared with the alternative method. The main reason is that when the loads are balanced, the impedances between the Live to Earth circuit and the Neutral to Earth circuit are equal. When the loads are balanced, using the new injection technique (Earth wire being the reference wire) with a voltage injected between the Live and Earth wires and the Neutral and Earth wires, the system is completely balanced. For the alternative injection technique (Neutral wire being the reference wire), the injection of the voltage is between the Live and Neutral wires and the Earth and Neutral wires, the system is not completely balanced. This unbalance in the system does not allow the electromagnetic fields to be totally cancelled compared with the new injection technique. This explains why the new injection technique produces better results.

Also, in terms of the CM current in the Earth wire, it is higher for the alternative injection technique as compared to the new injection technique.

In most countries, household power supply is usually single-phase electric power. In this type of electrical system, the Live wire brings alternating current into the house and the Neutral wire returns it to the power supply. Many plugs and sockets also include a connection to the Earth wire which carries current in case of an insulation fault in the connected equipment. Figure 5.19 shows the connection of the Live, Neutral and Earth

wires to the electrical equipment. The Earth wire is connected to the exposed metal frame of the electrical equipment. The purpose of the Earth wire is to hold all the exposed metal frames in a building to the same voltage potential to prevent electric shocks when touching two metal frames at the same time.

Knowing the purposes of the Earth wire, we understand that we should not inject any voltage into the Earth wire as this may cause a difference in voltage potential between two metal frames and hence leading to a possibility of an electric shock when a person touches them. We also understand that during normal operation of any electrical equipment, we should not have high current flowing in the Earth wire as it may lead to the tripping of the residual current circuit breaker. With these understandings, we analyze our proposed new injection method to see if it compromises the safety of the electrical system. For our new injection method, a signal voltage of 1.5 V_{pp} is injected into the Live and Earth wires and a compensating signal voltage of 1.5 V_{pp} into the Neutral and Earth wires.

Figure 5.19. Connection of the Live, Neutral and Earth wires to the electrical equipment.

Here, the Earth wire is used as the reference, which is at zero voltage (i.e. earth potential). There will not be any voltage injected into the Earth wire. From the voltage point of view, our proposed injection method is definitely safe.

In Singapore, the Earth wire usually runs parallel and close to the Live and Neutral wires. The implementation of the new injection technique is a good solution in reducing the level of radiation. However, this technique cannot be applied throughout the world. In

some countries, the only available power plugs in used are the Type C (European 2-pin plugs). This type of plug is unearthed and has two round, 4 mm pins [54]. Examples of such countries which use only such plugs are Egypt, Iran, North Korea, Somalia and some other developing countries [55]. For these countries, the alternative and the new injection techniques cannot be deployed at all due to the absence of the Earth wire. Also, in some developing countries, due to absence of strict and proper regulations, the Earth wire may not even be properly connected in some cases. In these countries, even if we employ the alternative or the new injection techniques, it may not lead to reduction in the radiation level.

Chapter 6

INVESTIGATION OF COMMON-MODE COUPLING PATH IN POWER LINE MODEMS

The power line network has been viewed as a convenient and inexpensive communication medium for data and voice transmissions but its unbalance nature leads to the generation of common-mode (CM) noise currents in the power line network. This in turn generates electromagnetic interference (EMI) radiation. The resulting EMI radiation is a concern as it interferes with the existing wireless communications users. Since one has no control over the imbalances of the power lines caused by the electrical loads connected to them, we can only look at the RF source (the PLC modem) to solve the problem. One possible way to control the EMI radiation is to ensure that the PLC modem has very high CM impedance. With high CM impedance, the CM current propagating through the power line network is kept to as low a level as possible.

Many works have been done to reduce the CM noise current propagating over the analog front end (AFE) onto the power line network. This can be achieved by inserting a CM choke in series with the isolation transformer or replacing the AFE with an optical isolation circuitry [52], [53] to increase the CM source impedance. There is also an attempt to use the CM choke to reduce the transmission loss deviation resulting from different states of electrical appliances by simply inserting the CM choke between the AC mains line (or modem) and the electrical appliances [56]. The aim of this chapter is to investigate the effectiveness of reducing CM noise current generated in the power line by increasing the CM impedance across the power supply circuitry in a PLC modem. Measurements are carried out to verify that there is a reduction in radiated emissions due to a reduction in CM noise current in the power line network. This is important as it will provide essential information for researchers to explore more EMI mitigation techniques to suppress the radiated emissions from the power line networks.

6.1 IDENTIFICATION OF THE COMMON-MODE COUPLING PATH

Figure 6.1. Building blocks of a PLC modem.

Figure 6.1 shows the basic building blocks of a PLC modem. As seen from the figure, other than the AFE, the power supply building block also shares the same power line. The AFE uses the power line as a channel to communicate whereas the power supply block uses the power line to power up the various circuitries in the PLC modem. Therefore, it is suspected that other than the AFE, it is possible that the CM noise current can couple through the power supply building block onto the power line.

Figure 6.2. Power line modem system block diagram.

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At the AFE, an isolation transformer and a class X2 capacitor are used to couple the PLC signals onto the power line without any physical connection to the mains. They also serve to protect and isolate the modem's circuitry from the 230V AC. The wideband DM data signal flows from the power line through the isolation transformer and the class X2 capacitor. In the power supply building block, a power transformer and a voltage rectifying circuit are present. The voltage is being stepped down by the power transformer to a level that is able to be used by the digital circuitries in the power line modem.

DM signal flows between Live and Neutral wires, and electrically isolated from the Earth wire. In the ideal scenario, there will not be any CM current flowing as the CM impedance from modem to Earth is infinite, even if a CM noise source is present. However, in practice, there will always be imperfection in winding a transformer and this gives rise to parasitic inter-winding capacitance formed between the primary and secondary windings of the power transformer. Due to this parasitic capacitance [57], a CM noise path is formed between the digital signal processing circuitry and the power line network. This results in the CM noise current being able to propagate onto the power line via the parasitic capacitance between the primary and secondary windings of the transformer. The parasitic capacitance provides a low impedance path between the power line modem and the isolated Earth line at higher frequency range. As the CM noise current propagates over the power lines, an increased level of radiated emission is expected.

In general, the generation of CM current in the power lines is best explained by an equivalent circuit model consisting of a transmitting PLC modem, a receiving PLC modem, the power line network as propagation medium and some electrical appliances connected to the power line network as loads, as shown in Figure 6.3. The transmitting modem has a differential voltage source (V_{DM}) between its Live and Neutral wires with source impedance Z_{TXDM} . Similarly, the receiving modem is represented by the differential impedance Z_{RXDM} between its Live and Neutral wires. The transmitting modem excites a differential signal while the receiving modem receives the signal across

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the Live and Neutral wires. Both the transmitting and receiving modems are not physically connected to the Earth wire. However, at high frequency, the parasitic capacitances represented by Z_{TXCM} and Z_{RXCM} within the two modems provide noise propagation path between the modems and the Earth wire. Also, there are mutual couplings between the Live, Neutral and the Earth wires, which are due to mutual inductances and capacitances between the three lines. In the frequency range from 1 to 30 MHz, mutual inductive couplings between the wires are usually more dominant compared to mutual capacitive couplings.

Figure 6.3. Equivalent circuit model of a basic PLC network.

The equivalent electrical model of the transmitting modem can be represented by two differential voltage sources of $V_{DM}/2$, and two source impedances of $Z_{TXDM}/2$ each, as illustrated in Figure 6.3. The imbalance due to the power line network (usually the existing electrical appliances connected to the network) causes a CM noise voltage, which can be represented by an equivalent CM voltage source V_{CM} .

The parasitic impedances, Z_{TXCM} and Z_{RXCM} , within the modems provide CM paths between the modems and the Earth wire. As shown in Figure 6.4, the parasitic impedances, which usually are high impedance in nature, provide finite impedances at high frequency. These parasitic impedances present between the power supply and the power line as well as between the analog front end (AFE) and the power line.

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Figure 6.4. Building blocks of a PLC modem with parasitic impedances.

Figure 6.5. CM circuit model of a PLC modem.

Figure 6.5 is the proposed CM noise model for existing PLC modems. To simplify the analysis, it is assumed that there is only one CM noise source in a PLC modem, which is V_{CM} . There are many parasitic capacitances in the PLC modems. C_1 , C_2 , C_3 , C_4 , C_5 and C_6

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are only some of the parasitic capacitances shown. We shall only limit our discussion to C_2 , C_3 and C_4 as they are in the main paths in which CM noise currents are coupled onto the AC mains. As we can see, V_{CM} originates from the AFE whereby the data is coupled onto the power line. Due to C_2 , some of the V_{CM} is coupled onto the secondary side of the isolation transformer where the wanted signals start propagating. Some of the noise current can also be coupled onto the secondary side of the isolation transformer via capacitive coupling from C_3 and C_4 . Besides these paths, there may be other paths available for the coupling of the V_{CM} onto the secondary side of the isolation transformer. The noise currents appearing on the secondary side of the isolation transformer is the main cause of radiated emissions.

Figure 6.6. CM impedance of the power transformer.

Figure 6.7. CM impedance of the isolation transformer.

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The measured inter-winding capacitances of the power transformer and the isolation transformer respectively are shown in Figures 6.6 and 6.7. The maximum impedance of the power transformer and isolation transformer are 100 k Ω and 1000 k Ω respectively. These impedances are parasitic in nature with parasitic capacitance value of 10 pF for power transformer and 1 pF for isolation transformer.

Using simple circuit analysis, we see that as impedance is increased, the amount of current flow will be reduced. Since the power transformer and isolation transformer are put in parallel with each other, its equivalent impedance will be almost equal to the smaller impedance of 100 k Ω which is the impedance of the power transformer. Therefore, the CM impedance between the power lines and CM noise source is smaller when power transformer is inside the modem. If the power transformer is removed from the modem, the CM impedance between the power lines and CM noise source will increase. CM noise current that is propagating through the isolation transformer will decrease because of the increased CM impedance. The reduction of conducted emission will lead to the reduction of radiated emission.

A power line network test bed is set up to analyze the levels of conducted and radiated emissions due to change in impedance in the power supply network. Experimental measurements are carried out to measure and compare the conducted and radiated emissions due to the inter-winding capacitance across the primary and secondary windings of the power transformer. This is achieved through the removal of the power transformer and relying on batteries to power up the PLC modems. Measurements were taken in a semi-anechoic chamber in the School of Electrical and Electronic Engineering, Nanyang Technological University to measure the conducted emissions with a clamp-on current probe and the radiated emissions with a loop antenna and a rod antenna.

6.2 EXPERIMENTAL VERIFICATION OF CM CURRENT REDUCTION BY REMOVING THE POWER TRANSFORMER

Figures 6.8 and 6.9 show the experimental setup of the test bed for measuring the conducted and radiated emissions respectively. The experimental setup is the same for both cases except the equipment used to measure the noise emitted from the power line network. The test bed is constructed in a semi-anechoic chamber. In order not to complicate the analysis, the test bed is a straight 3-m long power line cable consisting of Live, Neutral and Earth wires. Branches are created at every 1 m such that loads can be inserted at every 1 m of the test bed. Loads 1 and 2 have an impedance of $200\ \Omega$ between the Live and Earth wires and between the Neutral and Earth wires respectively. Load 3 has an impedance of $150\ \Omega$ between the Live and Earth wires and another $15\ \Omega$ between the Neutral and Earth wires. Load 3 is inserted to create an imbalance in the power line test bed to facilitate the generation of CM current in the power line. Loads 1 and 2 are required to ensure that the impedance of the power line does not become more dominant than the imbalance created by Load 3. This will also ensure that the impedance of the test bed is comparable to the average impedance of the actual power line network.

Figure 6.8. Setup of test bed for measuring conducted emissions.

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Figure 6.9. Setup of test bed for measuring radiated emissions.

The measurements on the CM noise current are carried out using the Schaffner Chase EMC CSP-8425-1 RF clamp-on current probe, the EMCO 3301B rod antenna (see details in Appendix A.5) and the EMCO 6509 passive loop antenna to measure the CM noise current, E-field and H-field generated respectively [58]. The spectrum analyzer used for the measurements is the Agilent 4395A Network/Impedance/Spectrum Analyzer. The model of PLC modems we are using is HL100E HomePlug Ethernet Adaptor (see Appendix A.6 for more details). The measurements using the antennas are carried out at a distance of 3 m from the centre of the test bed. The clamp-on current probe is placed at the center of the test bed for measuring the CM noise current in the test bed. In this research, only the measured worst-case radiated emissions using the passive loop and the active rod antennas are shown. Communication between the transmitter and the receiver of the PLC modems is carried out using a developed program. The program allows pre-specified bytes of data to be sent out continuously from the transmitting workstation to the receiving workstation. The “max-hold” option of the spectrum analyzer is turned on such that the measured data can be assumed to be equivalent to a peak detector.

6.3 MEASUREMENT RESULTS AND DISCUSSIONS

The measured data presented are obtained after adjustments have been made to the stored data from the spectrum analyzer with the transfer impedance/conductance and the antenna factor of the probe and antenna.

Figure 6.10. Measured current in the Live wire.

Figure 6.10 shows the spectrum for the net current transmitted in the Live wire. As can be seen, the net current is almost constant throughout the transmission band. This will allow us to analyze the effects of the CM noise current generated throughout the transmission spectrum from 4 to 21 MHz. According to the EMI analysis, the amount of currents flowing in a pair of wires can be decomposed into CM and DM current components as shown in (6.1) and (6.2).

$$I_L = I_{DM} + I_{CM} \quad (6.1)$$

$$I_N = I_{DM} - I_{CM} \quad (6.2)$$

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It is observed from Figure 6.10 that there are some frequencies in the transmission spectrum with minimum signal strength measured. This is because, in the frequency band of 1 to 30 MHz, other than PLC there are also other licensed services operating in this frequency band. In order to prevent PLC from causing unnecessary interference to them, the PLC modems are programmed to attenuate the signals at these frequencies through the use of notch filters.

Figures 6.11, 6.12 and 6.13 show the measured CM currents, E-fields and H-fields for the following three cases – (a) Ambient, (b) Original PLC modems and (c) Modified PLC modems with power transformer removed.

For the ambient case (a), the personal computers (PCs) and PLC modems were turned on. However, no communication was carried out between PLC Modem 1 and PLC Modem 2. For cases (b) and (c), batteries were used to power up the PLC modems. However, for case (c), the power transformer was removed from both the transmitting and receiving PLC modems. By removing the power transformer, the impedance between the power line and digital circuitries is increased.

Figure 6.11. Measured CM noise currents for the 3 cases.

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Figure 6.12. Measured E-fields for the 3 cases.

Figure 6.13. Measured H-fields for the 3 cases.

From the measured results, it can be observed that there is a relationship between the radiated emissions and the CM noise current in the 3-m power line test bed. It can also be concluded that the removal of the power transformer results in a decreased CM noise current in the test bed throughout the whole transmission band which may directly or indirectly lead to a decreased radiated emissions. However, the extent of the

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improvement varies with the frequency. The improvement can be as high as 25 dB reduction in the frequency band from 4 to 15 MHz. In the next frequency band of 15 to 18 MHz, there is only an observable 5-10 dB reduction. In the last frequency band of 18 to 21 MHz, there is not much noticeable reduction in conducted and radiated emissions. From the 3 plots, the average reduction in conducted and radiated emissions observed is 15 dB.

From the measurement results, it has shown that it is possible to reduce the radiated emission from communicating over the power line by removing the power transformer from the PLC modem's circuitry. However, it is not practical to remove the power transformer and resort to batteries as current battery capacity is only able to provide slightly less than 3 hours of operation time. This leads us to propose a new solution to the modem circuitry. Now we know that the power transformer is a possible coupling path for the CM noise current, we can increase the CM impedance of the power transformer by adding a CM choke in series with it. Figure 6.14 shows the proposed solution. This method of adding the CM choke is different from simply inserting the CM choke between the AC mains line (or modem) and the electrical appliances [56]. By inserting the CM choke between the AC mains line and the electrical appliances, the CM and DM impedances of the whole system will be greatly affected. However, the adding the CM choke in series with the power transformer does not affect the CM and DM impedances of the whole system much. This is due to the fact that from a circuit point of view, the power transformer is parallel with the isolation transformer. With the CM choke in series with the power transformer, the CM and DM impedances follow that of the isolation transformer. The net effect is almost equivalent to the removal of the power transformer.

Figure 6.14 illustrates the implementation of a CM choke in series with the power transformer in power line modem. By winding the two signal wires around a single ferrite core with equal number of turns a CM choke or inductor is produced. To allow the CM current propagating in each winding to create fluxes that are equal in magnitude and phase, the windings are secured tightly on the core. These magnetic fluxes within the

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Figure 6.14. Power line modem with CM choke in series with power transformer.

ferrite core will accumulate and strengthen in the ideal case. With the magnetic fluxes flowing and strengthening within the ferrite core, the CM choke will effectively function as an inductor against the CM current. This method provides a more effective noise suppression technique compared to using the ferrite inductor, which has only one winding around the core, on each conductor [59]. For the inductor, the core is often expected to support the entire current on each conductor without saturation. Thus, to avoid saturation effect, the inductor is often made with a core of a low effective permeability material such as gapped ferrites or powder cores. The CM choke, on the other hand, can use a high permeability material and obtain a very high inductance on a relatively small core.

Figure 6.15. Magnetic fluxes within (a) an ideal and (b) a practical CM choke.

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The DM current propagates in each winding of the CM choke with equal magnitude but opposite in phase. This creates fluxes within the core that are equal in magnitude but opposite in phase. These two fluxes will cancel each other resulting in a balanced state for the core. Ideally, the cancellation of fluxes will result in a zero inductance and this will have no effect on the DM current propagating towards the load. However, in practice, the CM choke will have some leakage fluxes between the two windings that result in an incomplete cancellation. This is due to the imperfect coupling of the windings around the core [60]. The creation of the leakage fluxes, which do not link with all the turns, causes the windings on both sides of the core to have some form of self-inductance. These leakage fluxes alternately store and discharge magnetic energy with each electrical cycle and thus effectively acts as an inductor in series with the mutually-coupled transformer windings. In addition, the windings also have some winding resistances, which would have some effect on the transmitted DM signals. Figure 6.15(a) and Figure 6.15(b) compare the working principles of an ideal CM choke and a practical one respectively.

6.4 EXPERIMENTAL VERIFICATION FOR ADDING A CM CHOKE IN SERIES WITH THE POWER TRANSFORMER

Figure 6.16. CM and DM impedances of a CM choke.

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The experimental setups for measuring the conducted and radiated emissions are the same as that shown in Figures 6.8 and 6.9. This time round, we are adding a CM choke in series with the power transformer, and we are also studying whether the presence of loads has any effect on the reduction of CM noise as well. The CM choke used is Multicomp's 10 mH CM choke with current rating of 0.1A, part number: SU9V-01100.

Figure 6.16 shows the measured CM and DM impedances of the CM choke. The communication bandwidth for the power line modem is from 4 to 21 MHz. The CM impedance is 20 k Ω at 4 MHz and it drops to 4 k Ω at 21 MHz. The DM impedance is lower compared to CM impedance. The DM impedance is 10 k Ω at 4 MHz and it drops to 600 Ω at 21 MHz.

The conducted and radiated emissions are measured without any load connected first. Then the conducted and radiated emissions are measured with Loads 1, 2 and 3 connected to some home appliances to simulate the real PLC environment. The results are then analyzed and discussed.

6.4.1 Measurements of conducted and radiated emissions with no loads connected at Loads 1, 2 and 3

Figure 6.17. Comparison of CM current of the modem without and with CM choke.

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Figure 6.17 shows the CM current for the original modem and the modem after adding CM choke in series with power transformer. The maximum CM current reduction occurs at 19 MHz. The measured value is only 15 dB after adding the choke, while it is 34 dB for the original modem. The average reduction of CM current between 10 and 21 MHz is around 15 dB. Before 10 MHz, the CM current for original modem is low, that explains why the CM current reduction is not significant.

A higher reduction of the CM current above 10 MHz has been achieved by adding the CM choke at the power transformer in the power line modem. This shows that the parasitic capacitance of the power transformer in the modem has contributed largely to the generation of CM noise, and this has been solved by the addition of CM choke to the power transformer.

The modem without CM choke generates a high amount of CM current due to the winding imperfection of the power transformer. This forms parasitic capacitance between the windings and results in a CM noise path between the power line and digital circuitries in the power line modem. Therefore, CM noise current flows through the parasitic capacitance to the power line. Isolation transformer also has a small amount of parasitic capacitance between its input and output. This contributes to the generation of CM noise from the modem to the power line.

Figure 6.18. Comparison of the DM current of the modem without and with CM choke.

Chapter 6 Investigation of Common-mode Coupling Path in Power Line Modems

Figure 6.18 shows the DM current for the modem without and with CM choke. The measurement results show the attenuating effects of the CM choke leakage inductance on the level of the DM signal transmitted into the impedance network. The DM current for the original modem at 11 MHz is 55 dB and it is reduced to 48 dB after adding the CM choke, hence a reduction of 7 dB.

Figure 6.19 shows the reduction in CM and DM currents for the modem without and with CM choke. The reduction for CM current is higher between 10 MHz and 21 MHz because the amount of CM current produced below 10 MHz is low, hence the reduction after adding the CM choke is not significant. The highest CM reduction is 20 dB at 19 MHz. The DM current is reduced at a relatively constant level of 10 dB. The DM current is reduced due to the CM choke's leakage fluxes between the windings that result in imperfect cancellation.

Figure 6.19. Reduction in CM and DM currents for modem with CM choke.

Figure 6.20. Comparison of radiated emissions of the modem without and with CM choke.

Figure 6.20 shows the comparison of the radiated emissions between the modem without and with CM choke. The reduction in radiated emission is consistent with the reduction in conducted emission (CM current). The highest reduction is as high as 18 dB at 18.5 MHz, where the original modem's radiated emission is 60 dB and it reduces to 42 dB after adding the CM choke. The average radiated emission reduction across the whole frequency range is 11 dB.

6.4.2 Measurements of conducted and radiated emissions with loads connected at Loads 1, 2 and 3

Power line network includes many electrical appliances and switches. These components will cause electrical unbalance on the power line network and increase radiated emissions. The electrical unbalance of a transmission line is caused by unbalance at the transmitter, receiver, and on the cable itself [61].

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PC (Load 1) and fan (Load 2) are put on the 3-m test bed. There is no load connected at Load 3. Measurements of CM, DM currents and radiated emission with the presence of these two loads were taken. This experiment was done to further study the effectiveness of putting the CM choke in series with the power transformer in the power line modem with the presence of loads commonly used at homes and offices.

Many devices are connected to the mains but not all of them are switched on. This is true in indoor environments such as homes and offices. Devices that are switched on have low impedance connection between Live-to-Earth and Neutral-to-Earth. Hence, balanced devices plugged to the mains and switched on do not cause any impedance imbalance. But, if these balanced devices are connected to the mains but not switched on, there will be a high impedance from Live-to-Earth, but low impedance from Neutral-to-Earth which will cause impedance imbalance.

Figure 6.21. Comparison of the CM current of the modem without and with CM choke with loads on test bed.

Figure 6.21 shows the CM current of the modem without and with CM choke with the presence of PC and fan as loads on the test bed. The maximum CM current reduction is

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seen at 17 MHz where the original modem CM current is at 29 dB and it is reduced to 14 dB after the addition of CM choke, hence a reduction of 15 dB is observed. The average CM current reduction across the whole frequency range is 10 dB.

Since the power transformer's primary winding is connected to the power line, the parasitic capacitance that is formed between the primary and secondary windings offers a low impedance path for the high frequency signal to propagate to the power line, which results in higher amount of CM current induced. At high frequencies, there is a low CM impedance path for the high frequency noise to bypass the transformer and induce CM current on the power line.

Comparison of DM Current (With Loads)

Figure 6.22. Comparison of the DM current of the modem without and with CM choke with loads on test bed.

Figure 6.22 shows the comparison of the DM current of the modem without and with CM choke with loads on test bed. Between 4 and 11 MHz, the DM current reduction is 9.5 dB,

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while between 11 and 21 MHz the reduction is lower at 6.5 dB. The average DM current reduction across the whole frequency range is 7.5 dB.

Figure 6.23 shows the reduction in CM and DM currents before and after adding CM choke on the power line modem. The reduction in DM current is higher compared to reduction in CM current between 4 and 11 MHz, because the CM current for the original modem between 4 and 11 MHz is low, hence CM current reduction in this frequency range is not significant. Between 11 and 18 MHz, the reduction in CM current is higher compared to DM current.

Figure 6.23. Reduction in CM and DM currents for modem with CM choke.

Figure 6.24. Comparison of the radiated emissions of the modem without and with CM choke with loads on test bed.

Radiated emission of a PLC system is directly related to electrical unbalance of the power line network. CM current is generated in the power line resulting in electromagnetic radiation. Figure 6.24 shows the radiated emissions for the modem without and with the addition of CM choke. The highest radiated emission reductions are at 12 dB at 11 MHz and 17.8 MHz. The average reduction in radiated emission across the whole frequency range is 8 dB.

Chapter 7:

CONCLUSIONS AND RECOMMENDATIONS FOR FUTURE RESEARCH

7.1 CONCLUSIONS

PLC Technology is a convenient and economical alternative for broadband access, multimedia communication and home automation. Current PLC technology has issues hindering its full swing commercialization. One of the main issues is the significant levels of radiated emissions from PLC systems which has the potential to interfere with radio receivers and other electronic devices. Many radiated emission field studies have shown that the radiation emissions from PLC systems are about 20 dB higher than the FCC limits which is the least stringent regulatory limit.

Such significant levels of radiated emissions from PLC systems is the primary factor which is inhibiting its full-scale commercialization. Hence the main motivation of the project was to understand the radiation mechanisms so as to implement new signal injection techniques and other EMI mitigation techniques to solve the EMI issue so that widespread use of PLC technology is possible.

In this thesis, we have proposed a complete current propagation model for a three-wire power line cable based on the Multi-conductor Transmission Line (MTL) theory. This complete model consists of a DM current propagation model and a CM current propagation model. The frequency of interest is from 1 to 30 MHz. The complete current propagation model can be used to characterize and understand the behavior of DM and CM in the PLC network. This model uses a three-wire power line cable as it is a more accurate representation of the actual power line network. The distributed per-unit-length DM and CM propagation parameters for the power line cable are derived based on the properties of the power line cable. Next, the chain parameter matrix for any length of the power line cable can be calculated. Once we know the DM and CM currents and voltages

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at one end of the power line, the levels of DM and CM voltages at any point of the power line cable for different loading conditions can be estimated with reasonable accuracy. From the modeling of the current distribution along the power line cable, we can find out the maximum levels of DM and CM currents flowing. With these values, we can estimate the worst case electric field strength generated by these currents. This complete model allows us to predict the level of EMI radiation based on the DM and CM currents flowing in the power line cable.

We have also proposed a new signal injection technique that utilizes the existing Earth wire to achieve the cancellation of electromagnetic fields so as to achieve reduction of the radiation. As compared to the alternative injection technique proposed by other authors and the normal injection technique, we can achieve significant reduction in the radiation for the balanced case and slight reduction in the radiation for the unbalanced case. The same conclusion can also be made for the total CM current in the LNE wires. This suggests a direct relationship between the total CM current in the LNE wires and the level of radiated emission. Although our new injection technique and the alternative injection technique can both achieve reduction in the level of radiation, our injection technique results in a lower level of CM in the Earth wire. With a lower CM current in the Earth wire, the resultant radiation emitted from the actual system may be lower. Also, our new injection technique proves to be a safer technique as we are using the Earth wire as the reference conductor. The alternative injection technique uses the Neutral wire as the reference conductor and may cause a voltage rise on the Earth wire. Although our new injection technique proves to be a good solution to mitigate the radiation level, this technique can only be used in countries which have the Live, Neutral and Earth wires running close and parallel to each other.

In order to explore other methods of mitigation of the EMI radiation level, the generation of parasitic capacitance between the primary and secondary windings of the power transformer and isolation transformer of the power line modem is also studied. From the measurement results, it has been shown that the large parasitic capacitance of the power transformer is the major cause of CM coupling path to the power line network. The

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smaller parasitic capacitance of the isolation transformer is another alternative CM coupling path, though it does not contribute as significantly as the power transformer. From the experiments conducted, it has been shown that it is possible to reduce the conducted and radiated emissions from communicating over the power line by removing the power transformer from the PLC modem's circuitry. An average reduction of 15 dB can be seen across the whole PLC bandwidth with a maximum of around 25 dB in the 4 to 15 MHz range and around 5 to 10 dB in the 15 to 18 MHz range. However, it is not practical to remove the power transformer and resort to batteries as the current battery capacity is only able to provide slightly less than 3 hours of operation time. This leads us to propose a new solution to the modem circuitry, that is, to implement a CM choke in series with the power transformer. After implementing the CM choke, an average CM current reduction of 14 dB in conducted emission and 11 dB in radiated emission are achieved without the presence of loads on the test bed. While, with the presence of PC and fan on the test bed, an average reduction of 10 dB in conducted emission and 8 dB in radiated emission are achieved. But, due to the leakage inductance of CM choke and stray capacitance between the CM choke wires connected with the Live and Neutral wires, DM current is also affected. Hence, in order to maintain the original level of DM current, more research needs to be done.

7.2 RECOMMENDATIONS FOR FUTURE RESEARCH

The followings are some recommendations and issues opened for future research and development in the PLC systems:

1. The scope of the proposed complete current propagation model is only restricted to a straight line configuration within the power line network of a residential house. In a high-rise building, there exist many branches from the cable risers going to different floors. At each floor, from the main distribution board, it is again branched and connected to several houses. The different types of cables and devices present make the topologies of the whole network very complex. A

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more comprehensive model is required to account for all the variants such as the rectangular bus bars, other cable structures and other ways of laying the cables.

2. Currently, for the modeling of the power line cables, we are only considering the skin effect for the whole wires only and not taking into consideration the proximity effect as we assume the wires are widely separated. This assumption is valid as the wires in house wirings are not tied together. More detailed modeling of the power cables, wires and stranded conductors could be done to incorporate skin effect and proximity effect for frequencies between 1 MHz and 30 MHz when calculating ac resistance of stranded conductors. We could also study the impact on the final results due to taking into consideration the skin effect and proximity effect for stranded conductors.
3. For the new injection method, we could only implement it to reduce the radiation level from indoor broadband PLC in countries where the three wires run together. This can only be done if we can find ways to design the modem such that in the event of a wiring fault, the fault current that is flowing in the Earth wire would not damage the PLC modem. So the design of a protection circuit against fault current flowing in the Earth wire can be carried out in order to implement this injection method.
4. For the new injection method, we are currently limiting our scope of investigation to a simple 3-m long power line cable. Further investigations could be carried out in a practical environment which involves the Earth wire as the ground connection. These investigations could be carried out with a coupler that is designed to inject the required signals into the Live and Earth wires and Neutral and Earth wires simultaneously. More work could be done to investigate the new injection method on different lengths of cables and also on a network of cables. Different configurations of networks could also be explored. The effects of reflections at line discontinuities (junctions) using the new injection method could also be studied in a more complex configuration. Besides practical

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measurements, a modeling of the proposed solution could also be done. Also, the experiments could be taken to a new level by conducting them in an apartment and even in a building with real power line networks.

5. Placing a CM choke in series with the power transformer provides an excellent CM current reduction, but the DM current reduction is also significant, resulting in a loss of data signal. One recommendation in implementing this solution is by adding a discrimination network between the Live and Neutral lines and a CM choke in the modem. The discrimination network would be able to separate the CM and DM signals. Hence, the CM signal will be forced to go through the CM choke to achieve maximum attenuation and the DM signal will go through another path, bypassing the CM choke, hence the data signal is maintained.

REFERENCES

- [1] Available: <http://www.internetworldstats.com/stats.htm>.
- [2] Available: <http://www.ida.gov.sg>.
- [3] Available: <http://www.singstat.gov.sg> (for broadband).
- [4] J. Newbury, Broadband Power Line Communications.
Available: http://www.metering.com/archive/033/32_1.htm.
- [5] D. Hansen, EMC: The Impact of Power Line Communications, Part 1.
Available: <http://www.ce-mag.com/archive/03/ARG/hansen1.html>
- [6] J. J. Fahie, "Edward Davy," The Electrician, pp. 181-227, 7 Jul 1883.
- [7] J. Routin and C. E. L. Brown, "Power line signalling electricity meters," UK Patent Office, British Patent No. 24833, 1897.
- [8] K. M. Dostert, "Error Control Coding for Mainsborne Communications," IEEE International Symposium on Information Theory and Its Applications (ISITA'96), pp. 135-138, Sep 1996.
- [9] N. Pavlidou, A. J. Han Vick, J. Yazdani and B. Honary, "Power line communications: state of the art and future trends," IEEE Communications Magazine, vol. 41, pp. 34-40, 2003.
- [10] Available: <http://www.homeplug.org>.
- [11] Joint Press Release by Singapore Power and Singapore Polytechnic (1 Nov 2000), Singapore Power First to put PLC on Trial in Asia at Singapore Polytechnic.
Available:
http://www.singaporepower.com.sg/publish/PR/Files/Approve/PR29_12179.html.
- [12] Joint Press Release by SP Telecom, Pacific Internet and Little Green Apples (11 Apr 2002), Singapore Power Kicks Off Market Trial. Available:
http://www.singaporepower.com.sg/publish/PR/Files/Approve/PR11_4239.html.
- [13] S. Roper, "Some issues in the near and far field radiated emission benchmarking of PLT systems," International Symposium on Power-Line Communications and Its Applications (ISPLC), pp. 185-190, 2001.

- [14] J. M. Silva and B. Whitney, "Evaluation of the potential for power line carrier (PLC) to interfere with the use of nationwide differential GPS network," *IEEE Transactions on Power Delivery*, vol. 17, no. 2, pp. 348-352, Apr 2002.
- [15] C. Muto, N. Mori, and T. Kondoh, "On radio interference assessments of access PLC system," *International Symposium on Power-Line Communications and Its Applications (ISPLC)*, pp. 67-72, 2003.
- [16] K. Asano, M. Shin, Y. Matsuura, K. Maegawa, S. Yamamoto, and N. Nakayama, "Experimental study of the electromagnetic radiation caused by PLC with frequency band signals," *International Symposium on Power-Line Communications and Its Applications (ISPLC)*, pp. 143-148, 2004.
- [17] E. Marthe, F. Rachidi, M. Ianoz, and P. Zwiacki, "Indoor radiated emission associated with power line communication systems," *IEEE International Symposium on Electromagnetic Compatibility*, vol. 1, pp. 517-520, 2001.
- [18] C. Hensen and S. Schwarze, "Electromagnetic compatibility for power-line communications on the medium voltage power grid," *International Symposium on Power-Line Communications and Its Applications (ISPLC)*, pp. 155-160, 2001.
- [19] R. Vick, "Radiated emission caused by in-house PLC-systems," *International Symposium on Power-Line Communications and Its Applications (ISPLC)*, pp. 197-202, 2001.
- [20] Australian Communications Authority, Spectrum Planning Team, Radio Frequency Planning Group, Broadband Powerline Communication Systems, A Background Brief, Document SP 11/03, Sep 2003.
- [21] B. Adebisi, and B. Honary, "Comparisons of indoor PLC emissions measurement results and regulations standards," *International Symposium on Power-Line Communications and Its Applications (ISPLC)*, pp. 319-324, Mar 2006.
- [22] Radio Communications Agency, MPT 1570 Radiation Limits and Measurement Specification, Jan 2003.
- [23] "FCC Rules for Radio Devices", Code of Federal Regulations, 47 CFR Part 15, 7 Aug 1998.
- [24] H. Hrasnica, A. Haidine, and R. Lehnert, *Broadband Powerline Communications: Network Design*, John Wiley & Sons, 2004.

- [25] M. D'Amore and M. S. Sarto, "Electromagnetic field radiated from broadband signal transmission on power line carrier channels," *IEEE Transactions on Power Delivery*, vol. 12, no. 2, pp. 624-631, Apr 1997.
- [26] K. Dostert, "EMC aspects of high speed power line communications," *International Wroclaw Symposium and Exhibition on Electromagnetic Compatibility*, pp. 98-102, 2000.
- [27] M. Ianoz, *Electromagnetic Effect due to PLC and Work Progress in Different Standardization Bodies*, Final Report, EMC Group – Swiss Federal Institute of Technology, Nov 2002.
- [28] R. Vick, "Radiated emission of domestic main wiring caused by power-line communications systems," *International Wroclaw Symposium and Exhibition on Electromagnetic Compatibility*, pp. 111-115, 2000.
- [29] I. P. Macfarlane, "A probe for measurement of electrical unbalance of networks and devices," *IEEE Transactions on Electromagnetic Compatibility*, vol. 41, no. 1, pp. 3-14, Feb 1999.
- [30] C. Hensen, "Measuring the longitudinal conversion loss value of in-house low voltage power lines," *International Symposium on Power-Line Communications and Its Applications (ISPLC)*, pp. 161-165, 2004.
- [31] H. Furukawa and S. Tsuzuki, "A low interference PLC modem of selecting subcarriers based on LCTL measurement," *International Symposium on Power-Line Communications and Its Applications (ISPLC)*, pp. 149-154, 2004.
- [32] M. Zimmermann and K. Dostert, "A multipath model for the powerline channel," *IEEE Transactions on Communications*, vol. 50, no. 4, pp. 553-559, Apr 2002.
- [33] L. T. Tang, P. L. So, E. Gunawan, Y. L. Guan, S. Chen, and T. T. Lie, "Characterization and modeling of in-building power lines for high-speed data transmission," *IEEE Transactions on Power Delivery*, vol. 18, no. 1, pp. 69-77, Jan 2003.
- [34] H. Philipps, "Modeling of powerline communication channels," *International Symposium on Power-Line Communications and Its Applications (ISPLC)*, pp. 14-21, 1999.

- [35] T. Bostoen and O. V. de Wiel, "Modeling the low voltage power distribution network in the frequency band from 0.5MHz to 30MHz for broadband powerline communication (PLC)," Zurich Seminar on Broadband Communications, pp. 171-177, Feb 2000.
- [36] T. C. Banwell and S. Galli, "A new approach to the modeling of the transfer function of the power line channel," International Symposium on Power-Line Communications and Its Applications (ISPLC), pp. 319-324, 2001.
- [37] T. Esmailian, F. R. Kschischang, and P. G. Gulak, "Characteristics of in-building power lines at high frequencies and their channel capacity," International Symposium on Power-Line Communications and Its Applications (ISPLC), pp. 52-59, 2000.
- [38] S. Tsuzuki, T. Takamatsu, H. Nishio, and Y. Yamada, "An estimation method of the transfer function of indoor power-line channels for Japanese house," International Symposium on Power-Line Communications and Its Applications (ISPLC), pp. 55-59, 2002.
- [39] H. Meng, S. Chen, Y. L. Guan, C. L. Law, P. L. So, E. Gunawan, and T. T. Lie, "A transmission line model for high-frequency power line communication channel," PowerCon, pp. 1290-1295, Oct 2002.
- [40] D. Anastasiadou and T. Antonakopoulos, "An experimental setup for characterizing the residential power grid variable behaviour," International Symposium on Power-Line Communications and Its Applications (ISPLC), pp. 65-70, 2002.
- [41] H. Meng, S. Chen, Y. L. Guan, C. L. Law, P. L. So, E. Gunawan and T. T. Lie, "Modeling of transfer characteristics for the broadband power line communication channel," IEEE Transactions on Power Delivery, vol. 19, no. 3, pp. 1057-1064, Jul 2004.
- [42] T. Sartenar and P. Delogne, "Power cables modeling for broadband communications," International Symposium on Power-Line Communications and Its Applications (ISPLC), pp. 331-338, 2001.
- [43] K. Dostert, Powerline communications, Prentice Hall, 2001.

- [44] R. P. Rickard, Coupling communications signals to a power line, Patent US6037678, 14 March 2003.
- [45] M. Rubinstein, J.L. Bermudez, A. Vukicevic, F. Rachidi, M. Schneider, E. Marthe "Discussion on the assessment and mitigation of radiation from PLC networks," International Symposium on Power-Line Communications and Its Applications (ISPLC), pp. 215-218, Apr 2005
- [46] N. Korovkin, E. Marthe, F. Rachidi, E. Seilna, "Mitigation of electromagnetic field radiated by PLC systems in indoor environment," International Journal of Communication Systems, Vol. 16, pp.417-426, May 2003.
- [47] C. R. Paul, Analysis of Multiconductor Transmission Lines, John Wiley & Sons, 1994.
- [48] D. K. Cheng, Fundamental of Engineering Electromagnetics, Addison-Wesley Publishing, 1992.
- [49] R. M. van Maurick, "Potential common mode currents on the ISDN S and T-interface caused by cable unbalance," IEEE International Conference on Electromagnetic Compatibility, no. 362, pp. 202-206, 1992.
- [50] S. Kim and D. P. Neikirk, "Compact equivalent circuit model for the skin effect," IEEE-MTT-S International Microwave Symposium., Editor: R. G. Ranson, vol. 3, pp. 1815-1818, 1996.
- [51] K. Y. See, P. L. So, A. Kamarul, and E. Gunawan, "Radio-frequency common-mode noise propagation model for power-line cable," IEEE Transactions on Power Delivery, vol. 20, no. 4, pp. 2443–2449, Oct 2005.
- [52] K. Y. See, A. Kamarul, and P. L. So, "Reduction of CM noise emissions from PLC modem using optically coupled signaling," IEEE Transactions on Electromagnetic Compatibility, vol. 48, no. 4, pp. 648–653, Nov 2006.
- [53] H. L. Chen, Y. Kenji and M. Yuji, "Introduction of an optical isolator design for narrow band power line modems and feasibility study for optical isolator for broad band power line modems," International Symposium on Power-Line Communications and Its Applications (ISPLC), pp. 73-78, 2003.
- [54] Available: http://en.wikipedia.org/wiki/Domestic_AC_power_plugs_and_sockets.
- [55] Available: <http://users.pandora.be/worldstandards/electricity.htm>.

- [56] D. Hirata, N. Kuwabara, Y. Akiyama, and H. Yamane, "Influence of appliance state on transmission characteristics of indoor AC mains lines in frequency range used power line communication", IEEE International Symposium on Electromagnetic Compatibility, Vol. 1-3, pp. 715-720, 2005.
- [57] National Centre for Construction and Education and Research, Advanced Electrical Topics, Prentice Hall, pp. 55-56.
- [58] M. Ishihara, D. Umehara, and Y. Morihiro, "The correlations between radiated emissions and power line network components on indoor power line communications," International Symposium on Power-Line Communications and Its Applications (ISPLC), pp. 314-318, 2006.
- [59] K. Naishadham, "A rigorous experimental characterization of ferrite inductors for RF noise suppression," IEEE Radio and Wireless Conference, pp. 271-274, 1 Aug 1999.
- [60] M. J. Nave, "On modeling the common mode inductor," International Symposium on Electromagnetic Compatibility, pp. 452-457, 12 Aug 1991.
- [61] J. A. Malack, "Statistical correlation between conducted voltages on the powerline and those measured with a line-impedance stabilization network," IEEE Transactions on Electromagnetic Compatibility, pp. 346-349, May 1978.

APPENDICES

This appendix chapter presents the two-current-probe methodology which is used widely throughout the course of this research and also the commonly used equipment that is used in conducting the experiments.

A.1 Two-Current-Probe Methodology

Figure A.1. Two-current-probe measurement methodology [51], [52].

Figure A.1 illustrates the basic setup of the two-current-probes approach to measure any unknown impedance Z_X . The measurement system consists of an injecting current probe, a monitoring current probe and a network analyzer. A high-frequency coupling circuit which is made up of two current probes and a decoupling capacitor C is required to avoid direct connection between the network analyzer and the high voltage power mains. A pair of short wires with length d from the reference plane b' to the reference plane b connects the coupler to Z_X . Port 1 of the network analyzer induces a continuous wave (CW) signal in the closed loop through the injecting current probe. Port 2 of the network analyzer measures the generated current propagating in the closed loop through the monitoring current probe.

Figure A.2. Equivalent circuit of the injecting current probe [51], [52].

Figure A.2 shows how the injecting current probe can be modeled as an equivalent transformer circuit. In this equivalent circuit, V_{p1} is the source voltage and Z_p is the source impedance of the injecting signal source from port 1 of the network analyzer. When a signal is injected, the injecting probe induces a voltage V_w , this results in a current I_w circulating in the closed loop. L_p , L_w , and M are the primary self-inductance of the probe, the self-inductance of the wire in the probe volume, and the mutual inductance between the probe and the wire respectively. V_{p1} and V_w can be expressed as follows [51], [52]:

$$V_{p1} = (Z_p + j\omega L_p)I_p - j\omega M I_w \quad (\text{A.1})$$

$$V_w = -j\omega M I_p + j\omega L_w I_w \quad (\text{A.2})$$

By combining (A.1) and (A.2) and eliminating I_p , the induced voltage V_w can be expressed as [51], [52]:

$$V_w = Z_{M1} I_w - V_{M1} \quad (\text{A.3})$$

where

$$Z_{M1} = j\omega L_w + \left(\frac{(\omega M)^2}{Z_p + j\omega L_p} \right) \quad (\text{A.4})$$

$$V_{M1} = V_{p1} \cdot \left(\frac{j\omega M}{Z_p + j\omega L_p} \right) \quad (\text{A.5})$$

Equations (A.4) and (A.5) suggest that the injecting current probe can be represented at the reference plane a by a Thevenin equivalent circuit as shown in Figure A.3. Here, due to the existence of the monitoring current probe, Z_{M2} is the reflected impedance seen by the loop.

Figure A.3. Equivalent circuit for the two-current-probes setup [51], [52].

From (A.5), the ratio V_{M1}/V_{p1} , which is dependant on the properties of the injecting probe and its operating frequency ω , is given by [51], [52]

$$K_R = \frac{V_{M1}}{V_{p1}} = \frac{j\omega M}{Z_p + j\omega L_p} \quad (\text{A.6})$$

The transmission line effect can be ignored here as the typical wire length d between the reference planes b' and b is much shorter than the wavelength of the maximum frequency of interest (10 m at 30 MHz). With Z_{ip} being the impedance seen by the unknown impedance Z_X at the reference plane b , V_{M1} can be redefined as [51], [52]

$$V_{M1} = (Z_{ip} + Z_X)I_w \quad (\text{A.7})$$

where $Z_{ip} = Z_{M1} + Z_{M2} + Z_C$. Substituting V_{M1} from (A.7) into (A.6), Z_X can be evaluated by

$$Z_X = (K_R Z_{T2}) \cdot \left(\frac{V_{p1}}{V_{p2}} \right) - Z_{ip} \quad (\text{A.8})$$

where $Z_{T2} = [V_{p2} / I_w]$ and V_{p2} is the transfer impedance of the monitoring current probe and the voltage measured by the monitoring probe at Port 2 respectively. The product $K_R Z_{T2}$ is a frequency dependent coefficient that is independent of the unknown load Z_X . Using the network analyzer, the ratio of V_{p1} / V_{p2} can be obtained through the S-parameters measurement as follows [51], [52]:

$$\frac{V_{p1}}{V_{p2}} = \frac{S_{11} + 1}{S_{21}} \quad (\text{A.9})$$

In a pre-measurement calibration process, the following steps can be used to obtain the product $K_R Z_{T2}$: Firstly, Z_X is removed and Z_{ip} is measured using an impedance analyzer. Then, Z_X is replaced with a known precision standard resistor R_{std} and V_{p2} / V_{p1} is measured again using the S-parameter measurement function of the network analyzer. The coefficient $K_R Z_{T2}$ is finally obtained by [51], [52]

$$K_R Z_{T2} = \frac{Z_{ip} + Z_{std}}{\left(\frac{V_{p1}}{V_{p2}} \right) \big|_{Z_X = Z_{std}}} \quad (\text{A.10})$$

When $K_R Z_{T2}$ and Z_{ip} are determined, the two-current-probe setup is ready to be used to determine any unknown impedance (Z_X). Z_X can be found based on (A.8) after measuring V_{p1} and V_{p2} using the network analyzer.

A.2 Physical Characteristics and Parameters of Power Line Cables

Figure A.4. Cross section view of power line cable used.

Figure A.4 shows the cross section view of the power line cable used in this thesis. The power line cable that is used throughout the thesis is a single-core PVC (Polyvinyl Chloride) insulated, non-sheathed cable with a voltage rating of 450/750 V. Its specifications match those of cables commonly used in the wiring of homes. The power line cable is 3-m in length, and made up of 3 wires, namely Live, Neutral and Earth. Each wire is insulated with PVC and is made up of 7 stranded copper conductors. The wire has a cross-sectional area of 2.5 mm^2 , and the diameter of a single copper strand is 0.67 mm. The radius of a single strand of copper conductors r_{cond} works out to be 0.335 mm and the radius of a single wire works out to be 1.005 mm.

Table A.1. Electrical properties of power line cable.

Material	Relative Permittivity, ϵ_r	Relative Permeability, μ_r	Conductivity, σ (S/m)
Air	1	1	$0.3 \text{ to } 0.8 \times 10^{-14}$
PVC	4	1	$\sim 1 \times 10^{-6}$
Copper	-	1	5.813×10^7

Table A.1 shows the electrical properties of the power line cable. Using the above information, the per-unit-length resistance, inductance, capacitance and conductance matrices in Chapter 4, Section 2 can be derived.

Figure A.5. Per-unit-length resistance values.

Figure A.6. Per-unit-length inductance values.

Figure A.7. Per-unit-length capacitance values.

Figure A.8. Per-unit-length conductance values.

Figure A.9. Attenuation constant.

Figure A.10. Phase constant.

Figure A.5 shows the frequency dependence of per-unit-length resistance values using (4.44) and (4.60). R_w is the simpler approximation which is based on the skin effect on the solid conductor while R_{w-corr} takes into account the gaps at the circumference of the conductor strands.

Figure A.6 shows the frequency dependence of per-unit-length inductance values using (4.65), (4.66), (4.68) and (4.69). L_{ii} and L_{ij} are derived based on the simpler approximation which is based on the skin effect on the solid conductor while $L_{ii-corr}$ and $L_{ij-corr}$ takes into account the gaps at the circumference of the conductor strands.

Figures A.7 and A.8 show the per-unit-length capacitance and conductance values respectively. It can be seen that these two values are independent of frequency.

The CM and DM attenuation constants and phase constants are shown in Figures A.9 and A.10 respectively.

A.3 Line Impedance Stabilization Network (LISN)

Figure A.11. Line impedance stabilization network.

Most standards require LISN to couple the radio frequency from the cable and provide repeatable impedance at the measured frequencies when measuring EMC emissions conducted from the equipment under test. A LISN has several functions. It can be used to isolate the measurement from external mains interference. Besides that, it can provide a known fixed impedance at radio frequency to the equipment under test. Finally, it can facilitate the coupling of radio frequency from the mains to an output suitable for connection to spectrum analyzer [61]. LISN is usually placed between the power mains and equipment under test to stabilize the line impedance, provide a 50-ohm RF connection and eliminate unwanted RF signals from the mains.

A.4 Schaffner Chase EMC CSP-8425-1 RF Clamp-On Current Probe

Figure A.12. Current probe.

Current probes are used to measure electrical current in the conductor without making physical contact with it. Measurements can be made on single or multi-conductor cables. Current probes measure the flux field generated by the movement of electrons through a conductor. Within the range specifications of the current probe, the flux field surrounding a conductor is converted to a linear voltage output that can be displayed and analyzed on the spectrum analyzer.

A.5 Agilent 4395A Network/ Spectrum/ Impedance Analyzer

Figure A.13. AGILENT 4395A Network/ Spectrum/ Impedance Analyzer.

The Agilent 4395A Spectrum Analyzer operates from 10 Hz to 500 MHz. A fully-synthesized local oscillator allows stable and accurate frequency analysis. It is a frequency-selective voltmeter calibrated to display the measured voltage results. It has the capability to present data in the frequency domain, those which are not recognizable in the time domain. This project covers the frequency range of 4 to 21 MHz which is the communication bandwidth for the power line modem.

A.6 EMCO 3301B Active Monopole

Figure A.14. EMCO 3301B Active Monopole.

The EMCO 3301B Active Monopole is a broadband, high sensitivity electric-field receiving antenna. It is designed to provide reception of electric field in the signal band without tuning from 30 Hz to 50 MHz. There are three main components: sensing rod, ground plane and broadband high input impedance pre-amplifier. The rod and ground plane works together as electrically short antenna over the ground plane. The pre-amplifier provides impedance transformation from the extremely high impedance at the base of the rod to 50 ohms that is required by most receiving systems, and provides power gain to allow the sensing of very low level signals.

A.7 HL100E HomePlug Ethernet Adaptor

Figure A.15. Power line modem.

Power line modem converts communication signals from digital to analog and transmit these signals through the power lines, by superimposing these modulated analog signals onto the 50 or 60 Hz alternating current, while the receiving modem converts these analog signals back to digital ones.

HL100E HomePlug Ethernet Adaptor manufactured by Aztech uses the 10Base-T Ethernet under the IEEE 802.3 standard with transfer rate of 14 Mbps with OFDM, DQPSK, DBPSK modulation techniques. The frequency band for communication is 4.3 MHz to 20.9 MHz. It is able to support communication up to a range of 460 square meters.

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AUTHOR'S PUBLICATIONS

Portions of the research work reported in this thesis have appeared in the following international publications:

Journal Publications:

1. **T. S. Pang**, P. L. So, K. Y. See, and A. Kamarul, "Modeling and analysis of common-mode current propagation in broadband power-line communication networks," IEEE Transactions on Power Delivery, vol. 23, no. 1, pp. 171-179, Jan 2008.
2. L. B. Wang, P. L. So, K. Y. See, M. Oswal, and **T. S. Pang**, "Investigation of radiated emissions in power line communications networks," International Journal of Emerging Electric Power Systems, vol. 9, iss. 3, Art. 4, pp. 1-14, Apr 2008.
3. **T. S. Pang**, P. L. So, and K. Y. See, "Feasibility study of a new injection method for EMI reduction in indoor broadband PLC networks," IEEE Transactions on Power Delivery, vol. 25, no. 4, pp. 2392-2398, Oct 2010.

Journal Papers submitted / under review:

1. **T. S. Pang**, P. L. So, and K. Y. See, "Modeling and analysis of different-mode and common-mode current propagation in broadband PLC networks," submitted to IEEE Transactions on Electromagnetic Compatibility, 2010.

Conference Papers Publications:

1. **T. S. Pang**, P. L. So, K. Y. See, and A. Kamarul, "Common-mode current propagation in power line communication networks using multi-conductor transmission line theory," IEEE International Symposium on Power Line Communications and its Applications (ISPLC 2007), Pisa, Italy, pp. 517-522, Mar 2007.
2. L. B. Wang, K. Y. See, P. L. So, **T. S. Pang**, and M. Oswal, "DC power supply design for broadband PLC modem with low common-mode EMI coupling," International Symposium on Integrated Circuits (ISIC 2007), Singapore, Sep 2007.
3. L. B. Wang, P. L. So, K. Y. See, M. Oswal, and **T. S. Pang**, "Investigation of radiated emissions in power line communication network," 8th International Power Engineering Conference (IPEC 2007), Singapore, Dec 2007.
4. **T. S. Pang**, P. L. So, and K. Y. See, "Modeling of radio frequency electromagnetic disturbances in power line communication networks," Asia-Pacific Symposium on Electromagnetic Compatibility and 19th International Zurich Symposium on Electromagnetic Compatibility, Singapore, May 2008.