Modeling and analysis of electrical power lines for broadband power line communications

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**MODELING AND ANALYSIS OF ELECTRICAL POWER LINES
FOR BROADBAND POWER LINE COMMUNICATIONS**

Meng Han

School of Electrical & Electronic Engineering

**A thesis submitted to the Nanyang Technological University
in fulfillment of the requirement for the degree of
Doctor of Philosophy**

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Statement of Originality

I hereby certify that the work embodied in this thesis is the result of original research and has not been submitted for a higher degree to any other University or Institution.

17 May 2005
Date

Meng Han

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SUMMARY

As demand for broadband access and home networking increase, new technologies are being created to tap into this growing broadband communication market. Among them, the power line communication (PLC) technology has recently attracted many attentions with its convenience of wide availability and low cost advantage. However, like all other technologies, the power line communication also faces its own set of obstacles and technical challenges. The communication medium of this technology, the power lines, is designed to transmit high electrical power at low frequency, and it appears as a harsh environment for the low-power high-frequency communication signals. With regard to communication purposes, the channel parameters, namely the transfer function and the input impedance are found to be highly variable, and the system is also subjected to much electrical noise. Hence, in order to design a well-adapted communication system over the power lines, a good understanding of these channel parameters is necessary. This research seeks to expand the knowledge in the characteristics of power lines and to develop a channel model that can be used for designing PLC scheme. The model incorporates all the three important channel parameters and takes into consideration of power networks with different topologies and under different loading conditions.

The modeling of the channel transfer function and the network input impedance is developed for both tree and ring topology networks. The proposed model is based on microwave theory, whereby the power line is analyzed as a multi-conductor transmission line. Based on this theory, the two important line parameters, namely the characteristic impedance and the propagation constant, are determined. These two line parameters serve as the basic parameters in the scattering parameter matrices technique for deriving the transfer function and the input impedance of the power network. The accuracy of the proposed model has been verified by several measurements using actual power networks. In addition, the sensitivity of the model towards any deviation in the network configuration is also studied.

The model also considers the effect of power line noise in terms of background and impulsive noises. The background noise is modeled based on statistical analysis of the

results from two long-term measurements conducted at two different sites - a laboratory and a residential apartment. On the other hand, the impulsive noise is modeled based on direct measurements from its dominant sources - various electrical household appliances. In addition, using the proposed channel transfer function model, the amount of background and impulsive noises that reaches a receiver is determined. This combined noise at the receiver is an essential factor for the design of PLC system. This proposed noise model at the receiver has also been verified by measurements.

Finally, to show the versatility of the proposed channel and noise models, several applications of the model in designing a PLC system are discussed. Specifically, the channel transfer function model is first used to determine the parameters of the modulation scheme. The noise model is later applied to predict the system performance. Combining the channel characteristics revealed by the transfer function model and the noise model, the system is optimized to enhance its performance. It is shown that multi-carrier scheme with bit-loading adaptation is able to achieve the necessary BER (bit error rate) with lower SNR (signal to noise ratio).

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LIST OF ABBREVIATIONS

(Alphabet sequence)

AMR:	Automatic Meter Reading
AWGN:	Additive White Gaussian Noise
BER:	Bit Error Rate
BPSK:	Binary Phase Shift Keying
CDF:	Cumulated Distribution Function
CEA:	Consumer Electronic Association
CENELEC:	European Committee for Electrotechnical Standardization
CP:	Cyclic Prefix
CSMA/CA:	Carrier Sensing Multiple Access with Collision Avoidance
CToP:	Carrier Transmission over Powerline
CTP:	Centralized Token Passing
DB:	Distribution Board
DBPSK:	Differential Binary Phase Shift Keying
DMT:	Discrete Multi-Tone
DQPSK:	Differential Quadrature Phase Shift Keying
DSL:	Digital Subscriber Line
DSMA:	Datagram Sensing Multiple Access
DSSS:	Direct Sequence Spread Spectrum
DS-CDMA:	Direct Sequence - Code Division Multiple Access
EM:	Electromagnetic
EMC:	Electromagnetic Compatibility
EMI:	Electromagnetic Interference
ETSI:	European Telecommunications Standards Institute
FCC:	Federal Communication Commission
FEM:	Finite Element Method
FFT:	Fast Fourier Transform
FSK:	Frequency Shift Keying
GI:	Guard Interval
GPIB:	General Purpose Interface Board

List of Abbreviations

HV:	High Voltage
IAT:	Inter Arrival Time
IDA:	Infocomm Development Authority
IEC:	International Electrotechnical Commission
ISDN:	Integrated Service Digital Networks
ISI:	Inter-Symbol Interference
LCL:	Longitudinal Conversion Loss
LCTL:	Longitudinal Conversion Transfer Loss
LMS:	Least Mean Square
LV:	Low Voltage
MAC:	Medium Access Control
MQAM:	M-Ary Quadrature Amplitude Modulation
MTL:	Multi-conductor Transmission Line
MV:	Middle Voltage
OC:	Open Circuit
OFDM:	Orthogonal Frequency Division Multiplexing
PDF:	Probability Density Function
PLC:	Power Line Communication
PSD:	Power Spectral Density
PVC:	Polyvinyl Chloride
QAM:	Quadrature Amplitude Modulation
QoS:	Quality of Service
QPSK:	Quarter Phase Shift Keying
RBW:	Resolution Bandwidth
RCS:	Ripple Carrier Signaling
SIG:	Special Interest Group
SNR:	Signal to Noise Ratio
SW:	Short Wave
S-matrix:	Scattering matrix
S-parameters:	Scattering parameters
TEM:	Transverse Electromagnetic
T-matrix:	Cain Scattering matrix
UTP:	Unshielded Twisted Pair
VoIP:	Voice over Internet Protocol

Chapter 1

INTRODUCTION

1.1 MOTIVATION

In just a few short years, the Internet has become an indispensable fact of life. The whole family is going online to download information, correspond with friends, get the latest news, purchase the amazing variety of goods and services, play interactive games, listen to music and manage their investment portfolios. At the same time, more and more families include the wage earners rely on the Internet for a home-based business, work in home one or more days each week, or travel on business and access to office resources and email. Students of all ages are also talking about Web-based courses offered through employers, local and remote universities, or online training companies.

With the increase of Internet usage, consumers demand nothing less than broadband access. Home broadband access and home networking go hand in hand because users can maximize their single high-speed connection and share it with all the connected devices and PCs. This enables access to bandwidth-intensive applications like multimedia streaming, voice over IP (VoIP), and multiple-player gaming. In fact, according to Parks Associates, of the 30 million U.S. households predicted to have broadband Internet connections by 2004, 17 million of them are projected to have home networks [1]. On the other hand, home access and networking depend not only on the penetration of broadband and multiple PCs but also the pervasiveness of devices that drive consumer demand for networking technology. As more and more applications penetrate the market, consumers are interested in a way to simplify and coordinate the abundance of products.

At the convergence of these trends, there is a clear demand within the technology industry for a home access and networking solution. Currently, there are several ways to do this, using either wired or wireless methods. The conventional wired broadband systems include Digital Subscriber Line (DSL), cable, and fiber optics. Recently, a new candidate has joined this family --- Power Line Communication (PLC).

1.1.1 Development of PLC

PLC is to use the existing power cable infrastructure to serve the communication purpose. Actually, power supply networks began to be used additionally for data transmission soon after the full coverage of electrification. As shown in Figure 1.1, a typical power network structure consists of three distinct levels: the High Voltage (HV) level (110-380kV), the Medium Voltage (MV) level (10-30kV), and the Low Voltage (LV) level (0.4kV), each adapted to the bridging of certain distances. The earliest PLC implementations include the Carrier Transmission over Powerline (CToP) on HV networks in 1920s and the Ripple Carrier Signaling (RCS) on MV and LV networks in 1930s. However, after that, this technology has not been carried forward very much. Until recently, the only worthwhile application had been seen in niche areas like building automation, lighting control, and remote meter reading [2-4]. All these applications belong to narrow band PLC since their signaling concentrates on the low frequency band (<500kHz), which technically limits the data transfer rate in the range of kbps. Extremely restrictive regulations with regard to the usable frequency ranges and admissible transmission levels have discouraged PLC developments. There has not been massive economic interest to drive the formation of a PLC lobby like those in other technological fields, which would be able to exert influence on standards and regulations.

Figure 1.1. Structure of the Typical Electric Energy Supply System

This situation changed radically with the deregulation of the telecommunications and energy markets. One immediate consequence is the demand for alternative fast data

connections on the local loop level. PLC could offer ideal solutions, mainly because the power supply infrastructure is denser than any other communication network. The full infrastructure from the provider to the home wall plug is there, ready for use without any additional installation cost. This means, for example, that fast Internet access from the wall plug will not remain just a vision but will gradually become reality. Development will advance at a pace because future users will demand to be provided with comprehensive information in easily understandable form. In 1995, NOR.WEB [5] demonstrated that data communications are possible in the high frequency region ($>1\text{MHz}$), which showed the technical feasibility of the PLC systems over the high frequency band. This system is referred as broadband PLC operating in the MHz frequency band and transferring data in Mbps range.

Although the narrow band PLC has been developed for all voltage levels in the power network, broadband PLC researchers tend to put their interests into the LV network. Different voltage levels are interconnected by transformers, designed in such a way that the energy loss is as low as possible at the power frequency (50 Hz). At the high frequencies typically used for data communication, the transformers are natural obstacles, which cause a perfect separation between multiple voltage levels. Additionally, in different levels, different power cables are used, which appear to have different channel characteristics for high frequency communication. As a result, it won't be easy to develop a PLC system which covers all the three voltage levels of power network. Presently, the development of the broadband PLC is concentrated mainly in the LV network due to its large coverage area and close connectivity to users.

1.1.2 Overview of Current PLC Industry

From 1995 to present, the PLC has experienced fairly fast development but until now the technology is still not matured. In different continents and counties, many organizations and companies in the PLC area are addressing different aspects of the challenge.

1.1.2.1 PLC Organizations and Standards

Currently, there are four organizations working towards a power line standard, each focusing on different types of applications [6]. EchoNet is an Asian organization that has

been chartered to provide a cost-effective power line solution to address the control and automation market. PLC Forum is responsible for making recommendations on PLC technology to the two European standards organizations, European Telecommunications Standards Institute (ETSI) and European Committee for Electrotechnical Standardization (CENELEC). This group has the strongest root in the “last mile” access industry but it is just a beginner in home networking. Consumer Electronic Association (CEA) R7.3 Working Group is primarily made up of consumer electronic companies as well as various power line proponents. The goal of the group is to endorse a single power line technology as the CEA-approved standard for communications between consumer appliances. Since there is a tremendous focus on consumer electronic appliances, the group’s requirements tend to focus on scalability across many devices, overall affordability and providing guaranteed multimedia streams in the form of QoS (Quality of Service). Finally, the HomePlug Special Interest Group (SIG) is focusing on providing an “Ethernet-class” solution for home networking, with the emphasis on speed and connectivity. It is noteworthy that in June of 2001, the SIG released their home networking specification (HomePlug spec 1.0) which is an Orthogonal Frequency Division Multiplexing (OFDM) implementation targeting at 10+ Mbps of raw data speed with payload throughput ranging from 500 Kbps to 5 Mbps depending on the number of active nodes and line conditions. Basically, HomePlug does not address the access networks and the consumer electronic applicability is also of secondary importance.

Although the HomePlug specification is the first active approach to standardize the PLC technology, it is still not complete. For example, it does not cover the coexistence between the access PLC and the home PLC. Most importantly, it does not provide any specific regulations on the Electromagnetic Interference (EMI) and Electromagnetic Compatibility (EMC) for PLC. It simply follows the Federal Communication Commission (FCC) Part 15, which is an American regulation for radio frequency devices. However, since the power distribution networks have different configurations in different countries, the EMC regulation in one country may not be suitable for other countries. In this case, some countries have taken the lead to champion their own national EMC requirements. For example, Germany has enacted NB30 and United Kingdom has defined MPT1570. The International Electrotechnical Commission (IEC) also suggested CISPR. Furthermore, in order to address the EMC problem and facilitate high bit rate PLC technology, ETSI and CENELEC are currently collaborating to produce new standards. The goal of this joint

work is to produce a standard with a common radiated emissions limit for the wired communication technologies.

1.1.2.2 Various PLC Technologies

Because of the non-standardized condition, the current PLC market is a mixture of different communication technologies. Some companies, like Polytrax in German and Mitsubishi in Japan, still make use of the narrow band (< 500 KHz). Their products closely follow the frequency bands and signal levels specified in European standard EN 50065, which is originally initiated for the power line control and automation. These companies claimed that the broadband companies, which use frequency band larger than 1 MHz, would finally meet the potential EMI/EMC problems. They argued that although currently there is no standard limiting the frequency bands and signal levels in power line in the MHz frequency range, that doesn't mean none will be enforced in the future. However, majority of the PLC companies still put their interests on the development of the broadband products since the high speed is always the most attractive point to the costumers. These companies (mainly come from US, like Intellon and Inari) claimed that they strictly follow the FCC CFR47 Part 15,

In the following table, the technologies of three typical PLC companies are compared.

Developer (Vendor)	Organization	Access Method (Protocol)	Modulation	Product Transfer Rate (bps)	Carrier Bandwidth
Polytrax	Member of PLCforum, EchoNet, HomePlug	In development	DMT with OFDM/QAM	2M	≤450KHz
Intellon	Former Member of HomePlug Member of PLCforum	CSMA/CA	OFDM	14M	3.5~16.5MHz
Inari	Member of CEA R7.3	DSMA and CTP	Multi-carrier DBPSK/DQPSK	2M~12M	>1MHz

CSMA/CA: Carrier Sensing Multiple Access with Collision Avoidance
DMT: Discrete Multi-Tone
QAM: Quadrature Amplitude Modulation
DSMA: Datagram Sensing Multiple Access
CTP: Centralized Token Passing
DBPSK: Differential Binary Phase Shift Keying
DQPSK: Differential Quadrature Phase Shift Keying

Table 1.1. Comparison of Typical PLC Technologies

Basically, Polytrax is the representative for the narrow band PLC. Intellon and Inari are broadband ones but they belong to different organizations. From the comparison of the transfer rate and the bandwidth, Polytrax's product has already optimize the transfer rate in the mentioned bandwidth, which means that, in this limited frequency band, the transfer rate can not be further increased anymore under the signal level limitation of the EN 50065. Comparing with Polytrax, Intellon and Inari have the advantage in data rate (both larger than 10Mbps) since they have a much broader bandwidth to use. But because of the different applications they focus on, they are using different access protocols. Intellon addresses mainly on home networking and it uses the CSMA/CA protocol (IEEE 802.11), which is similar to the Wireless LAN (Local Area Network). This protocol is more suitable for the data networking. Inari uses a hybrid protocol DSMA and CTP, which leverages the most effective features of existing protocols and make it suitable for both data networking and equipment control and automation.

Presently, it is very hard to say which company's products will capture the market in the future because beside the technology used, another important factor for the market is the applications. In the following sub sections, some popular PLC applications are formulated and analyzed.

1.1.3 PLC Applications

With the varying technical visions, different types of the PLC applications have been targeted in the markets.

1.1.3.1 Control and Automation

This application was started many years ago. The low bit rate implementations like power utility Automatic Meter Reading (AMR), lighting control and power line intercoms are widely used nowadays. With the idea of home automation coming in, people are thinking of using PLC to control the shades, garage door, heating systems and other devices in the house. These may be the future mainstreams for people. However, the development of a common protocol for controlling multiple appliances is the major challenge for this application.

1.1.3.2 Access

Access or “last-mile” applications require relatively high data speeds to assure delivery of DSL/Cable-like speeds for household users. This can provide a real alternative to the national telecommunication networks. Among the other access mediums, the optical fiber has the advantage of high throughput but the disadvantage of high cost. The coaxial cables and twisted pair cables have medium cost in installation but lower throughput than optical fiber. In practice, the optical fiber is normally used to connect zone access point to the network backbone and the distribution from the zone access point to the household users is generally done via the coaxial or twisted pair cables.

Figure 1.2. Illustration of PLC Access system

Compared with the other access medium, the advantage of PLC is the low installation cost since the power lines have already been laid in every house. In addition, PLC provides the convenience for the users since all of the power sockets can be used as the access terminals. A typical PLC access system is illustrated in Figure 1.2.

1.1.3.3 Home Networking

The power line network is intended for connecting home appliances and services to the electricity source. Home appliances, in addition to be connected to the electricity source through the power line network, are also interconnected to each other through that same network. This is a very important value for home networking since home networking is for interconnecting PCs, appliances and devices at home in order to share information between them, as illustrated in Figure 1.3.

Figure 1.3. Illustration of PLC Home Networking

	Phone Line	Ethernet	Power Line	Wireless
Speed (bps)	100K~10M	10M~100M	>10M	700K~54M
Estimated Cost (In US dollars)	\$50~\$130	\$75~\$200	\$70~\$150	\$200~\$300
Advantages	Convenient, simple (no new wires), secure	Fastest, most secure and reliable	Convenient, simple (no new wires)	Convenient, mobile, simple (no wires), secure
Requirements	Need computers and peripherals near phone jackets on the same phone line	Requires special cabling; best in new home installations or remodels	Need computers and peripherals near power outlets on the same power circuit	Network components must be within a 250-foot range

Table 1.2. Home Networking Technology Comparison

In Table 1.2, power line is compared to other home networking technologies such as phone line, Ethernet and wireless systems [7]. Among the four technologies, Ethernet has the fastest speed, but its installation is troublesome and expensive since it needs the additional cabling. Wireless systems should be the most convenient one, because all the components can be moved freely without the limitations of the wires. However, the wireless systems have expensive access units (for both Bluetooth and Home RF) and the distance limitations are quite obvious when compared to wired systems. Both power line and phone line do not need additional cabling. Phone line has the advantage of low cost, but in term of speed and system requirements, phone line is not as good as power line.

Recently, with the appearance of the fast PLC products, the concept for PLC home networking has been greatly expanded. It not only allows PCs and its peripherals to

communicate with each other, but also introduce new home entertainments like power line gaming, audio and video systems. As a PLC application, home networking has two potential advantages: firstly, 10Mbps is sufficient for intra-networking inside a house; secondly, the distance between power line access points inside a house is normally short resulting in lower signal attenuation.

1.1.4 Suitable Broadband PLC Implementation in Singapore

Singapore is one of the countries which are on the fore front of implementing broadband PLC technologies. PLC made its debut in Singapore in the form of a one-year technology trial conducted by SP Telecom at the Singapore Polytechnic campus from November 2000 to October 2001. The test results obtained were satisfactory and mirrored the success of more than a dozen corresponding trials in Europe. The trial results also demonstrated that the PLC can function safely without compromising the power supply and normal telecommunications service standards. After this, a half-year market trial starting from April of 2002 was conducted to test out the technical implementation of PLC on a wider scale and to evaluate its commercial viability. Although the market trial also seems to be successful, the commercial deployment was delayed by the Infocomm Development Authority (IDA) of Singapore due to the uncertain EMI regulation of the technology [8]. This is because unlike telecommunication cables, electrical cables are not properly shielded and balanced and therefore have potential for unintended radio frequency emissions. But this will not be a long-term problem since IDA has already started working on this and they claimed that a specific PLC EMI regulation for Singapore will be finalized by the end of year 2003.

However, there is another problem, which is how the PLC competes with the existing telecommunication services in Singapore. The key point of this problem is to find out suitable applications. To do this, the structure of the distribution system needs to be studied. Figure 1.4 shows a typical distribution system in Singapore. After the distribution transformer, the system is in a ring form and supplies to about 5~6 high-rise buildings, each having around 100 households. According to this structure, a possible PLC application model can be suggested. From the system point of view, it can be estimated that it will not be feasible to provide the access solution just after the LV distribution transformer. As the power line is a shared medium, the large number of households in the

distribution ring will drastically limit the data rate for each user. Suppose there are five high-rise buildings in the supply ring (as in Figure 1.4), since there are about 100 households in each building, which means there will be around 500 households sharing the power line medium after the transformer. Assume that there is one user per household and the PLC master modem has the speed of 27Mbps (the downstream data rate for DS2 modem), then each user can obtain a mere 54Kbps, which is not a competitive speed.

Figure 1.4. Simplified Structure of the Power Network in Singapore after Distribution Transformer

As a result, the PLC “last mile” for Singapore residential quarters is suggested to be in the intra-building scope only (as shown in Figure 1.5). The PLC range will be limited from the main distribution board of each building to all homes within the building. In this case, the averaging transfer rate for each user can go up to 270Kbps (since there are only around 100 users per building), which is much faster than the speed of the ISDN (Integrated Services Digital Networks) line.

Figure 1.5. A PLC Access System with Master Access Point after the Main Distribution Board

There is one potential advantage for Singapore to deploy PLC --- the house-service power lines are normally laid inside the metal conduits. This has provided a natural protection for

the EMC problem since the grounded metal conduit can be regarded as the Electromagnetic (EM) shield. In addition, the largest telecommunication company in Singapore --- Singtel, is now pushing the “Fiber to House” project [9]. It aims to provide a complete island-wide network of fiber optic cables that will connect housing estates and commercial buildings by the year 2005. This has provided a very good precondition for the implementation of the intra-building PLC access system. For instance, the PLC access system can be easily connected to the fiber network in the housing estate and provide a very cheap “last mile” solution from fiber to the house.

1.1.5 Research Motivations

As a telecommunication service, PLC can be very attractive, but as a technology, it still faces some challenges. As mentioned in the previous section, the use of electric network is not unlimited. When occupying a frequency band from about 9kHz to 30MHz, existing services such as long wave, medium-wave, and short-wave radio and amateur radio bands are overlapped. To minimize these interferences, the frequency allocation and the power limits for the broadband PLC need to be standardized. As a result, the main objective of many PLC researchers is how to optimize the PLC performance under the constraints stipulated by the various standards. This requires better understanding of the PLC channel --- the power line.

Unlike other wired communication mediums like the optical fiber, Unshielded Twisted Pair (UTP) and coaxial cables, LV power lines are designed to distribute the electrical energy, operating at 120/230V and 50/60Hz. The physical topology of the home wiring, the physical properties of the electrical cabling, and the appliances connected all play a significant part in the use of power line as a communication medium. From the literatures, the power line medium presents an extremely harsh environment on which to operate a communication system. The three channel parameters of noise, impedance and attenuation are highly variable with time, frequency and location of the transceivers [10-17]. These are the major hindrances for reliable communication through power lines. In order to overcome these difficulties, some assessments on the characteristics of the power line medium itself is important for the design and implementation of any PLC system.

Furthermore, it is necessary to determine how these characteristics impact on the performances of any communication schemes. When designing a communication system, it is always convenient to construct models that can reflect the important characteristics of the medium and also be flexible enough to cope with the changes in the channel. With the help of the channel model, different modulation schemes and coding techniques can be tested on the channel and their performances can be evaluated.

In this research, a computer model of the power line channel has been developed based on the following motivations:

1. To give thorough understanding into the characteristics of the power line channel.
2. To discover the causes of the various uncertainties in the channel and to find solutions to overcome them.
3. To provide a controllable channel model of the power network that can be used for designing PLC scheme.
4. To optimize the communications over the channel with the help of an accurate representation of the power line channel.

1.2 OBJECTIVES

To meet the motivations mentioned above, the primary objective of this research is to systematically determine the reasons behind the unfavorable responses from the power line networks to the high frequency communication signals. It is envisaged that a more superior communication scheme can only be achieved with an accurate and flexible channel model since the model can clearly identify the problems and defects in the channel so that the system designers can avoid or combat them. This objective can be translated into more detailed targets as follow:

1. Analyze the power line responses based on the multi-conductor transmission line theory and derive its line parameters.
2. Derive the transfer characteristics and the input impedances of the power line channel under different network configurations and load conditions.
3. Study the accuracy, applicability and sensitivity of the derived parameters to the errors in estimating the physical quantities of the power lines.

4. Statistically study the noise behavior on the power line and derive its corresponding mathematical model.
5. Form a PLC channel model based on the derived transfer characteristic, input impedance, and noise. Apply the model to study the properties of PLC channel encountered in the Singapore environment.
6. Using the knowledge of the channel behavior, design modulation schemes that are best matched to the channel conditions.
7. Designing the necessary hardware and software for conducting practical measurements on the power line.

1.3 MAJOR CONTRIBUTIONS OF THE THESIS

The broadband PLC research team in Nanyang Technological University of Singapore is formed by quite a few professors and research students from different fields. Currently, the researches of the team are concentrating on the following four topics: (1) Channel modeling and characterization. (2) MAC (Medium Access Control) layer design and implementation. (3) EMI/EMC study. (4) Home control and automation. The focus of this research is on the first topic, channel modeling and characterization. The major contributions of this thesis can be summarized as follows:

- (i) The idea of using the multi-conductor transmission line to analyze the power line is proposed. In the actual installations, power networks are normally constructed using different types of power cables. The analysis based on multi-conductor transmission line has provided a general and accurate solution for modeling various types of power lines with different number of conductors and line structures.
- (ii) A frequency domain network model to determine the transfer function and the input impedance of the tree topology network is proposed based on the scattering parameter method. Measurements have been conducted to verify the accuracy and the flexibility of the model. The sensitivity of the model has also been studied according to different network configuration parameters.

- (iii) A network model for the ring topology network has been proposed to fill the gap in studying this type of network. The model is derived based on the derived tree topology network model and the concept of the scattering parameters. The model is able to consider all the complex signal reflections in the ring topology network and verification measurements have also demonstrated the accuracy of the model.
- (iv) The noise on the power line is modeled using a statistical method. Specifically, the amplitude spectrum of power line background noise at each single frequency is proposed to be modeled by the Nakagami-m PDF (Probability Density Function). The total appliance noise at a receiver is based on the measurement of the noise spectrum of every single appliance and the derived transfer function model of the network.
- (v) Some possible uses of the derived channel model in designing of OFDM system over the power network are suggested. These include how to apply the transfer function model to determine the OFDM system parameters, how to apply the noise model to predict its performance, and how to apply both the transfer function model and the noise model to optimize the system overall performance.
- (vi) Hardware devices, including signal coupler and line trap, for high-frequency measurement on the real power network have been developed. The function of the signal coupler is to prevent the power signal (230V, 50Hz) from damaging the measurement instrument, while at the same time allow the high frequency communication signal to pass through. The line trap is actually a high pass filter. It provides isolation between the network under test and the external power network. In addition, in the measurement of the impedance of a single appliance, it provides a high impedance reference.

1.4 ORGANISATION OF THE THESIS

There are ten chapters in this thesis. The first two chapters are introducing chapters. An overview of the current PLC industry, some suggestions on possible PLC implementation

in Singapore, the motivations and objectives of the research, and the major contributions of this thesis are outlined in Chapter 1. Chapter 2 gives detailed descriptions on the PLC channel parameters in terms of transfer function, input impedance, and noise. Literature reviews and the modeling approach proposed in the research to derive the three channel parameters are also given in this chapter.

Chapters 3 to 8 are the main body of the thesis describing in detail the various channel modeling approaches. Based on [56], Chapter 3 lays out the fundamentals of Multi-Conductor Transmission Line (MTL) equations and shows how the power line can be analyzed as MTL. Based on these equations, the expressions of the two intrinsic line parameters namely, the characteristic impedance and the propagation constant are derived. Chapter 4 focuses on the determination of the line parameters according to per-unit-length primary parameters of the line. Chapter 5 describes how to use the line parameters to derive the transfer function and input impedance models for tree topology networks. Verifications and the sensitivity study of the derived model are also stated in this chapter. The corresponding model derivation for ring topology networks is described in Chapter 6. In Chapter 7, a PLC noise model is developed based on frequency domain approach. As an ending chapter of the main body, the use of the derived channel model in the design of an OFDM system for PLC is described in Chapter 8.

Finally, the conclusions of this work and the recommendations for future research are presented in Chapter 9. The information will be useful for readers who wish to conduct further research in this field.

In Appendix A.1, the hardware and software designs for high frequency measurement on power network are introduced. In addition, the mathematical derivation to simplify the progressive matrices summation, which serves as a reference for the derivations described in Chapter 6, is also given in Appendix A.2.

Chapter 2

MODELING APPROACHES OF CHANNEL PARAMETERS

To provide an overview of the modeling approaches of PLC channel, this chapter focuses on introducing the characteristics of LV power line channel in terms of transfer function, input impedance, and noise. At the same time, the detailed literature works on methods and techniques for modeling these channel parameters are described and new modeling techniques are proposed.

The PLC channel is characterized by the three channel parameters namely, transfer function, input impedance, and noise. Among the three parameters, transfer function and noise are important for designing the modulation schemes and the coding techniques; while input impedance is the essential factor for designing the modem matching circuits. Although these parameters are highly variable with respect to time, frequency and locations in the power network, it is not difficult to show that they mainly depend on the configurations and load conditions of the power network. The network configuration typically remains unchanged except when short and/or open circuits happen. However, there are frequent turning-on and turning-off of electrical devices and appliances, and this will change the impedance distributions in the network, so that both the transfer function between transceivers and the network input impedances seeing by the transceivers can vary greatly. In addition, any load in the network can be a source of noise, and changes in the loading condition may also change the noise pattern. To better understand the PLC channel, detailed characteristics of the three channel parameters and the corresponding literature reviews are introduced in the following sections.

2.1 TRANSFER FUNCTION AND INPUT IMPEDANCE

2.1.1 Characteristics of Transfer Function and Input Impedance

The channel transfer function is the complex gain or attenuation to the communication signal between two points in a channel. Since power line is generally passive in the high

frequency band, the transfer function in power line normally means attenuation. The following are the major types of attenuations in PLC channel:

- Line attenuation, which is mainly caused by the heat loss and radiation loss on the power line. This attenuation is always present and it depends on the length of the wave path and the frequency of the wave.
- Mismatch attenuation, which is due to the impedance mismatches between the power network and the PLC transceiver. This type of attenuation can always be minimized through proper design of matching circuit in the transceiver.
- Multi-path attenuation, which is caused by multiple reflections at the points of impedance discontinuities in the channel. The reflected wave from these unmatched points will interfere the original wave. The interferences can be constructive or destructive; if destructive, it will cause attenuation.

The focus of this research is on line attenuation and multi-path attenuation since the mismatch attenuation can always be controlled by matching circuits in the PLC transceivers.

The power line input impedance is a highly fluctuating quantity. It not only depends on the specific loads that are connected in parallel to the power line at specific times, but also on the characteristic impedance of the cables used. This impedance is important for channel characterization since its uncertainties directly affects the uncertain characteristics of the channel transfer function.

2.1.2 Network Modeling Techniques

The modeling of transfer function and input impedance is normally referred as network modeling because both parameters can be determined with the knowledge of power network properties like cable characteristics, network topologies and loading conditions. Previous works on PLC network modeling can be generally divided in two phases:

Phase 1: Line modeling

Phase 2: Network modeling according to topologies

2.1.2.1 Line Modeling

In the line modeling phase, model parameters of the power line such as characteristic impedance and propagation constant are obtained. From the ways these model parameters are obtained, the line modeling approach can be classified as empirical approach or analytical approach. In the empirical approach, the model parameters are retrieved from measurements [17, 19-24]. This approach is also called top-down approach in [24]. This 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 [25-27], and it is called bottom-up approach in [24]. In this approach, the parameters are derived according to the physical properties of the power line, including the properties of its conducting material and insulating dielectric material, and its orientation. Although this approach requires more computational efforts compared to the empirical approach, it however describes 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.

2.1.2.2 Network Modeling according to Topologies

In the second phase, the model parameters obtained in the first phase are used to determine the network parameters of the power networks with different network topologies. Two types of network topology have been studied in this research, which are tree topology and ring topology, since they are commonly used in power networks.

For the tree topology network, depending on the algorithms used, the network parameters can be determined in the time domain [17, 19, 24, 25] or frequency domain [20-23, 26, 27]. For the time domain algorithm, the power line channel is regarded as a multi-path environment and an echo model is developed to represent this physical characteristic. However, this modeling approach is not accurate enough since normally only the dominant paths will be considered in the algorithm. This can be based on the approximation that the backward reflections from impedance discontinuities are negligible [24, 25] or based on the assumption that the non-peak impulse responses are ignored [17,

19]. In addition, the echo model also becomes complex when multiple branches are connecting at the same joint. For the frequency domain algorithm, 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. The main advantage of the frequency domain algorithm is its ability to consider all the signals reflected from the discontinuities regardless of the complexity of the network.

In this research, for the tree topology network, analytical approach is adopted due to its accuracy and flexibility. Both time and frequency domain algorithms are studied and their performances are compared. The frequency domain algorithm in the literature is based on transmission matrix method [20-23, 26] but in this research, the scattering matrix method [27] is proposed instead, due to the following reasons.

Figure 2.1. Comparison of Transmission Matrix and Scattering Matrix of 2-Port Network

As Figure 2.1 shows, the transmission matrix gives the relationship of the voltage (V) and current (I) at the two terminals while the scattering matrix gives the relationship of the incident (a) and reflected (b) waves. The use of scattering matrix will be more convenient in the modeling of network parameters at high frequency. Take the transfer function as an example, the transfer function derived from the transmission matrix is basically V_2/V_1 , which is the transfer factor of the standing waves (including both incident and reflected waves). However, PLC is only interested in the transfer function in the forward direction, which is the ratio of the incident power into the receiver over the power injected by the transmitter. This can readily be expressed by b_2/a_1 or S_{21} in the scattering matrix.

With the above arguments, a novel analytical approach based on the frequency domain modeling using scattering matrix is proposed in this research. The two model parameters, the characteristic impedance and the propagation constant, are first derived using the multi-conductor transmission line theory. Then the frequency domain algorithm based on the scattering matrices is applied to account for the many branches in the power line.

2.2 NOISE

2.2.1 Characteristics of noise

In low-voltage power lines, the source of noise can be from internal (inside the power network) or external (outside the power network). Specifically, it is distinguished into the following types in [18]:

- Coloured background noise with a relatively low Power Spectral Density (PSD), varying with frequency. This type of noise is mainly caused by summation of numerous noise sources with low power. Its PSD varies over time in terms of minutes or even hours.
- Narrow band noise, mostly sinusoidal signals, with modulated amplitudes caused by ingress of broadcasting stations. The level is generally varying over the day.
- Periodic impulsive noise asynchronous to the mains frequency with a repetition rate between 50 to 200 kHz, with a discrete line spectrum spaced according to the impulse repetition rate. This type of noise is mostly caused by switched power supplies.
- Periodic impulsive noise synchronous to the mains frequency with a repetition rate of 50 or 100 Hz. The impulses are of short duration and have a PSD decreasing with frequency. This type of noise is caused by power supplies, mainly by the switching action of rectifier diodes, which occurs synchronously with the mains cycle.
- Asynchronous impulsive noise is caused by switching transients in the network. The impulses have durations of some microseconds up to a few millisecond with random occurrence. The PSD of this type of noise can reach values of more than 50 dB above the background noise.

This classification of power line noise is popular and widely used by researchers of PLC. For the sake of simplicity, the first two types of noise can be summarized as background noise, and the last three types as impulsive noise.

2.2.2 Noise Modeling

Noise modeling is quite independent and different from modeling of the other two channel parameters, channel transfer function and input impedance. The power line noise modeling is based mainly on measurements and currently there is no complete theoretical model because of the complex nature of the noise.

From the previous section, it is known that generally the power line noise can be classified into background noise and impulsive noise. Since the impulsive noise is actually generated by different appliances in the power network, the power line noise can also be regarded as the combination of the background noise and the sum of the noises from all appliances. Since noise is very hard to be characterized through purely analytical derivation, all of the existing noise models are based on measurements.

According to the measurement in different domains, the noise modeling approaches in the literature can be classified into frequency-domain approach and time-domain approach. The frequency-domain approach is based on the measurement of noise spectrum, while the time-domain approach is based on the measurement of real noise waveforms. In the past researches, the background noise is mainly modeled in the frequency domain, while the impulsive noise is characterized in both the frequency and time domains.

There are two methods to model the background noise in the frequency domain. The first one is the curve fitting method [22, 29-31], which is to fit the measured PSD [22, 30, 31] or voltage spectrum density [29] of the background noise into certain functions of frequency. However, this method only illustrates the rough changes of the noise spectrum according to frequency, and it does not give any information on the random behavior of the noise at each single frequency. To show the properties of the background noise at each frequency, another method, the statistical analysis method [32-34], needs to be used. This method characterizes the background noise in single frequency into certain PDFs so that with the corresponding statistical parameters like mean and standard deviation, these PDFs

can fully describe the statistical characteristics of the noise at an individual frequency. The proposed PDF in the existing noise models include the “sum of two Rayleigh” distribution [32], the lognormal distribution [33], and the Gaussian distribution [34].

The frequency-domain approach for modeling the impulsive noise is purely based on measurements [35-37]. The noise spectrum generated by some household appliances are measured so that their patterns in the frequency domain can be studied. Although this frequency-domain approach can not represent the random changing features of the impulsive noise, it however gives sufficient information on the impact of the impulsive noise in the PLC operating frequency range. This will be very useful in determining the system Signal to Noise Ratio (SNR) under different bit loading conditions especially when the multi-carrier modulation technique is used.

The time-domain impulsive noise model is characterized by three parameters namely, the pulse amplitude, the pulse width, and the Inter Arrival Time (IAT). From the literature, the time-domain modeling of impulsive noise is also focus on finding the statistical characteristics of these three parameters. Like in [38-41], the probability distribution curves of the three parameters are constructed from the measurement results. Recently, based on the assumption that the pulse width and the IAT correspond to the superposition of several exponential distributions, some researchers proposed to use the partitioned Markov chain [42] to model the statistical properties of pulse width and the IAT [18, 43]. The transition probability matrices used in the model are retrieved from measurement. Comparing with those methods [38-41] based on direct measurement, this method retrieves more information from measurement. It provides not only the probability distributions of these two parameters but also the time transition properties of the distributions. In addition, some research works [44-46] directly applied the Middleton’s class A noise model [47] to model the amplitude distribution of power line impulse noise. However, since this model is originally developed for the manmade impulse interference, its accuracy in modeling the impulsive noise in the power network is still lacking detailed study.

Besides all the above-mentioned models, several Japanese researchers have proposed using the cyclo-stationary noise model [48-50] to characterize the background noise and the impulsive noise together. This model is based on the finding that the noise

characteristic in power line changes synchronously with a half cycle of the power supplying voltage, so that the noise can be regarded as the sum of sinusoidal waves with different amplitudes and frequencies.

In this research, since the frequency-domain approach is adopted though out the whole analysis, the focus will is still on the frequency-domain characterization of the power line noise. The statistical analysis method is used to characterize the background noise while the impulsive noise is studied from the measured appliances' noise spectrums. Although the discussion above is mainly on the background noise and the impulsive noise found on the power line, but for a communication system designer, the most important noise is the noise appearing at the receiver. In order to make the noise model more practical, the measured appliances' noise spectrums are combined with the network transfer function model to obtain the noise reaching the receiver. As a result, with the knowledge of the configuration and the loading condition of a particular power network, the proposed model as described in Chapter 7 is able to provide the noise spectrum at any position within the power network.

In the following Chapters 3~7, the details of the proposed modeling techniques for transfer function, input impedance and noise are discussed.

Chapter 3

THEORY OF MULTI-CONDUCTOR TRANSMISSION LINE

The theory used in the first phase of the network modeling, which is the line modeling, is discussed in this chapter. Specifically, the LV power line is analyzed as a Multi-conductor Transmission Line (MTL) and the corresponding transmission line equations and line 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 cables (live, neutral and earth) are usually laid inside metal conduits embedded inside the wall for the purpose of protection, as shown in Figure 3.1. These cables are made up of stranded copper conductors with PVC (Polyvinyl Chloride) insulation. Although this typical system is used as an example in this thesis, cables with different conductor sizes, strands, and layouts can also be modeled by just changing the parameters inside the model.

Figure 3.1. Cross-Sectional View of the House Service Power Line

This structure of power line can be approximated as an MTL when it is used to transmit the high frequency communication signals. The term MTL typically refers to a set of $(n+1)$ parallel conductors that serve to transmit electrical signals between sources and loads. 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 whose frequencies range from dc to where the line cross-sectional dimension becomes a significant fraction of the wavelength.

The first assumption for the MTL approximation of power line, which is also the fundamental assumption for all transmission line analyses, is that the wave propagation on the power line follows TEM mode. This requires the line structure to have electrically small cross-sectional dimensions. In other words, the distance of separation between the conductors is small in comparison with the operating wavelength. This requirement can be easily fulfilled because in the PLC, the operating wavelength can be obtained by

$$\lambda = \frac{\lambda_{air}}{\sqrt{\epsilon_r}} \quad (3.1)$$

where λ_{air} is the wavelength in free space and ϵ_r is the relative permittivity of the dielectric material. The dielectric between the conductors is a mixture of air and insulation material. Since the PLC frequency range in this research is taken as between 1MHz-30MHz (following ETSI requirements for 1st generation PLC [51]), λ_{air} has the value of 300m ~ 10m and ϵ_r for the PVC insulation is from 3.8 ~ 2.7782 [52]. Since the ϵ_r value for air is 1 in the frequency range, the mixture's ϵ_r must be less than that of the PVC. It is quite obvious that the operating wavelength is typically much larger than the distance between cables, which is in the order unit of several mm.

The other requirement of the TEM mode propagation is that the power line should be in an uniform structure, which means that, the conductors in the structure should have uniform cross section along their length and are parallel to each other. In the actual installation, the power cables are simply pulled through the conduit and the separation between them is not uniform at all. However, the metal conduit normally has small cross-sectional area and this limits the variation of the separation between the cables. Hence, the assumption of uniform structure is reasonable in this case.

Based on the above considerations, the transmission line equations and the two important line parameters, the characteristic impedance and propagation constant, are derived [56].

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. In this chapter, the Maxwell equations will be used to derived MTL equations.

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

Figure 3.2 shows the general $(n+1)$ -conductor line to be considered. It consists of n conductors and a reference conductor (earth conductor), which is denoted as 0^{th} conductor in the figure. Recall the Faraday's law in integral form [53]:

$$\oint_c \vec{E} \cdot d\vec{l} = -\mu \frac{d}{dt} \int_s \vec{H} \cdot d\vec{s} \quad (3.2)$$

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

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

c : Contour of field volume

s : Surface of field volume

μ : Permeability of conductor (H/m)

Applying this to the contour 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 [56]

$$\int_a^{a'} \vec{E}_t \cdot d\vec{l} + \int_a^{b'} \vec{E}_z \cdot d\vec{l} + \int_b^b \vec{E}_t \cdot d\vec{l} + \int_b^a \vec{E}_z \cdot d\vec{l} = \mu \frac{d}{dt} \int_{s_i} \vec{H}_t \cdot d\vec{s} \quad (3.3)$$

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 of $\vec{H}_t$, and the right-hand rule, the minus sign on the right-hand side of (3.2) is absent in (3.3).

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

$$V_i(z, t) = - \int_a^{a'} \vec{E}_t(x, y, z, t) \cdot d\vec{l} \quad (3.4a)$$

$$V_i(z + \Delta z, t) = - \int_b^{b'} \vec{E}_t(x, y, z + \Delta z, t) \cdot d\vec{l} \quad (3.4b)$$

The integrals along the z direction are zero if the conductors are considered to be perfect conductors. It was pointed out that the TEM mode can not 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 drop 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,

$$\int_a^{a'} \vec{E}_z \cdot d\vec{l} = R_i \cdot \Delta z \cdot I_i(z, t) \quad (3.5a)$$

$$\int_b^b \vec{E}_z \cdot d\vec{l} = R_0 \cdot \Delta z \cdot \sum_{k=1}^n I_k(z, t) \quad (3.5b)$$

Substituting (3.4) and (3.5) into (3.3) yields

$$-V_i(z, t) + R_i \cdot \Delta z \cdot I_i(z, t) + V_i(z + \Delta z, t) + R_0 \cdot \Delta z \cdot \sum_{k=1}^n I_k(z, t) = \mu \frac{d}{dt} \int_{s_i} \overline{H}_t \cdot ds \quad (3.6)$$

Dividing both sides by Δz , and rearranging the equation gives

$$\begin{aligned} & \frac{V_i(z + \Delta z, t) - V_i(z, t)}{\Delta z} \\ &= -R_0 I_1 - R_0 I_2 - \dots - (R_0 + R_i) I_i - \dots - R_0 I_n + \mu \frac{1}{\Delta z} \frac{d}{dt} \int_{s_i} \overline{H}_t \cdot ds \end{aligned} \quad (3.7)$$

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 on the conductors. This total magnetic flux can be written as

$$\begin{aligned} \psi_i &= -\mu \lim_{\Delta z \rightarrow 0} \frac{1}{\Delta z} \int_{s_i} \overline{H}_t \cdot ds \\ &= L_{i1} I_1 + L_{i2} I_2 + \dots + L_{ii} I_i + \dots + L_{in} I_n \end{aligned} \quad (3.8)$$

$L_{i1}, L_{i2}, \dots, L_{ii}, \dots$ and L_{in} are the per-unit-length inductances in (H/m). Substituting (3.8) into (3.7) and taking the limit as $\Delta z \rightarrow 0$ gives,

$$\begin{aligned} \frac{\partial V_i(z, t)}{\partial z} &= -R_0 I_1 - R_0 I_2 - \dots - (R_0 + R_i) I_i - \dots - R_0 I_n \\ &\quad - L_{i1} \frac{\partial I_1(z, t)}{\partial t} - L_{i2} \frac{\partial I_2(z, t)}{\partial t} - \dots - L_{ii} \frac{\partial I_i(z, t)}{\partial t} - \dots - L_{in} \frac{\partial I_n(z, t)}{\partial t} \end{aligned} \quad (3.9)$$

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

$$\frac{\partial}{\partial z} \mathbf{V}(z, t) = -\mathbf{R} \cdot \mathbf{I}(z, t) - \mathbf{L} \frac{\partial}{\partial t} \mathbf{I}(z, t) \quad (3.10)$$

where the voltage and current vectors are defined as

$$\mathbf{V}(z, t) = [V_1(z, t) \quad \dots \quad V_i(z, t) \quad \dots \quad V_n(z, t)]^T \quad (3.11a)$$

$$\mathbf{I}(z, t) = [I_1(z, t) \quad \dots \quad I_i(z, t) \quad \dots \quad I_n(z, t)]^T \quad (3.11b)$$

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

$$\mathbf{R} = \begin{bmatrix} (R_1 + R_0) & R_0 & \dots & R_0 \\ R_0 & (R_2 + R_0) & \dots & R_0 \\ \vdots & \vdots & \ddots & \vdots \\ R_0 & R_0 & \dots & (R_n + R_0) \end{bmatrix} \quad (3.12)$$

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.

$$\mathbf{L} = \begin{bmatrix} L_{11} & L_{12} & \cdots & L_{1n} \\ L_{21} & L_{22} & \cdots & L_{2n} \\ \vdots & \vdots & \ddots & \vdots \\ L_{n1} & L_{n2} & \cdots & L_{nn} \end{bmatrix} \quad (3.13)$$

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 $\hat{s}$ around the i -th conductor as shown in Figure 3.3.

Figure 3.3. Definition of the Surface for Derivation of the Second MTL Equation

The portion of the surface over the end caps is denoted as $\hat{s}_e$ while the portion over the sides is denoted as $\hat{s}_s$. Recalling the continuity equation of conservation of charge [53],

$$\oint_{\hat{s}} \vec{J} \cdot d\vec{s} = -\frac{\partial Q}{\partial t} \quad (3.14)$$

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

Over the end caps (surface $\hat{s}_e$), it can be easily written that [56]

$$\iint_{\hat{s}_e} \vec{J} \cdot d\hat{s} = I_i(z + \Delta z, t) - I_i(z, t) \quad (3.15)$$

Over the sides of the surface (surface $\hat{s}_o$), there are two currents: conduction current, $\vec{J}_c = \sigma \vec{E}_t$, and 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

$$\iint_{\hat{s}_o} \vec{J}_c \cdot d\hat{s} = \sigma \iint_{\hat{s}_o} \vec{E}_t \cdot d\hat{s} \quad (3.16)$$

This can again 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,

$$\begin{aligned} \sigma \lim_{\Delta z \rightarrow 0} \frac{1}{\Delta z} \iint_{\hat{s}_o} \vec{E}_t \cdot d\hat{s} &= G_{i1}(V_i - V_1) + \dots + G_{ii}V_i + \dots + G_{in}(V_i - V_n) \\ &= -G_{i1}V_1(z, t) - G_{i2}V_2(z, t) - \dots \\ &\quad + \sum_{k=1}^n G_{ik}V_k(z, t) - \dots - G_{in}V_n(z, t) \end{aligned} \quad (3.17)$$

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

$$Q = \epsilon \iint_{\hat{s}_o} \vec{E}_t \cdot d\hat{s} \quad (3.18)$$

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

$$\begin{aligned} \epsilon \lim_{\Delta z \rightarrow 0} \frac{1}{\Delta z} \iint_{\hat{s}_o} \vec{E}_t \cdot d\hat{s} &= C_{i1}(V_i - V_1) + \dots + C_{ii}V_i + \dots + C_{in}(V_i - V_n) \\ &= -C_{i1}V_1(z, t) - C_{i2}V_2(z, t) - \dots \\ &\quad + \sum_{k=1}^n C_{ik}V_k(z, t) - \dots - C_{in}V_n(z, t) \end{aligned} \quad (3.19)$$

Substituting (3.15), (3.16), and (3.18) into (3.14), and dividing both sides by Δz gives

$$\frac{I_i(z + \Delta z, t) - I_i(z, t)}{\Delta z} + \sigma \frac{1}{\Delta z} \iint_{S_0} \vec{E}_i \cdot d\vec{s} = -\varepsilon \frac{1}{\Delta z} \iint_{S_0} \vec{E}_i \cdot d\vec{s} \quad (3.20)$$

Substituting (3.20) into (3.17) and (3.19), and taking the limit as $\Delta z \rightarrow 0$ yields

$$\begin{aligned} \frac{\partial I_i(z, t)}{\partial z} = & G_{i1} V_1(z, t) + \dots - \sum_{k=1}^n G_{ik} V_k(z, t) + \dots + G_{in} V_n(z, t) \\ & + \frac{\partial}{\partial t} \left\{ C_{i1} V_1(z, t) + \dots - \sum_{k=1}^n C_{ik} V_k(z, t) + \dots + C_{in} V_n(z, t) \right\} \end{aligned} \quad (3.21)$$

Equation (3.21) can be written into the compact form with matrix notation, which forms the second MTL equation.

$$\frac{\partial}{\partial z} \mathbf{I}(z, t) = -\mathbf{G} \cdot \mathbf{V}(z, t) - \mathbf{C} \frac{\partial}{\partial t} \mathbf{V}(z, t) \quad (3.22)$$

where $\mathbf{V}$ and $\mathbf{I}$ are defined by (3.11). The per-unit-length conductance matrix, which represents the conduction current flowing between the conductors in the transverse plane, is defined as

$$\mathbf{G} = \begin{bmatrix} \sum_{k=1}^n G_{1k} & -G_{12} & \dots & -G_{1n} \\ -G_{21} & \sum_{k=1}^n G_{2k} & \dots & -G_{2n} \\ \vdots & \vdots & \ddots & \vdots \\ -G_{n1} & -G_{n2} & \dots & \sum_{k=1}^n G_{nk} \end{bmatrix} \quad (3.23)$$

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

$$\mathbf{C} = \begin{bmatrix} \sum_{k=1}^n C_{1k} & -C_{12} & \dots & -C_{1n} \\ -C_{21} & \sum_{k=1}^n C_{2k} & \dots & -C_{2n} \\ \vdots & \vdots & \ddots & \vdots \\ -C_{n1} & -C_{n2} & \dots & \sum_{k=1}^n C_{nk} \end{bmatrix} \quad (3.24)$$

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

Equation (3.10) and (3.22) 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 two line parameters, the characteristic impedance and the propagation constant, which describe the behavior of the wave propagation on the line, are 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.

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

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

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

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

$$\hat{I}_i(z) = I_i(z) \angle \phi_i(z) = I_i(z) e^{j\phi_i(z)} \quad (3.26b)$$

Substituting the phasor form of line voltage and current into MTL equations (3.10) and (3.22),

$$\frac{d}{dz} \hat{\mathbf{V}}(z) = -\mathbf{Z} \cdot \hat{\mathbf{I}}(z) \quad (3.27a)$$

$$\frac{d}{dz} \hat{\mathbf{I}}(z) = -\mathbf{Y} \cdot \hat{\mathbf{V}}(z) \quad (3.27b)$$

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

$$\mathbf{Z} = \mathbf{R} + j\omega\mathbf{L} \quad (3.28a)$$

$$\mathbf{Y} = \mathbf{G} + j\omega\mathbf{C} \quad (3.28b)$$

Combining (3.27a) and (3.27b) into a matrix form,

$$\frac{d}{dz} \hat{\mathbf{X}}(z) = \mathbf{A} \cdot \hat{\mathbf{X}}(z) \quad (3.29)$$

where

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

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

The solution of (3.29) 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 [56]. Firstly, the mode voltage $\hat{\mathbf{V}}_m(z)$ and current $\hat{\mathbf{I}}_m(z)$ can be defined as

$$\hat{\mathbf{V}}(z) = \mathbf{T}_V \cdot \hat{\mathbf{V}}_m(z) \quad (3.30a)$$

$$\hat{\mathbf{I}}(z) = \mathbf{T}_I \cdot \hat{\mathbf{I}}_m(z) \quad (3.30b)$$

The $n \times n$ complex matrices $\mathbf{T}_V$ and $\mathbf{T}_I$ 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 $\mathbf{T}_I$ must be nonsingular, i.e. $\mathbf{T}_V^{-1}$ and $\mathbf{T}_I^{-1}$ must exist. Substitute mode voltage and current into (3.29)

$$\frac{d}{dz} \begin{bmatrix} \hat{\mathbf{V}}_m(z) \\ \hat{\mathbf{I}}_m(z) \end{bmatrix} = \begin{bmatrix} 0 & -\mathbf{T}_V^{-1} \mathbf{Z} \mathbf{T}_I \\ -\mathbf{T}_I^{-1} \mathbf{Y} \mathbf{T}_V & 0 \end{bmatrix} \begin{bmatrix} \hat{\mathbf{V}}_m(z) \\ \hat{\mathbf{I}}_m(z) \end{bmatrix} \quad (3.31)$$

If $\mathbf{T}_V$ and $\mathbf{T}_I$ can be obtained such that $\mathbf{T}_V^{-1} \mathbf{Z} \mathbf{T}_I$ and $\mathbf{T}_I^{-1} \mathbf{Y} \mathbf{T}_V$ are diagonal as follows,

$$\mathbf{T}_V^{-1} \mathbf{Z} \mathbf{T}_I = \mathbf{Z}' = \begin{bmatrix} Z'_1 & 0 & \dots & 0 \\ 0 & Z'_2 & \ddots & \vdots \\ \vdots & \ddots & \ddots & 0 \\ 0 & \dots & 0 & Z'_n \end{bmatrix} \quad (3.32a)$$

$$\mathbf{T}_I^{-1} \mathbf{Y} \mathbf{T}_V = \mathbf{Y}' = \begin{bmatrix} Y'_1 & 0 & \dots & 0 \\ 0 & Y'_2 & \ddots & \vdots \\ \vdots & \ddots & \ddots & 0 \\ 0 & \dots & 0 & Y'_n \end{bmatrix} \quad (3.32b)$$

Then the MTL equations can be decoupled as

$$\begin{cases} \frac{d}{dz} V_{m1}(z) = -Z'_1 \cdot I_{m1}(z) & \frac{d}{dz} I_{m1}(z) = -Y'_1 \cdot V_{m1}(z) \\ \vdots & \vdots \\ \frac{d}{dz} V_{mn}(z) = -Z'_1 \cdot I_{mn}(z) & \frac{d}{dz} I_{mn}(z) = -Y'_1 \cdot V_{mn}(z) \end{cases}$$

The essential question becomes: How to find the similarity transformation matrices of $\mathbf{T}_V$ and $\mathbf{T}_I$ to make $\mathbf{Z}'$ and $\mathbf{Y}'$ diagonal? Combining with (3.27a) and (3.27b)

$$\frac{d^2}{dz^2} \hat{\mathbf{V}}(z) = \mathbf{Z} \mathbf{Y} \hat{\mathbf{V}}(z) \quad (3.33a)$$

$$\frac{d^2}{dz^2} \hat{\mathbf{I}}(z) = \mathbf{Y} \mathbf{Z} \hat{\mathbf{I}}(z) \quad (3.33b)$$

Substituting in the similarity transformations as given in (3.30), gives

$$\begin{aligned}\frac{d^2}{dz^2} \hat{\mathbf{V}}_m(z) &= \mathbf{T}_V^{-1} \mathbf{Z} \mathbf{Y} \mathbf{T}_V \hat{\mathbf{V}}_m(z) = \mathbf{T}_V^{-1} \mathbf{Z} \mathbf{T}_I \mathbf{T}_I^{-1} \mathbf{Y} \mathbf{T}_V \hat{\mathbf{V}}_m(z) \\ &= \mathbf{Z}' \mathbf{Y}' \hat{\mathbf{V}}_m(z)\end{aligned}\quad (3.34a)$$

$$\begin{aligned}\frac{d^2}{dz^2} \hat{\mathbf{I}}_m(z) &= \mathbf{T}_I^{-1} \mathbf{Y} \mathbf{Z} \mathbf{T}_I \hat{\mathbf{I}}_m(z) = \mathbf{T}_I^{-1} \mathbf{Y} \mathbf{T}_V \mathbf{T}_V^{-1} \mathbf{Z} \mathbf{T}_I \hat{\mathbf{I}}_m(z) \\ &= \mathbf{Y}' \mathbf{Z}' \hat{\mathbf{I}}_m(z)\end{aligned}\quad (3.34b)$$

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

$$\begin{aligned}(\mathbf{T}_V^{-1} \mathbf{Z} \mathbf{Y} \mathbf{T}_V)^T &= (\mathbf{T}_V)^T (\mathbf{Y})^T (\mathbf{Z})^T (\mathbf{T}_V^{-1})^T \\ &= (\mathbf{T}_V)^T \mathbf{Y} \mathbf{Z} (\mathbf{T}_V^{-1})^T \\ &= \mathbf{Z}' \mathbf{Y}' \\ &= \mathbf{Y}' \mathbf{Z}'\end{aligned}\quad (3.35)$$

Since the required $\mathbf{Z}'$ and $\mathbf{Y}'$ are diagonal, comparing (3.35) to (3.34b), the following relationship can be obtained,

$$\mathbf{T}_I^T = \mathbf{T}_V^{-1} = \mathbf{T} \quad (3.36)$$

Substituting (3.36) into (3.34b),

$$\frac{d^2}{dz^2} \hat{\mathbf{I}}_m(z) = \mathbf{T}^{-1} \mathbf{Y} \mathbf{Z} \mathbf{T} \hat{\mathbf{I}}_m(z) = \boldsymbol{\gamma}^2 \hat{\mathbf{I}}_m(z) \quad (3.37)$$

where $\boldsymbol{\gamma}$ is the diagonal matrix that represents the propagation constants between different conductors and the reference conductor.

$$\boldsymbol{\gamma}^2 = \mathbf{T}^{-1} \mathbf{Y} \mathbf{Z} \mathbf{T} = \begin{bmatrix} \gamma_1^2 & 0 & \dots & 0 \\ 0 & \gamma_2^2 & \ddots & \vdots \\ \vdots & \ddots & \ddots & 0 \\ 0 & \dots & 0 & \gamma_n^2 \end{bmatrix} \quad (3.38)$$

It should be noted that element inside the matrix $\boldsymbol{\gamma}$, 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 $\mathbf{Z}$. β_i is the phase constant (in rad/m) and it is related to the admittance matrix $\mathbf{Y}$.

From (3.37), the general solution for the mode current is

$$\hat{\mathbf{I}}_m(z) = e^{-\gamma z} \hat{\mathbf{I}}_m^+ - e^{\gamma z} \hat{\mathbf{I}}_m^- \quad (3.39)$$

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

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

Similarly, by substituting (3.36) into (3.34a), the decoupled second-order differential equation for the mode voltage can be obtained

$$\begin{aligned} \frac{d^2}{dz^2} \hat{\mathbf{V}}_m(z) &= \mathbf{T}^{-1} \mathbf{Z} \mathbf{Y} \mathbf{T} \hat{\mathbf{V}}_m(z) = \mathbf{T}^T \mathbf{Z} \mathbf{Y} (\mathbf{T}^T)^{-1} \hat{\mathbf{V}}_m(z) \\ &= \gamma^2 \hat{\mathbf{V}}_m(z) \end{aligned} \quad (3.41)$$

The general solution for the mode voltage is

$$\hat{\mathbf{V}}_m(z) = e^{-\gamma z} \hat{\mathbf{V}}_m^+ - e^{\gamma z} \hat{\mathbf{V}}_m^- \quad (3.42)$$

The actual voltage can be obtained by multiplying this result by the transformation, $\mathbf{T}_V=(\mathbf{T}_I^{-1})=(\mathbf{T}^{-1})^T$ to give

$$\hat{\mathbf{V}}(z) = (\mathbf{T}^{-1})(e^{-\gamma z} \hat{\mathbf{V}}_m^+ - e^{\gamma z} \hat{\mathbf{V}}_m^-) \quad (3.43)$$

The solutions in (3.40) and (3.43) are actually related. To determine this relation, (3.40) is substituted into the second MTL equation (3.27b),

$$\begin{aligned} \hat{\mathbf{V}}(z) &= -\mathbf{Y}^{-1} \frac{d}{dz} \hat{\mathbf{I}}(z) = \mathbf{Y}^{-1} \mathbf{T} \gamma (e^{-\gamma z} \hat{\mathbf{I}}_m^+ + e^{\gamma z} \hat{\mathbf{I}}_m^-) \\ &= (\mathbf{Y}^{-1} \mathbf{T} \gamma \mathbf{T}^{-1}) \mathbf{T} (e^{-\gamma z} \hat{\mathbf{I}}_m^+ + e^{\gamma z} \hat{\mathbf{I}}_m^-) \\ &= \mathbf{Z}_0 \mathbf{T} (e^{-\gamma z} \hat{\mathbf{I}}_m^+ + e^{\gamma z} \hat{\mathbf{I}}_m^-) \end{aligned} \quad (3.44)$$

where the characteristic impedance matrix can be defined as [56]

$$\begin{aligned}
 \mathbf{Z}_0 &= \mathbf{Y}^{-1} \mathbf{T} \boldsymbol{\gamma} \mathbf{T}^{-1} \\
 &= \mathbf{Z} \mathbf{T} \boldsymbol{\gamma} \mathbf{T}^{-1}
 \end{aligned}
 \tag{3.45}$$

Note that the propagation constant matrix $\boldsymbol{\gamma}$ and characteristic impedance $\mathbf{Z}_0$ are characteristic properties of the MTL whether or not the line is infinitely long. They depend on the primary line parameters matrices $\mathbf{R}$, $\mathbf{L}$, $\mathbf{G}$, $\mathbf{C}$ and ω but not the length of the line. Normally the similarity transmission matrix $\mathbf{T}$ is chosen as the combination of eigen vectors of the product $\mathbf{Z} \cdot \mathbf{Y}$ [54] so that $\boldsymbol{\gamma}$ and $\mathbf{Z}_0$ can be easily determined.

In the next chapter, the determination of the per-unit-parameter matrices for $\mathbf{R}$, $\mathbf{L}$, $\mathbf{G}$, $\mathbf{C}$ will be discussed so that $\boldsymbol{\gamma}$ and $\mathbf{Z}_0$ for the power line structure as shown in Figure 3.1 can be determined.

Chapter 4

DETERMINATION OF LINE PARAMETERS

The two line parameters, the characteristic impedance Z_0 and the propagation constant matrix γ , serve as the model parameters in this research. In this chapter, the derivation of these two parameters is discussed. In order to do this, from (3.10) and (3.22), the four primary line parameters, R , L , G and C are determined first.

The power line structure to be studied is shown in Figure 3.1. The communication signal will be transmitted between live and neutral cables. It is for sure that differential mode transmission will be used since the common mode transmission will result in much serious EMI/EMC problem. Figure 4.1 illustrates the MTL representation for the power line.

Figure 4.1. MTL Representation of Power Line

The earth cable and the metal conduit are both grounded. However, the earth cable is only grounded at certain points in the power network, in the high frequency transmission, this can not be regarded as perfect grounding. The metal conduit has direct contact with the wall or the earth through out its length and it can be approximated as a perfect grounding plane. As a result, in this research, the power line is represented by a (2+1) MTL. With the neutral cable as the reference conductor, the live cable and the earth cable can be regarded

as conductor 1 and 2 respectively. Our interests are on the characteristic impedance and the propagation constant between the live and neutral cables (between conductor 1 and 0). In the following sections, the details of the derivations for the four primary line parameters are described.

4.1 DETERMINATION OF PRIMARY LINE PARAMETERS

4.1.1 Resistance

When AC current flows in a conductor, the self-inductance within the conductor causes more current to flow near to the outer surface of the wire instead of towards the center. This phenomenon is called the skin effect [53]. This effect causes an increase in the effective resistance of the cable and it worsens as the current frequency increases. Although the current flow is still distributed throughout the cross-section of the cable, when calculating the resistance, it is normal to assume that all the current flows within the “skin depth” of the cable. The skin depth (δ) is a function of frequency and can be calculated using the equation below,

$$\delta = \sqrt{\frac{1}{\pi f \mu_c \sigma_c}} \quad (4.1)$$

where σ_c and μ_c are the conductivity and permeability of the conductor respectively. So, for a round cross-section solid core conductor, the resistance can be obtained,

$$R_{solid} = \frac{1}{\pi a \delta \sigma_c} \text{ (}\Omega/\text{m)} \quad (4.2)$$

where a is the radius of the solid conductor.

If the conductor has a solid composition then by using the above method for calculating the skin depth and knowing the radius of the conductor, the effective cross sectional area of current flow can be determined. However, for power line, the cable conductor is stranded and hence, the area of the current flow is again reduced because of the gaps at the circumference of the conductor strands [55]. This is shown in Figure 4.2. δ in the diagram stands for the skin depth, and the shaded area indicates the area of current flow.

Figure 4.2. Area of Current Flow on Outer Surface of Stranded Core

To determine the resistance under this type of current flow, the ratio of the effective current flow area of the solid-core conductor and stranded core conductor will be determined first. Specifically, our purpose is to determine the ratio of the shade area to the area of the ring shape in Figure 4.2. The current flow area for solid conductor can be given by

$$\begin{aligned} S_{solid} &= \pi \cdot r_{cable}^2 - \pi \cdot (r_{cable} - \delta)^2 \\ &= \pi \cdot (2r_{cable}\delta - \delta^2) \end{aligned} \quad (4.3)$$

where r_{cable} is the radius of the power cable conductor.

Figure 4.3. Area of Current Flow on One Surface Stranded Wire

The determination of the shaded area is not so straightforward. The shaded area in Figure 4.3 shows the area of current flow on one stranded wire. The following geometry derivation steps are used to determine this area, which can be regarded as the subtraction

of the sector $\widehat{OGJF}$ by the triangle OGF and the crescent shape $\widehat{HGIF}$. According to the Law of Cosines, from Figure 4.3, it can be obtained that

$$OG^2 = OA^2 + GA^2 - 2 \cdot OA \cdot GA \cdot \cos(\angle GAO)$$

From this,

$$\begin{aligned} \angle GAO &= \cos^{-1} \left(\frac{OA^2 + GA^2 - OG^2}{2 \cdot OA \cdot GA} \right) \\ &= \cos^{-1} \left[\frac{(r_{\text{cable}} - r_{\text{wire}})^2 + (r_{\text{cable}} - \delta)^2 - r_{\text{wire}}^2}{2 \cdot (r_{\text{cable}} - r_{\text{wire}}) \cdot (r_{\text{cable}} - \delta)} \right] \\ &= \cos^{-1} \left[\frac{2r_{\text{cable}}^2 - 2r_{\text{cable}}r_{\text{wire}} - 2r_{\text{cable}}\delta + \delta^2}{2 \cdot (r_{\text{cable}}^2 - r_{\text{cable}}r_{\text{wire}} - r_{\text{cable}}\delta + r_{\text{wire}}\delta)} \right] \end{aligned} \quad (4.4)$$

$$\angle GAF = 2 \cdot \angle GAO \quad (4.5)$$

From (4.4) and (4.5), the area of triangle AGF

$$\begin{aligned} S_{AGF} &= \frac{1}{2} AG \cdot AF \cdot \sin(\angle GAF) \\ &= \frac{1}{2} (r_{\text{cable}} - \delta)^2 \sin \left\{ 2 \cdot \cos^{-1} \left[\frac{2r_{\text{cable}}^2 - 2r_{\text{cable}}r_{\text{wire}} - 2r_{\text{cable}}\delta + \delta^2}{2 \cdot (r_{\text{cable}}^2 - r_{\text{cable}}r_{\text{wire}} - r_{\text{cable}}\delta + r_{\text{wire}}\delta)} \right] \right\} \end{aligned} \quad (4.6)$$

The area of the sector $\widehat{AGIF}$ can also be determined by

$$\begin{aligned} S_{\widehat{AGIF}} &= \frac{1}{2} AG \cdot AF \cdot \angle GAF \\ &= (r_{\text{cable}} - \delta)^2 \cos^{-1} \left[\frac{2r_{\text{cable}}^2 - 2r_{\text{cable}}r_{\text{wire}} - 2r_{\text{cable}}\delta + \delta^2}{2 \cdot (r_{\text{cable}}^2 - r_{\text{cable}}r_{\text{wire}} - r_{\text{cable}}\delta + r_{\text{wire}}\delta)} \right] \end{aligned} \quad (4.7)$$

From (4.6) and (4.7), the area of the crescent shape $\widehat{HGIF}$ can be obtained,

$$\begin{aligned} S_{\widehat{HGIF}} &= S_{\widehat{AGIF}} - S_{AGF} \\ &= (r_{\text{cable}} - \delta)^2 \cos^{-1} \left[\frac{2r_{\text{cable}}^2 - 2r_{\text{cable}}r_{\text{wire}} - 2r_{\text{cable}}\delta + \delta^2}{2 \cdot (r_{\text{cable}}^2 - r_{\text{cable}}r_{\text{wire}} - r_{\text{cable}}\delta + r_{\text{wire}}\delta)} \right] \\ &\quad - \frac{1}{2} (r_{\text{cable}} - \delta)^2 \sin \left\{ 2 \cdot \cos^{-1} \left[\frac{2r_{\text{cable}}^2 - 2r_{\text{cable}}r_{\text{wire}} - 2r_{\text{cable}}\delta + \delta^2}{2 \cdot (r_{\text{cable}}^2 - r_{\text{cable}}r_{\text{wire}} - r_{\text{cable}}\delta + r_{\text{wire}}\delta)} \right] \right\} \end{aligned} \quad (4.8)$$

To determine the area of triangle OGF ,

$$GH = AG \cdot \sin(\angle GAO) \quad (4.9)$$

From (4.4) and (4.9),

$$\begin{aligned} \angle GOF &= 2 \cdot \angle GOH = 2 \sin^{-1} \left(\frac{GH}{GO} \right) \\ &= 2 \sin^{-1} \left\{ \frac{(r_{cable} - \delta) \cdot \sin \left\{ \cos^{-1} \left[\frac{2r_{cable}^2 - 2r_{cable}r_{wire} - 2r_{cable}\delta + \delta^2}{2 \cdot (r_{cable}^2 - r_{cable}r_{wire} - r_{cable}\delta + r_{wire}\delta)} \right] \right\}}{r_{wire}} \right\} \end{aligned} \quad (4.10)$$

From (4.10), the area of the triangle GOF can be determined,

$$\begin{aligned} S_{GOF} &= \frac{1}{2} OG \cdot OF \cdot \sin(\angle GOF) = \\ &= \frac{1}{2} r_{wire}^2 \sin \left\{ 2 \sin^{-1} \left\{ \frac{(r_{cable} - \delta) \cdot \sin \left\{ \cos^{-1} \left[\frac{2r_{cable}^2 - 2r_{cable}r_{wire} - 2r_{cable}\delta + \delta^2}{2 \cdot (r_{cable}^2 - r_{cable}r_{wire} - r_{cable}\delta + r_{wire}\delta)} \right] \right\}}{r_{wire}} \right\} \right\} \end{aligned} \quad (4.11)$$

and the area of the sector $\widehat{OGJF}$ can also be obtained,

$$\begin{aligned} S_{\widehat{OGJF}} &= \frac{1}{2} OG \cdot OF \cdot \angle GOF = \\ &= r_{wire}^2 \sin^{-1} \left\{ \frac{(r_{cable} - \delta) \cdot \sin \left\{ \cos^{-1} \left[\frac{2r_{cable}^2 - 2r_{cable}r_{wire} - 2r_{cable}\delta + \delta^2}{2 \cdot (r_{cable}^2 - r_{cable}r_{wire} - r_{cable}\delta + r_{wire}\delta)} \right] \right\}}{r_{wire}} \right\} \end{aligned} \quad (4.12)$$

From (4.8), (4.11), and (4.12), the shaded current-flow area can be determined by

$$\begin{aligned} S_{shade} &= S_{\widehat{OGJF}} - S_{GOF} - S_{\widehat{HGI}} = \\ &= r_{wire}^2 \sin^{-1} \left\{ \frac{(r_{cable} - \delta) \cdot \sin \left\{ \cos^{-1} \left[\frac{2r_{cable}^2 - 2r_{cable}r_{wire} - 2r_{cable}\delta + \delta^2}{2 \cdot (r_{cable}^2 - r_{cable}r_{wire} - r_{cable}\delta + r_{wire}\delta)} \right] \right\}}{r_{wire}} \right\} - \end{aligned}$$

$$\begin{aligned}
& \frac{1}{2} r_{\text{wire}}^2 \sin \left\{ 2 \sin^{-1} \left\{ \frac{(r_{\text{cable}} - \delta) \cdot \sin \left\{ \cos^{-1} \left[\frac{2r_{\text{cable}}^2 - 2r_{\text{cable}}r_{\text{wire}} - 2r_{\text{cable}}\delta + \delta^2}{2 \cdot (r_{\text{cable}}^2 - r_{\text{cable}}r_{\text{wire}} - r_{\text{cable}}\delta + r_{\text{wire}}\delta)} \right] \right\}}{r_{\text{wire}}} \right\} \right\} - \\
& (r_{\text{cable}} - \delta)^2 \cos^{-1} \left[\frac{2r_{\text{cable}}^2 - 2r_{\text{cable}}r_{\text{wire}} - 2r_{\text{cable}}\delta + \delta^2}{2 \cdot (r_{\text{cable}}^2 - r_{\text{cable}}r_{\text{wire}} - r_{\text{cable}}\delta + r_{\text{wire}}\delta)} \right] + \\
& \frac{1}{2} (r_{\text{cable}} - \delta)^2 \sin \left\{ 2 \cdot \cos^{-1} \left[\frac{2r_{\text{cable}}^2 - 2r_{\text{cable}}r_{\text{wire}} - 2r_{\text{cable}}\delta + \delta^2}{2 \cdot (r_{\text{cable}}^2 - r_{\text{cable}}r_{\text{wire}} - r_{\text{cable}}\delta + r_{\text{wire}}\delta)} \right] \right\}
\end{aligned} \tag{4.13}$$

The total area of current flow on the outer surface of the stranded cable can be given by

$$S_{\text{stranded}} = N \cdot S_{\text{shade}} \tag{4.14}$$

where N is the number of the surface wires of the stranded conductor. Comparing with solid core conductor, the resistance correction factor for stranded core conductor can be given by

$$X_{\text{cor}} = \frac{S_{\text{stranded}}}{S_{\text{solid}}} \tag{4.15}$$

With this correction factor, the final resistance for the stranded cable is

$$\begin{aligned}
R_{\text{cable}} &= X_{\text{cor}} \cdot R_{\text{solid}} \\
&= X_{\text{cor}} \frac{1}{\pi r_{\text{cable}} \delta_{\text{cable}} \sigma_{\text{cable}}}
\end{aligned} \tag{4.16}$$

where δ_{cable} and σ_{cable} are the skin depth and the conductivity of the cable respectively.

From (4.16) and (3.12), with the neutral cable as the reference conductor, the per-unit-length resistance matrix for power line can be written as,

$$\mathbf{R} = \begin{bmatrix} 2R_{\text{cable}} & R_{\text{cable}} \\ R_{\text{cable}} & 2R_{\text{cable}} \end{bmatrix} \tag{4.17}$$

4.1.2 Inductance

To derive the per-unit-length inductance and capacitance matrices of the power line, knowledge of the static EM analysis needs to be used. Basically, the determination of inductance is based on Magnetostatics.

Before deriving the inductances for the power line, let's firstly consider the magnetic flux that penetrates a surface s that is parallel to a cable with current I as shown in Figure 4.4.

Figure 4.4. 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 of M_1 and M_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

$$\begin{aligned}\psi_m &= \oiint_s \vec{B}_t \cdot d\vec{s} = \Delta z \int_{M_1}^{M_2} \frac{\mu I}{2\pi r} dr \\ &= \Delta z \frac{\mu I}{2\pi} \ln\left(\frac{M_2}{M_1}\right)\end{aligned}\quad (4.18)$$

so that the per-unit-length magnetic flux is

$$\psi = \frac{\psi_m}{\Delta z} = \frac{\mu I}{2\pi} \ln\left(\frac{M_2}{M_1}\right) \text{ (Wb/m)} \quad (4.19)$$

To derive the per-unit-length inductances matrix for power line, the Method of Images is used. This method can be applied based on the previous assumption that the metal conduit is a perfectly conducting shield. Figure 4.5 shows how the method of images is applied to calculate the inductances.

Figure 4.5. Illustration of Calculating the Inductance by Method of Images

According to this method, if a directional current flows on one side of the conducting plane, then it will have an image current with opposite direction on the other side of the conducting plane. From [58], the distance between the center of the conducting shield and the center of the image conductor is given by $l_m = \frac{r_c^2}{l_d}$. From (4.19), the total per-unit-length magnetic flux between conductor i and its corresponding image can be obtained,

$$\begin{aligned} \psi_{ii} &= \frac{\mu}{2\pi I} \ln \left(\frac{r_c - l_d}{r_w} \right) + \frac{\mu}{2\pi I} \ln \left(\frac{l_m - l_d}{l_m - r_c} \right) \\ &= \frac{\mu}{2\pi I} \ln \left(\frac{r_c - l_d}{r_w} \right) + \frac{\mu}{2\pi I} \ln \left(\frac{r_c^2 / l_d - l_d}{r_c^2 / l_d - r_c} \right) \\ &= \frac{\mu}{2\pi I} \ln \left(\frac{r_c^2 - l_d^2}{r_c \cdot r_w} \right) \end{aligned} \quad (4.20)$$

and the self-inductance for conductor i is

$$L_{ii} = \frac{\Psi_{ii}}{I} = \frac{\mu}{2\pi} \ln \left(\frac{r_c^2 - l_d^2}{r_c \cdot r_w} \right) \text{ (H/m)} \quad (4.21a)$$

However, again the skin effects need to be considered. The skin effects not only cause an increase in the wire resistance but also a decrease in self-inductance. According to [79], (4.21a) can be corrected as,

$$\begin{aligned} L_{ii} &= \frac{1}{0.315 \times 0.53 \times \frac{r_c}{\delta}} \frac{\mu}{2\pi} \ln \left(\frac{r_c^2 - l_d^2}{r_c \cdot r_w} \right) \\ &\approx \frac{\delta \cdot \mu}{0.334 \cdot r_c \cdot \pi} \ln \left(\frac{r_c^2 - l_d^2}{r_c \cdot r_w} \right) \end{aligned} \quad (4.21b)$$

where δ is the skin depth from (4.1).

From (4.19), the total per-unit-length magnetic flux between conductor i and conductor j can be summed as,

$$\begin{aligned} \Psi_{ij} &= \frac{\mu}{2\pi I} \ln \left(\frac{s_1}{s_{ij}} \right) + \frac{\mu}{2\pi I} \ln \left(\frac{\hat{s}}{s_2} \right) \\ &= \frac{\mu}{2\pi I} \ln \left(\frac{1}{r_c \cdot l_d} \cdot \sqrt{\frac{l_d^4 + r_c^4 - 2l_d^2 r_c^2 \cos \theta_{ij}}{2 \sin \theta_{ij}}} \right) \end{aligned} \quad (4.22)$$

and the mutual inductance between conductor i and conductor j can be obtained,

$$L_{ij} = \frac{\Psi_{ij}}{I} = \frac{\mu}{2\pi} \ln \left(\frac{1}{r_c \cdot l_d} \cdot \sqrt{\frac{l_d^4 + r_c^4 - 2l_d^2 r_c^2 \cos \theta_{ij}}{2 \sin \theta_{ij}}} \right) \text{ (H/m)} \quad (4.23)$$

From (4.21b), (4.23) and (3.13) derived in previous chapter, the per-unit-length inductance matrix for the power line shown in Figure 4.1 can be written as

$$\mathbf{L} = \begin{bmatrix} L_{11} & L_{12} \\ L_{21} & L_{22} \end{bmatrix} \quad (4.24)$$

Since $L_{11}=L_{22}$, let them equal to L_{int} , and $L_{12}=L_{21}$, let them equal to L_{mut} , so (4.24) can be rewritten as

$$\mathbf{L} = \begin{bmatrix} L_{int} & L_{mut} \\ L_{mut} & L_{int} \end{bmatrix} \quad (4.25)$$

4.1.3 Capacitance

The derivation of the per-unit-length capacitance matrix for the power line is based on electrostatics. The equivalent coupling diagram inside the power line is shown in Figure 4.6. C_{pair} is the per-unit-length capacitance between paired cables without considering other conductors around them. $C_{conduit}$ is the per-unit-length capacitance between one cable and the metal conduit.

Figure 4.6. Equivalent Capacitance Diagram for Power Line (L: live N: neutral E: earth)

In the following sections, the derivations for C_{cable} and $C_{conduit}$ will be described.

4.1.3.1 Capacitance between Cables

Figure 4.7. Cross Section of Two Parallel Cables and the Corresponding Equivalent Line Charges

C_{pair} can be regarded as the capacitance between two parallel cables. Referring to the cross-sectional view as shown in Figure 4.7, 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 $(D - 2d_i) = d - d_i$ [53]. The potential difference between the two wires is that between any two points on the respective cables.

Let subscripts 1 and 2 denote the cables surrounding the equivalent line charges $+q$ and $-q$ respectively, then by using the path integration, the potential on cable 1 and cable 2,

$$V_2 = \frac{q}{2\pi\epsilon_0} \ln \frac{a}{d} \quad (4.26)$$

$$V_1 = -\frac{q}{2\pi\epsilon_0} \ln \frac{a}{d} \quad (4.27)$$

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

$$C = \frac{q}{V_1 - V_2} = \frac{\pi\epsilon_0}{\ln\left(\frac{d}{a}\right)} \quad (4.28)$$

According to cable geometry, d can be determined,

$$d = \frac{1}{2} \left(D + \sqrt{D^2 - 4a^2} \right) \quad (4.29)$$

With (4.28) and (4.29),

$$C_{pair} = \frac{\pi\epsilon}{\ln\left[\left(D/2a\right) + \sqrt{\left(D/2a\right)^2 - 1}\right]} \text{ (F/m)} \quad (4.30)$$

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

4.1.3.2 Capacitance between Cable and Conduit

The determination for $C_{conduit}$ is more complicated since the cable and the conduit are in the eccentric coaxial position. This kind of capacitance can be solved by conformal mapping method or perturbation method. In this research, a simpler analytic method based on segmentation is used.

As shown in Figure 4.8, taking the center of the inner conductor as the origin, the radial axes are used to segment the eccentric coaxial conductors into N sectors. When $N \rightarrow +\infty$, each sector can be approximated as one segment in the coaxial cylindrical conductors. The capacitance for the coaxial cylindrical conductor is given in [53],

$$C_{coaxial} = \frac{2\pi\epsilon}{\ln\left(\frac{b}{a}\right)} \text{ (F/m)} \quad (4.31)$$

where b is the inner radius of the outer conductor.

Figure 4.8. Segmentation of the Eccentric Coaxial Cylindrical Conductors

Based on the segmentation method, the capacitance between the cable conductor and metal conduit as shown in Figure 4.8 can be taken as the average capacitance of all segments,

$$C_{conduit} = \lim_{N \rightarrow \infty} \sum_{k=1}^N \frac{1}{N} \cdot \frac{2\pi\epsilon_k}{\ln\left(\frac{b_k}{a}\right)} \text{ (F/m)} \quad (4.32)$$

where ϵ_k and b_k are the permittivity of the dielectric material and the inner radius of the metal conduit for sector k .

From Figure 4.8, between any pair of cables, the high frequency communication signals are not only coupled by C_{pair} , but they are also coupled from one cable to the metal conduit and then coupled from conduit to another cable. Hence the total capacitance between any pair of cables can be summarized as

$$C_{cable} = C_{pair} + \frac{1}{2}C_{conduit} \quad (4.33)$$

From (4.33) and (3.24), the per-unit-length capacitance matrix for power line can be written as

$$\mathbf{C} = \begin{bmatrix} 2C_{cable} & -C_{cable} \\ -C_{cable} & 2C_{cable} \end{bmatrix} \quad (4.34)$$

4.1.4 Conductance

From [53], if the medium has the same space dependence or if the medium is homogeneous, the following equation holds,

$$\frac{C}{G} = \frac{\varepsilon}{\sigma} \quad \Rightarrow \quad G = \frac{\sigma \cdot C}{\varepsilon} \text{ (S/m)} \quad (4.35)$$

where σ is the conductivity of the dielectric material and G is the cable conductance.

For the house power cables as shown in Figure 4.1, the dielectric material, either between the cable conductors or between cable conductor and conduit, is inhomogeneous in both space (due to the round shape of the cable conductor) and contents (mixture of insulation and air). But since the cables are of close proximity to each other, the thickness of the insulation is comparable with that of the air between the conductors. In this research, the dielectric is assumed to be just a mixed content material and the effects of inhomogeneous in space are neglected to keep the model manageable. Based on this assumption and from (4.35),

$$\mathbf{G} = \frac{\sigma}{\varepsilon} \mathbf{C} = \frac{\sigma}{\varepsilon} \begin{bmatrix} 2C_{cable} & -C_{cable} \\ -C_{cable} & 2C_{cable} \end{bmatrix} \quad (4.36)$$

4.2 VERIFICATION OF LINE PARAMETERS

After deriving all the primary line parameter matrices, (3.38) and (3.45) can be applied to determine the characteristic impedance Z_0 and the propagation constant γ between live and neutral cables (conductor 1 and 0). With the line parameters, the communication channel

formed by live and neutral cables can be regarded as a two-port network. In this research, some measurements have been conducted to verify the derived line parameters between the live and neutral cables. From the transmission line theory, the input impedance looking into a line of length l and with a termination load of Z_L is

$$Z_{in} = Z_0 \frac{Z_L + Z_0 \tanh(\gamma \cdot l)}{Z_0 + Z_L \tanh(\gamma \cdot l)} \quad (\Omega) \quad (4.37)$$

If the load terminal is short-circuited, i.e. $Z_L=0$, (4.37) becomes

$$Z_{is} = Z_0 \tanh(\gamma \cdot l) \quad (\Omega) \quad (4.38)$$

Similarly, if the load terminal is open-circuited, i.e. $Z_L \rightarrow \infty$, (4.37) becomes

$$Z_{io} = Z_0 \cot(\gamma \cdot l) \quad (\Omega) \quad (4.39)$$

From (4.38) and (4.39), it can be derived that

$$Z_0 = \sqrt{Z_{is} \cdot Z_{io}} \quad (4.40)$$

$$\gamma = \frac{1}{l} \tanh^{-1} \sqrt{\frac{Z_{is}}{Z_{io}}} \quad (4.41)$$

Figure 4.9 shows the measurement setup for the line parameter verification. The input impedances are measured by using the Agilent analyzer combined with the S-parameter test set. The data are acquired into the computer through the General Purpose Interface Board (GPIB) interface. Z_{is} and Z_{io} are measured by connecting the Short Circuit (SC) and Open Circuit (OC) calibration kits to the end of the testing cable respectively. The measured power line's structure is similar to the one shown in Figure 4.1. The inner radius of the metal conduit is 15mm and the cross-sectional area of the cable conductor is 4.0mm².

It should be noted that when the cable length is equal to multiples of the quarter wavelength of the operating signal, then Z_{is} and Z_{io} will be either zero or infinity. This quarter wavelength effect can be easily viewed on the Smith's chart. At a certain frequency, the input impedance of a short termination (or open termination) will become open circuit (or short circuit) when the cable length is equal to odd multiples of the quarter wavelength and will become short circuit (or open circuit) when the cable length is equal to even multiples of the quarter wavelength. In such cases, measurements will be

erroneous and the results should not be considered. Because of this, in the verification measurements, several lengths of power lines are used and results affected by the above phenomenon are ignored.

Figure 4.9. Measurement Setup for Line Parameters Verification

Figures 4.10 and 4.11 compare the measured and the derived characteristic impedance (Z_0), attenuation constant (α) and group delay (τ), which represents the time needed for the wave-packet envelope (of a group of frequencies) to propagate a distance of 1m,

$$\tau = \frac{d\beta}{d\omega} \text{ (s/m)} \tag{4.42}$$

Figure 4.10. Comparison of the Measured and Derived Characteristic Impedance

Figure 4.11. Comparison of the Measured and Derived Attenuation Constant and Group Delay

From the comparisons, it can be seen that the derived line parameters match closely to the measured values most of the time. This confirms the validity of this modeling approach. As seen from the figures, the measured wave impedance and the group delay (which is a reflection of the group velocity) are roughly constant in the frequency range of interest and are frequency independent. These observations have confirmed that our previous assumption of TEM mode propagation in the power line channel is correct.

Z_0 and γ provide a very practical representation of the lines under different physical dimensions and material properties. The following network modeling chapter will show how to use these two line parameters to model the channel transfer function and network input impedance.

Chapter 5

TREE TOPOLOGY NETWORK MODELING

Since the majority of power distribution networks in Singapore are connected in radial structure, there is a strong need to study the characteristics of the PLC channel in tree-topology power networks. With the help of the characteristic impedance and propagation constant derived in the last chapter, the focus of this chapter is on network modeling. This includes transfer function modeling and input impedance modeling for the tree-topology network. The time domain modeling approach, which uses the echo algorithm, is discussed first. Then the frequency domain approach based on scattering parameters is explained. Using the hardware described in Appendix A.1, the impedances for some household appliances in the PLC frequency are measured to obtain the network transfer function and input impedance for verifying the derived model. From comparison with measurements, it can be seen that the frequency domain approach provides a more accurate and convenient result compared to the time domain approach. Since the network model is constructed based on the parameters of network configurations, so that in the last section of this chapter, error sensitivities of the derived frequency-domain model to these parameters are studied.

5.1 TIME DOMAIN MODELING APPROACH

Among the three types of attenuations introduced in Chapter 2, line attenuation always exists because the heat loss and the radiation loss on the power line cannot be avoided. The attenuation constant determined in the previous chapter is an indication of the line attenuation. However, it should be noted that by using the MTL model as described in Chapters 3 and 4, the derived attenuation constant only considers the heat loss on the cable. To incorporate the radiation loss, the complex charge distribution and EM field need to be closely analyzed. Anyway, if only the home network is considered, whose maximum cable length is normally less than 50m, the line attenuation will not be dominant. Besides line attenuation, impedance mismatch also results in signal attenuation. Fortunately, this

mismatch attenuation can be easily reduced through proper matching of the PLC transceivers to the network. The most unpredictable attenuation is the multi-path attenuation. When PLC waves propagate in the channel, at the branching points, they will be reflected and branched. And at the terminal points, they will be absorbed and reflected. This action will repeat until all the waves have been attenuated or absorbed.

It is observed that reflections from the impedance discontinuities and branches in the PLC channel result in echoes of the transmitted signal. Hence, the power line channel can be regarded as a multi-path environment. Each transmitted signal reaches the receiver not only on a direct path, but also delayed and in most cases attenuated portions are present at the receivers. An echo-based model can therefore be developed to represent this physical characteristic of the network.

Figure 5.1 shows a channel model with N echoes, to which the impulse response of the channel can be written as [70],

$$h(t) = \sum_{i=1}^N g_i \delta(t - \tau_i) \quad (5.1)$$

where g_i and τ_i are the complex attenuation factor and delay for path i , $\delta(\bullet)$ is the delta-sigma function.

Figure 5.1. Basic Structure of Echo Model

By applying Fourier Transform to (5.1), the transfer function of the channel is obtained,

$$H(f) = \sum_{i=1}^N g_i e^{-j2\pi f \tau_i} \quad (5.2)$$

For a N-branch power line network, if the backward reflections from the branches are ignored, the total transfer function of this channel can be approximated as a multiplication of the transfer functions of the N single-branch sub networks, as shown in Figure 5.2.

Figure 5.2. Approximation of Transfer Function for N-branched Network

With the above approximation, the transfer function of a simple single-branched network can be analyzed as shown in Figure 5.3.

Figure 5.3. Signal Propagation on a Single-branched Network

Path No.	Signal Path	Attenuation Factor (g_i)	Path Length (d_i)
1	$1 \rightarrow 2 \rightarrow 3$	T_{PJ}	$l_{12} + l_{23}$
2	$1 \rightarrow 2 \rightarrow 4 \rightarrow 2 \rightarrow 3$	$T_{PJ} \cdot \Gamma_{BT} \cdot T_{BJ}$	$l_{12} + 2l_{24}$
N	$1 \rightarrow 2 \rightarrow (4 \rightarrow 2)^{N-1} \rightarrow 3$	$T_{PJ} \cdot \Gamma_{BT} \cdot (\Gamma_{BT} \cdot \Gamma_{BJ})^{N-2} \cdot T_{BJ}$	$l_{12} + 2(N-1) \cdot l_{24} + l_{23}$

Table 5.1. Summary of Signal Paths for Deriving Transfer Function

This network is defined by terminals 1, 3, 4 and joint 2. It consists of three segments with lengths l_{12} , l_{23} , and l_{24} , having characteristic impedances of Z_{12} , Z_{23} , and Z_{24} respectively. Γ_{BT} is the reflection coefficient from the branch terminal. Γ_{PJ} and Γ_{BJ} are the reflection coefficients from path to joint and from branch to joint respectively, and T_{PJ} and T_{BJ} are the respective transmission coefficients. The signal propagation paths are summarized in Table 5.1.

The reflection and transmission coefficients can be obtained by the following equations,

$$\Gamma_{BT} = \frac{Z_4 - Z_{24}}{Z_4 + Z_{24}} \quad (5.3)$$

$$\Gamma_{PJ} = \frac{Z_{23} // Z_{24} - Z_{12}}{Z_{23} // Z_{24} + Z_{12}} \quad (5.4)$$

$$\Gamma_{BJ} = \frac{Z_{12} // Z_{23} - Z_{24}}{Z_{12} // Z_{23} + Z_{24}} \quad (5.5)$$

$$T_{PJ} = 1 + \Gamma_{PJ} \quad (5.6)$$

$$T_{BJ} = 1 + \Gamma_{BJ} \quad (5.7)$$

where Z_4 is the load impedance at terminal 4. And the delay τ_i of a path will be related to the length of the path,

$$\tau_i = \frac{d_i}{v_p} \quad (5.8)$$

where v_p is the phase velocity of the wave ($v_p = \frac{\beta}{\omega}$). Using the above derivations and (5.2), the transfer functions for any tree topology network can be easily determined.

Although the time domain approach looks simple as summarized in Table 5.1, it is still based on a “big” assumption that all the signal paths generated by the backward reflections are ignored. In other words, it assumed that there will be no reflections from any of the branch i to previous $(i-1)$ branches and the transmitter. This assumption may be valid for those networks, whose signal paths generated by backward reflections are very small and can be ignored, but may not be valid for those whose backward reflections are significant. Another disadvantage of the time domain approach is that it can not really be used to determine another important parameter, which is the input impedance of the network. In

the following section, the frequency domain approach is derived, and it is shown that both of these disadvantages in the time domain approach can be overcome in the frequency domain approach.

5.2 FREQUENCY DOMAIN MODELING APPROACH

Frequency domain modeling approach is suggested because it is very hard to use the time domain modeling approach to account for all the reflected and delayed paths in the power network. As explained in Chapter 2, the Scattering parameters (S-parameters) is used to characterize the channel. However, obtaining the Scattering matrix (S-matrix) of the whole tree network is quite complex and hence in this research, similar to the time domain approach, the N-branch network is divided into a group of N cascaded single-branch sub-networks as shown in Figure 5.4. The S-matrix for each single-branched sub-network is first derived and the S-matrix of the whole channel is then determined by using the chain scattering matrix method.

Figure 5.4. Simplified Representation of N-Branch Tree Topology Network

To analyze the S-parameters of a single-branched network, the following diagram shown in Figure 5.5 is used. In this derivation, the power line on the direct signal path (excluding the branches) is defined as the path line. Let l_1 be the cable length from the left end of the path line to the tap point, l_2 the cable length of the cable branched off from the tap point and l_3 the cable length from the tap point to the right end of the path line. Using transmission line theory, Z_{in1} , Z_{in2} , Z_{in} can be determined.

- Z_g : Source impedance (50Ω)
 Z_L : Load impedance at path end (50Ω)
 Z_b : Load impedance at branch end
 Z_{in1} : Input impedance of the network on the right of the tap
 Z_{in2} : Input impedance of the branched network
 Z_{in} : Input impedance of the single-branched network
 Γ_1 : Reflection coefficient from path end
 Γ_2 : Reflection coefficient from the tap point
 Line1: The path power line with line parameter (Z_0, γ)
 Line2: The branched power line with line parameter (Z'_0, γ')
 Line3: The transmission line with 50Ω characteristic impedance (negligible length)

Figure 5.5. Analysis Diagram for Single-Branched Network

$$Z_{in1} = Z_0 \cdot \frac{Z_L + Z_0 \cdot \tanh(\gamma \cdot l_3)}{Z_0 + Z_L \cdot \tanh(\gamma \cdot l_3)} \quad (\Omega) \quad (5.9)$$

$$Z_{in2} = Z'_0 \cdot \frac{Z_b + Z'_0 \cdot \tanh(\gamma' \cdot l_2)}{Z'_0 + Z_b \cdot \tanh(\gamma' \cdot l_2)} \quad (\Omega) \quad (5.10)$$

$$Z_{in} = Z_0 \cdot \frac{(Z_{in1} // Z_{in2}) + Z_0 \cdot \tanh(\gamma \cdot l_1)}{Z_0 + (Z_{in1} // Z_{in2}) \cdot \tanh(\gamma \cdot l_1)} \quad (5.11)$$

where γ and γ' are the propagation constants for the path line and for the branch line respectively and Z_0 is equal to 50Ω , which is the characteristic impedance of the negligible transmission line. With these input impedances, Γ_1 , and Γ_2 can be determined,

$$\Gamma_1 = \frac{Z_L - Z_0}{Z_L + Z_0} \quad (5.12)$$

$$\Gamma_2 = \frac{(Z_{in1} // Z_{in2}) - Z_0}{(Z_{in1} // Z_{in2}) + Z_0} \quad (5.13)$$

According to [58], the S_{11} and S_{21} of the network is given by

$$S_{11} = \left[\frac{b_1}{a_1} \right]_{a_2=0} = \frac{Z_m - Z_0}{Z_m + Z_0} \quad (5.14)$$

$$S_{21} = \left[\frac{b_2}{a_1} \right]_{a_2=0} = 2 \cdot \frac{V_3}{E_g} \quad (5.15)$$

Since it is not easy to obtain the ratio of E_g and V_3 in (5.15), S_{21} can be computed indirectly by

$$S_{21} = 2 \cdot \frac{V_3}{V_2} \cdot \frac{V_2}{V_1} \cdot \frac{V_1}{E_g} \quad (5.16)$$

where

$$\frac{V_1}{E_g} = \frac{Z_m}{Z_m + Z_0} \quad (5.17)$$

$$\begin{aligned} \frac{V_2}{V_1} &= \frac{V_2^+ (1 + \Gamma_2)}{V_1^+ (1 + \Gamma_2 \cdot e^{-2\gamma l_1})} \\ &= \frac{V_1^+ \cdot e^{-\gamma l_1}}{V_1^+} \cdot \frac{1 + \Gamma_2}{1 + \Gamma_2 \cdot e^{-2\gamma l_1}} \\ &= \frac{(1 + \Gamma_2) \cdot e^{-\gamma l_1}}{1 + \Gamma_2 \cdot e^{-2\gamma l_1}} \end{aligned} \quad (5.18)$$

Similarly,

$$\frac{V_3}{V_2} = \frac{(1 + \Gamma_1) \cdot e^{-\gamma l_3}}{1 + \Gamma_1 \cdot e^{-\gamma l_3}} \quad (5.19)$$

Then, from (5.16),

$$S_{21} = 2 \cdot \frac{Z_m}{Z_m + Z_0} \cdot \frac{(1 + \Gamma_2) \cdot e^{-\gamma l_1}}{1 + \Gamma_2 \cdot e^{-2\gamma l_1}} \cdot \frac{(1 + \Gamma_1) \cdot e^{-\gamma l_3}}{1 + \Gamma_1 \cdot e^{-\gamma l_3}} \quad (5.20)$$

From (5.14) and (5.20), S_{11} and S_{21} can be determined. Similar method can be used to determine S_{12} and S_{22} by simply swapping the locations of source and load.

After analyzing the single-branch network, the remaining task is the determination of the scattering matrix for a cascade of several single-branch networks. Using the microwave theory, there are generally two methods that can be used. The first method is to use the chain scattering matrix (or T-matrix), and the second is to use signal flow graph. For the sake of easier computation, the first method is used in this research.

In retrospect, the T-matrix for the whole network can be obtained by the product of all the T-matrices of the cascaded portions. The relation between the S-matrix and T-matrix is shown below [59],

$$\mathbf{T} = \begin{bmatrix} T_{11} & T_{12} \\ T_{21} & T_{22} \end{bmatrix} = \begin{bmatrix} \frac{1}{S_{21}} & -\frac{S_{22}}{S_{21}} \\ \frac{S_{11}}{S_{21}} & S_{12} - \frac{S_{11}S_{22}}{S_{21}} \end{bmatrix} \quad (5.21)$$

So the total T-matrix for the N-branched network will be

$$\mathbf{T}_{N\text{-branch}} = \prod_{k=1}^N (\mathbf{T}_k) \quad (5.22)$$

where $\mathbf{T}_k$ is the T-matrix for the k -th cascaded portion in the network. Finally, the S-matrix for the whole network can be obtained by using the following equation which convert the T-matrix back to S-matrix,

$$\mathbf{S} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} = \begin{bmatrix} \frac{T_{21}}{T_{11}} & T_{22} - \frac{T_{21}T_{12}}{T_{11}} \\ \frac{1}{T_{11}} & -\frac{T_{12}}{T_{11}} \end{bmatrix} \quad (5.23)$$

where the terms S_{21} and S_{12} give the network transfer functions from transmitter to receiver and from receiver to transmitter respectively; and the terms S_{11} and S_{22} can be used to calculate the input impedances at transmitter (Z_{in-tx}) and receiver (Z_{in-rx}) respectively. From (5.14),

$$Z_{in-tx} = Z_0 \cdot \frac{1 + S_{11}}{1 - S_{11}} \quad (5.24)$$

$$Z_{in-rx} = Z_0 \cdot \frac{1 + S_{22}}{1 - S_{22}} \quad (5.25)$$

The accuracy of both the time domain approach and frequency domain approach depend on how accurately the impedances of the connected loads are derived. Hence, before verifying the derived models, the next section describes how to determine the impedances of various household appliances. These appliances are important components in the low-voltage network. One major uncertainty in the PLC channel is caused by the on or off, plugged-in or unplugged states of these appliances. It is very hard to use theoretical model to characterize them because different appliances have different internal circuits and will appear as different impedances in the PLC frequency range. Hence, measurements were conducted to arrive at more accurate load impedances.

5.3 IMPEDANCE OF HOUSEHOLD APPLIANCES

In this section, a measurement scheme has been set up to characterize the impedances of several appliances with the help of designed signal coupler and line trap. The setup is illustrated in Figure 5.6.

Figure 5.6. Measurement Setup for Impedance Measurement

In the measurement, the impedance of the power network (Z_{net1}) is first measured when the appliance is not connected to the network. Then, the appliance is connected near the measured socket and the network impedance (Z_{net2}) is again measured. The impedance of the appliance with its supply cable (Z_{app1}) can be calculated as follow,

$$\frac{1}{Z_{net2}} = \frac{1}{Z_{net1}} + \frac{1}{Z_{app1}} \quad (5.26)$$

Rearranging (5.26),

$$Z_{app1} = \frac{Z_{net1} \cdot Z_{net2}}{Z_{net1} - Z_{net2}} \quad (5.27)$$

From these equations, the importance of using line trap can be realized. If the line trap is not used, then in some frequency range, the network impedance without the appliance (Z_{net1}) may be quite small compared to the appliance impedance in the same frequency range. Then in (5.26), Z_{net1} will dominate Z_{net2} but not Z_{app1} , which will increase the error sensitivity of the measured results.

It should be noted that our interest is on the impedance of the appliance itself, but not the appliance with its supply cable. Even for same type of appliances, those from different manufacturers will have different lengths of supply cables, and this will make them to appear as different impedance in the power network. In this research, the supply cable of the appliance is regarded as transmission line and by using impedance transformation, the impedance of the appliance itself (Z_{app1}) can be obtained,

$$Z_{app} = Z_{0_cable} \cdot \frac{Z_{app1} - Z_{0_cable} \cdot \tanh(\gamma_{cable} \cdot l)}{Z_{0_cable} - Z_{app1} \cdot \tanh(\gamma_{cable} \cdot l)} \quad (5.28)$$

Z_{0_cable} and γ_{cable} are the characteristic impedance and the propagation constant of the supply cable of the appliance respectively.

Figure 5.7 shows the amplitude of the impedances for several appliances when they are plugged in but not in operating condition. As the figure shows, most of the appliances' impedances decrease with the increase of frequency like dimmer, kettle, TV, fluorescent light, vacuum cleaner, microwave and fridge. This capacitive impedance property is reasonable since the appliances are in off state and the connection between the appliances and power line is open, which is acting as a capacitor. However, cooker and computer do show resonant properties having peaks in the frequency band.

Figure 5.7. Impedances for Some Particular Appliances in OFF State

Figure 5.8. Impedances for Some Particular Appliances in ON State

Figure 5.8 shows the impedance characteristics of these appliances in operating condition. Generally, it can be seen that most of the appliances' impedances in the on state will become smaller compared to those in the off state. The values of the impedances mainly

concentrate in the region of less than 500Ω . Among the measured appliances, dimmer, microwave, vacuum cleaner, fridge, and fluorescent light still show capacitive impedance property in this frequency band. Cooker, TV, kettle, and computer have obvious peaks in the frequency band and they behave as resonant circuits. The impedance of the electric fan shows a very small impedance ($<50\Omega$) for both on and off state in this frequency band.

5.4 VERIFICATION OF DERIVED MODELS

With the help of the measured appliances’ impedances above, the models derived using time-domain approach as described in Section 5.1 and frequency domain approach as detailed in Section 5.2 can be verified against measurements conducted on a real power network. Figure 5.9 shows a typical power network with three branches. Load 1, Load 2 and Load 3 in the network under test are a light dimmer, a vacuum cleaner and a desktop computer respectively, which are typical household appliances. All of them are in operating condition. Vacuum cleaner and computer are chosen since they show high resonant peaks in the frequency domain and these resonant peaks are useful for showing the presence of notches in the power line transfer function. The light dimmer is also included as it is readily available and can be easily incorporated into the test network. The lengths of the branches in the figure are the lengths of the supply cables of the loads. This network is only used to verify the transfer function and the input impedance models. It is important to note that the purpose of this network is not to mimic a universal power network. Hence, the obtained results are not general results as the network is assembled to verify the modeling technique only.

Figure 5.9. Configuration of Simulated Tree Topology Network

5.4.1 Verification of Transfer Function

Figure 5.10 shows the measurement setup for the transfer function verification. Two signal couplers are used to measure the transfer function of the network under test. The line trap is used to block the effects of the outside network to the network under test.

Figure 5.10. Measurement Setup for Transfer Function Measurement

Figure 5.11. Comparison of Derived and Measured Transfer Functions for the 3-Branch Network

In Figure 5.11, the amplitude and the phase angle of the derived transfer functions (using frequency-domain and time-domain approaches respectively) of this 3-branch network are compared to those obtained from measurements. The channel transfer function of the time domain approach is obtained from (5.2) and that of the frequency domain approach is obtained from (5.23). It can be seen that comparing with time-domain approach, the frequency-domain approach matched measurement results much more closely. The frequency-domain approach can accurately predict the channel transfer function of the PLC medium including the positions of attenuation notches in the frequency domain. The notches on the transfer function curve are mainly caused by the branch loads. For example, if the reflected signal from the branch load are totally out of phase (having phase difference of $\pm\pi$ rad) to the original transmitted signal at a certain frequency, then the sum of these two signals will have a very small amplitude at this frequency. This phenomenon will appear as a notch on the transfer function curve.

Figure 5.12. Comparison of Derived and Measured Transfer Functions for the 2-Branch Network

To demonstrate the flexibility and versatility of the derived model, Load 3 (desktop computer) is unplugged from the network. The newly derived and measured transfer functions are then compared as shown in Figure 5.12. The strength of the attenuation

notch near 20MHz has reduced to about -20dB. As can be observed, the frequency-domain approach still performs better than the time-domain approach and can match more closely to the measurement results. The deviation between the time-domain approach and the measurement results become smaller because when the number of branches reduces, the effects of the “no backward reflection” assumption in the time-domain approach will also lessen.

The above verifications have proved that the frequency-domain modeling approach is correct and demonstrate the capability of this model in predicting the behavior of the power line channel. Since the error of time-domain approach will increase with the number of branches while the frequency-domain approach does not have this problem. This is because in the time domain approach, when the network topology becomes complex, the number of reflections in the network increases tremendously. It will be impractical for this approach to consider all these reflections. Thresholding will be necessary to ignore those reflections with small amplitudes, but at the expense of its accuracy. On the other hand, the frequency domain channel transfer function is given by S_{21} and S_{12} in (5.23). Its accuracy is not affected by network topology. As a result, the frequency-domain approach is exclusively adopted in the following modeling sections of this thesis.

5.4.2 Verification of Input Impedance

Figure 5.13. Measurement Setup for Input Impedance Measurement

A similar verification measurement has been conducted for the derived input impedance model. The test network is the same as that in Figure 5.9. The measurement setup is illustrated in Figure 5.13. Port 1 of the test network will be connected to the impedance analyzer and the input impedance looking into port 1 will be measured. From the results shown in Figures 5.14 and 5.15, it can be seen that the model can clearly identify the amplitude peaks and the phase changes of the input impedance of the network under test. The results therefore confirm the applicability and accuracy of the proposed modeling approach.

Knowing the characteristics of the power line channel, like transfer functions and input impedances, will be of great help in analyzing the performance of different communication schemes as well as identifying potential difficulties. From the verification results in Figures 5.11 and 5.12, the frequency domain approach provides an accurate and flexible solution to model the power networks in different configuration, as long as the construction and dimensions of the power cables are known.

Figure 5.14. Comparison of Derived and Measured Input Impedances for the 3-Branch Network

Figure 5.15. Comparison of Derived and Measured Input Impedances for the 2-Branch Network

5.5 SENSITIVITY STUDY OF MODEL

After deriving and verifying the model, in this section, the sensitivity of the model with respect to some network configuration parameters is studied. These parameters include size of cable conductor, distance between cable conductors, length of direct signal path, and length of branch. Sensitivity is an important factor to decide if a model is applicable or otherwise. For example, people would like to know if there is any error in estimating the network configuration, will the model still be able to give the correct representation?

It should be noted that the sensitivity of our model is not only a function of frequency and parameter itself, it is also a function of the impedance of the branched loads. As a result, this becomes a 4-dimensional problem and is hard to demonstrate. To simplify this study, a single-branched network was used (as shown in Figure 5.16). This simplification is still valid as there is no difference in studying the parameters of a single-branched network or a multi-branched network. The OC (Open Circuit) terminal is assumed as the branch load. This assumption makes this network to have a perfect reflection branch since the reflection

coefficient of OC is 1. When studying the sensitivity to one parameter, the other parameters are fixed.

Figure 5.16. The Single-Branch Network Used for Sensitivity Study

Based on this simplified network, the error sensitivity of the outcomes of the derived model, the transfer function and the input impedance, were studied according to some typical network configuration parameters.

5.5.1 Sensitivity to Conductor Size

As the conductor size is an important factor to determine the line parameters, it is therefore critical to the accuracy of the model. Figure 5.17 compares the changes of transfer functions and input impedances according to different conductor sizes for the simplified network.

Figure 5.17. Comparison of (a) Transfer Functions (b) Input Impedances under Different Conductor Sizes

As seen from the figure, the increase of the conductor size will reduce the line attenuation due to the reduction of the resistance (refer to (4.2)). Because of this, the whole curve of the transfer function shifts down but the location of the attenuation notch remains unchanged. On the other hand, the input impedance curve shifts down with the increase of conductor size. This is because that the increase of the conductor size will cause the capacitance between cable conductors to increase, so that the characteristic impedance will decrease. Since the input impedance is positively proportional to characteristic impedance (refer to (5.24) and (5.25)), the overall input impedance will drop in this case.

By using partial differentiation, the sensitivity function of the transfer function to the size of the cable conductor can be expressed as

$$S_{H \rightarrow A} = \frac{\partial H}{\partial A} \text{ (dB/mm}^2\text{)} \quad (5.29)$$

in which H is the output transfer function in dB, and A is the cross-sectional area of the cable conductor in mm^2 . Similarly, the sensitivity function of the input impedance to the size of the cable conductor can be expressed as

$$S_{Z_{in} \rightarrow A} = \frac{\partial Z_{in}}{\partial A} \text{ (\Omega/mm}^2\text{)} \quad (5.30)$$

According to (5.29) and (5.30), the sensitivity of transfer function and input impedance with respect to the change of the conductor size can be obtained as a function of frequency and conductor size.

From Figure 5.18, it can be seen that generally, the sensitivity of transfer function is quite small and the absolute value of the sensitivity is within 5dB/mm^2 . However, the absolute value of the sensitivity of input impedance can be as high as $250 \text{ }\Omega$. Both sensitivity ranges increase with decreases of the conductor size. This is anticipated because a change in the conductor size will be more profound when the conductor size is small. The conclusion is that the derived model will be more prone to error when the conductor size is small.

(a)

(b)

Figure 5.18. Sensitivity of (a) Transfer Function (b) Input Impedance to the Conductor Size

5.5.2 Sensitivity to Distance between Cable Conductors

The distance between cables in the real installation is limited by the conduit size but it is not possible to determine the exact value since the separation between cables is always changing along the power line. In the derived model, an average distance is used and in this section, the effects of this distance to the channel transfer function and input impedance are studied.

Figure 5.19. Comparison of (a) Transfer Functions (b) Input Impedances under Different Distances between Cable Conductors

Figure 5.19 shows the comparison of the transfer functions and input impedances when the distances between cables are different. The conduit sizes for the two cases are same. It can be seen that as the distance differs, both the notch position and notch strength on the transfer function curve will change significantly. This phenomenon can also be observed for input impedances. Both the peak position and peak value of input impedance will change according to different distances between cables. These results imply that the accuracy of the derived model is highly dependent on this parameter.

Similar as (5.29) and (5.30), the sensitivity of the transfer function to the distance between conductors is

$$S_{H \rightarrow D} = \frac{\partial H}{\partial D} \text{ (dB/mm)} \quad (5.31)$$

where D is the distance between cable conductors. And the sensitivity of the input impedance to the distance between conductors is

$$S_{Z_{in} \rightarrow D} = \frac{\partial Z_{in}}{\partial D} \text{ (}\Omega\text{/mm)} \quad (5.32)$$

As Figure 5.20 shows, the range of the transfer function sensitivity is within ± 30 dB/mm and that for the input impedance sensitivity is from $-400 \Omega/\text{mm}$ to $+600 \Omega/\text{mm}$, which are much larger than the sensitivity to conductor size. Hence, this model should be used more carefully when the distances between the conductors are small.

(a)

(b)

Figure 5.20. Sensitivity of (a) Transfer Function (b) Input Impedance to the Distance between Cable Conductors

Fortunately, as mentioned in Chapter 4, although the cables are simply pulled through the metal conduit and the distance between cables conductors keep varying along the length, but because this distance is constrained by the size of the conduit, so on average, the actual distance between cables can still be roughly estimated. That is why the verification results in the previous section illustrate well-matched comparisons between measured and derived transfer functions and input impedances.

5.5.3 Sensitivity to Length of Direct Signal Path

The term direct signal path has been defined in Section 5.2. The length of the direct signal path can be easily overlooked or wrongly determined. This is because in the real installation, the power lines are normally laid inside the concrete wall making it hard to determine their length accurately.

Figure 5.21. Comparison of (a) Transfer Functions (b) Input Impedances under Different Length of Direct Path

From Figure 5.21, as the signal path becomes longer, the whole transfer function curve shifts down because the increase in length raises the line attenuation. However, the location of the attenuation notch for different lengths of signal path does not change with the path length. This also proves that the attenuation notch is generated by branch load and not the length of the direct signal path. However, the input impedances change a lot with the path length. This is because fundamentally the input impedance is a function of signal path (refer to (5.11)).

Similarly, the transfer function sensitivity to the length of direct signal path can be written as

$$S_{H \rightarrow L} = \frac{\partial H}{\partial L} \text{ (dB/m)} \quad (5.33)$$

where L is the path length. The input impedance sensitivity to the length of direct signal path can also be written as

$$S_{Z_{in} \rightarrow L} = \frac{\partial Z_{in}}{\partial L} \text{ (}\Omega/\text{m)} \quad (5.34)$$

(a)

(b)

Figure 5.22. Sensitivity of (a) Transfer Function (b) Input Impedance to the Length of Direct Signal Path

In this case, the absolute value of the transfer function sensitivity is within 6dB/m. The input impedance sensitivity at short path length is very critical, can be up to 1000Ω/m. The range of the sensitivity becomes large when the frequency is high and path length is short.

This is because both short wavelengths (high frequencies) and short path lengths are more sensitive to any error in the estimation of the power line length. As a result, the model is equally more sensitive to estimation error at high frequencies and when the signal path is short.

5.5.4 Sensitivity to Length of Branch

Figure 5.23. Comparison of (a) Transfer Functions (b) Input Impedances under Different Branch Lengths

As described in Section 5.2, the impedance of each of the branches is an important factor in deriving the transfer function of the channel. From (5.10), it is not only a function of the impedance of the terminal load, but also of the length of the branch cable. Normally in the actual power network, this length is the length of the power cord of the appliance, but this may not be always true since people may add extension cords in between appliances and sockets. In this section, the effects of the branched cable length to the model is studied.

Figure 5.23 again proves our previous statement on the effects of the branch loads on the transfer function. As can be seen, as the branch length changes, both the notch position and strength change. From the microwave theory, these changes are due to the amplitude and phase difference of the branch load as the length of branch cable changes. Similar effects can also be seen on the input impedance as shown in Figure 5.23 (b).

The transfer function sensitivity to the length of branch length can be written as

$$S_{H \rightarrow B} = \frac{\partial H}{\partial B} \text{ (dB/m)} \quad (5.35)$$

where B is the length of branch. The input impedance sensitivity to the length of direct signal path can also be written as

$$S_{Z_{in} \rightarrow B} = \frac{\partial Z_{in}}{\partial B} \text{ } (\Omega/\text{m}) \quad (5.36)$$

From Figure 5.24, it can be observed that both sensitivities of transfer function and input impedance will only be quite significant at a certain frequencies and branch lengths. These significant values tend to appear more frequently with the increase in frequency and cable length. The reason is that from (5.14) and (5.15), both transfer function and input impedance are determined by wave reflections in the network, and since in our simplified network as shown in Figure 5.16, the wave reflections are mainly caused by the branch, both the increase of frequency (decrease of wavelength) and increase of branch length will raise the amount of reflections.

(a)

(b)

Figure 5.24. Sensitivity of (a) Transfer Function (b) Input Impedance to the Branch Length

Among the above four network configuration parameters discussed, a change in the conductor size is observed to mainly cause vertical shift of both the transfer function and input impedance, implying that this model has relatively small sensitivities to this parameter. The length of direct path does not cause any significant change to the transfer function but it does have strong effect on the input impedance. Both the distance between cables and length of branch cable cause both the amplitude and the occurrence frequency of the transfer function notches and the input impedance peaks to change. This implies that the model sensitivities to these two latter parameters are relatively large. More attentions are needed to obtain these two parameters in order to ensure the accuracy of the model.

Chapter 6

RING TOPOLOGY NETWORK MODELING

To my knowledge, all of the existing modeling studies only consider power line networks in tree topology (radial circuits). However, according to the British Standard 7671 [28], the electrical power sockets in households can be wired in either ring or radial circuits. Actually, to improve system availability, the sockets in residential houses are normally connected in ring circuits. Therefore, a typical network model for ring-topology network is proposed in this research. It can be regarded as an extension of the proposed tree topology network model described in the previous chapter.

Figure 6.1. Plan Diagram for a Typical Residential House

The power line networks in many residential houses, commercial and industrial buildings do not always have the simple tree topology typically considered in the existing literatures. Figure 6.1 shows the power line wiring plan of a typical residential house in Singapore. It can be seen that after the energy meter, the power sockets are connected in two ring circuits from the distribution board (DB). As illustrated in the diagram, the sockets in the

living room, main bedroom and bedroom 1 are connected together in one ring, while the sockets in bedroom 2, kitchen, and balcony are connected in another ring. Although not shown in this diagram, the other portions of this power network supplying to lighting and high power devices are typically connected to the distribution board in radial circuits (tree topology). Hence the entire in-house power network can be regarded as a hybrid network, comprising of both tree and ring topologies.

The focus of this research is on the portion of the in-house power network where a PLC channel can be formed, i.e. between the power sockets. Since all sockets are arranged in ring circuits, it is therefore important to study and understand the channel characteristics of the ring topology power line communication networks. The next two sections will describe in detail our proposed method for modeling and analyzing the channel transfer function and network input impedance of ring-topology PLC networks. It will be seen that the overall characteristics of the ring network is not simply the superposition of the characteristics of the two paralleled tree networks between the transmitter and the receiver due to the inter-reflection signals in between the two tree networks.

6.1 MODELING OF TRANSFER FUNCTION FOR RING TOPOLOGY NETWORK

Figure 6.2. Simplified Channel Diagram for Transceivers in a Ring Network

The PLC channel in ring topology can be analyzed based on the simplified diagram shown in Figure 6.2. It can be seen that the other portion of the power line network that is

connected to the DB is treated as one of the many loads connected to this ring channel. The channel between the reference planes is no longer a 2-port network as in the tree topology network, but a 4-port network, which can be regarded as two parallel 2-port networks N1 and N2. Each of the N1 or N2 networks is in turn a tree network whose characteristics have been well studied in the literature. However, when these two tree networks form a ring, the complexity of the channel analysis will increase due to signal reflections and interactions between them. In this research, N1 and N2 are called the elementary tree networks and their S-parameter models can be derived from the tree model as described in previous chapter. In the following sections the detailed derivation and verification of the transfer function model for a ring topology network is described.

6.1.1 Derivation of Transfer Function Model

The conventional method to analyze ring type of communication network, which contains a loop inside as shown in Figure 6.2, is to find the short circuit or open circuit point so that the loop can be broken to simplify the 4-port network into two 2-port networks [57]. This method can easily be applied if the network is symmetrical. In such cases, the elementary networks of N1 and N2 should be either identical or following some symmetrical rules. In the case of power line network, the loads are mainly household appliances and their locations depend mainly on where the power sockets are situated, so N1 and N2 are typically not symmetrical to each other. As a result, the conventional method of ring network analysis will be difficult for PLC channel. Our research proposes an alternative analytical solution based on the concept of S-parameters.

Making use of the derived S-parameters of the N1 and N2 elementary tree networks, an equivalent diagram for the ring channel of Figure 6.2 can be obtained as shown in Figure 6.3. If the transmitter is regarded as port 1 and the receiver as port 2, then the channel transfer function is given by the scattering parameter S_{21} of the overall network.

In Figure 6.3, a_1 and a_2 denote the incident waves injected into port 1 and port 2 respectively; b_1 and b_2 denote the reflected waves from port 1 and port 2 respectively. By definition,

$$S_{21} = \left. \frac{b_2}{a_1} \right|_{a_2=0} \quad (6.1)$$

In this research, perfect matching is assumed at the receiver side so that $a_2=0$.

Figure 6.3. Equivalent Diagram of Ring-Topology Network for S_{21} Derivation

When a signal with a certain power is injected from port 1 into the network, part of the signal power will flow to the N1 sub-network while the remaining will flow to the N2 sub-network. For the sake of simplicity, the joints where the transceivers are connected to the ring network are taken as symmetrical power dividers with equal voltage gain to all its ports. Assume the voltage gain is equal to g , then with this assumption, the signal injected from the transmitter to the network will be equally divided to N1 and N2 with the gain g , and similarly the signal injected from N1 (N2) to the power divider 1 will be equally divided between the transmitter and N2 (N1) with the gain g . This assumption is made in this research for the sake of simpler presentation and hence, a voltage gain factor of g to denote equal division is used in subsequent analysis. In general, a variable division factor can be easily incorporated into the proposed model to cater for cases when the joint results in asymmetric division of power.

Considering the signal that flows into N1, part of it ($g \times S_{21_N1}$) will transmit through N1, while the other part ($g \times S_{11_N1}$) is reflected from N1. For the signal that transmits through N1, part of it will flow to the receiver, and the other part will flow to N2 and continue to be transmitted and reflected. In the same way, for the signal that is reflected from N1, part of it will flow back to the transmitter and part of it will transmit through N2 and come to the receiver side. Once again, part of the signal that flows through N2 will flow into the receiver, while the other part flows to N1 and continues to be transmitted and reflected.

These actions will continue until all the signal power has been attenuated or absorbed. The initial signal that flows to N2 will also exhibit similar behavior.

In order to consider all the wave actions inside the network, four directional waves namely, A , B , C and D , are introduced as shown in Figure 6.3. A and C are the incident waves onto the two terminals of N1; while B and D are the incident waves on N2. Although these incident waves will be reflected at the N1 and N2 terminals resulting in more waves, it will be shown later that these four wave definitions are sufficient to represent all the wave behaviors inside the ring-topology network.

There are two rules that describe the relationship between these four waves. The first rule is that: the four waves form loops in between each other. Specifically, the cause of one wave is related to the other two waves, while at the same time, one wave will result in another two waves. For example, wave D is formed from the combination of the transmitted wave A (through N1) and the reflected wave C (by N1). On the other hand, the transmitted wave D (through N2) can result in wave A while the reflected wave D (by N2) can result in wave C .

The second rule is that: these loops in between the waves are infinite. Therefore, it is necessary to differentiate the waves at different stages of the loops. For example, wave D that is resulted from wave A_n (wave A at stage n) is defined as D_{n+1} (wave D at stage $n+1$). With these definitions, four recursive equations that relate the waves at stage n and stage $n+1$ can be written as follows,

$$\begin{cases} A_{n+1} = g \cdot S_{11_N2} \cdot B_n + g \cdot S_{12_N2} \cdot D_n \\ B_{n+1} = g \cdot S_{11_N1} \cdot A_n + g \cdot S_{12_N1} \cdot C_n \\ C_{n+1} = g \cdot S_{21_N2} \cdot B_n + g \cdot S_{22_N2} \cdot D_n \\ D_{n+1} = g \cdot S_{21_N1} \cdot A_n + g \cdot S_{22_N1} \cdot C_n \end{cases} \quad (6.2)$$

All the “ g ” terms in (6.2) are based on the assumption that the power dividers are symmetrical and have equal gains for all ports. If the power divider is not symmetrical, then the division factor can be obtained from measurement. After measured the division factor through the network analyzer, this factor can be substituted for the assumed g division factors in the equations. From (6.1), if $a_1=1$ and $a_2=0$, then $S_{21}=b_2$. Due to power divider 2 in Figure 6.3, b_2 wave will be formed whenever there are C and D waves. i.e.,

$$S_{21} \Big|_{a_1=1, a_2=0} = b_2 = \sum_{i=0}^{\infty} (C_i + D_i) \quad (6.3)$$

If port 1 is transmitting while port 2 is receiving, i.e. $a_1=1$ and $a_2=0$, the initial (stage 0) values of the four waves can be written as

$$\begin{cases} A_0 = g \\ B_0 = g \\ C_0 = g \cdot S_{21_N2} \cdot B_0 \\ D_0 = g \cdot S_{21_N1} \cdot A_0 \end{cases} \quad (6.4)$$

Combining (6.2) and (6.4), S_{21} in (6.3) for this ring type network can be determined.

Note that the summation in (6.3) is infinite but it can be simplified. To achieve this, a vector $\mathbf{X}_n$ at stage n is defined as,

$$\mathbf{X}_n = [A_n \ B_n \ C_n \ D_n]^T$$

and (6.2) can be re-written compactly in matrix form

$$\mathbf{X}_{n+1} = \mathbf{U} \cdot \mathbf{X}_n \quad (6.5)$$

where

$$\mathbf{U} = \begin{bmatrix} 0 & g \cdot S_{11_N2} & 0 & g \cdot S_{12_N2} \\ g \cdot S_{11_N1} & 0 & g \cdot S_{12_N1} & 0 \\ 0 & g \cdot S_{21_N1} & 0 & g \cdot S_{22_N2} \\ g \cdot S_{21_N1} & 0 & g \cdot S_{22_N1} & 0 \end{bmatrix}$$

Then the infinite summation for all stages of wave C can be expressed as

$$C_{tot} = \sum_{i=0}^{\infty} C_i = [0 \ 0 \ 1 \ 0] \cdot [(\mathbf{I} + \mathbf{U} + \mathbf{U}^2 + \mathbf{U}^3 + \dots) \cdot \mathbf{X}_0] \quad (6.6)$$

in which $\mathbf{I}$ is a 4×4 identity matrix. It can be shown that (6.6) can be simplified to

$$C_{tot} = [0 \ 0 \ 1 \ 0] \cdot [(\mathbf{I} - \mathbf{U})^{-1} \cdot \mathbf{X}_0] \quad (6.7)$$

if the infinite sum of the geometric progression of matrix $\mathbf{U}$ in (6.6) converges. Similarly, the total summation of wave D is

$$D_{tot} = [0 \ 0 \ 0 \ 1] \cdot [(\mathbf{I} - \mathbf{U})^{-1} \cdot \mathbf{X}_0] \quad (6.8)$$

As a result, (6.3) can be expressed in a closed form,

$$S_{21} = C_{tot} + D_{tot} = \begin{bmatrix} 0 & 0 & 1 & 1 \end{bmatrix} \cdot \left[(\mathbf{I} - \mathbf{U})^{-1} \cdot \mathbf{X}_0 \right] \quad (6.9)$$

It is important to note that (6.9) is valid if and only if the infinite sum of the geometric progression of matrix $\mathbf{U}$ converges. This is proven to be always true for electrical power networks in Appendix A.2.

6.1.2 Verification of Transfer Function Model

A test ring-topology power network with three branches of load as shown in Figure 6.4 has been constructed to verify the above transfer function derivations. The corresponding measurement set up is same as shown in Figure 5.10.

Figure 6.4. Configuration of Simulated Ring Topology Network

Figure 6.5 compares the amplitude and phase responses of the derived transfer function of the network under test against those obtained from measurements. As shown in the amplitude response plot, the analytical model accurately identifies the positions of the spectral notches while its magnitude tracks the measured values closely throughout the entire frequency range. Similarly, the phase response also shows close agreement between the analytical model and measurement, demonstrating that the proposed modeling approach is accurate.

Figure 6.5. Comparison of Derived and Measured Transfer Functions

It should be noted that the above results are applicable to this simple test network only. They are meant for verifying the appropriateness of the modeling technique only and should not be taken as universal for all possible ring networks.

6.2 MODELING OF INPUT IMPEDANCE FOR RING TOPOLOGY NETWORK

The input impedance not only provides the basis for designing impedance matching circuits for the transceiver, it also serves an essential role in the transfer function modeling of large PLC networks. For example in Figure 6.1, there are two rings in the residential network. If the PLC transceivers are on the same ring, then the other ring with its connected appliances can be treated as a load impedance connected to the DB. Before determining the S-parameters of each of the elementary tree networks, the input impedance of the other ring network must be obtained first and regarded as a separate load impedance connected to the DB.

6.2.1 Derivation of Input Impedance Model

From the transmission line theory, the input impedance is given by

$$Z_{in} = Z_0 \cdot \frac{1 + S_{11}}{1 - S_{11}} \quad (6.10)$$

where Z_0 is the characteristic impedance of the measurement system, which is assumed to be the typical value of 50Ω in the proposed modeling approach and verification measurements described in this chapter. To determine Z_{in} , the model in Figure 6.6 is used to derive S_{11} first. The whole network is now treated as a one-port network with a loop inside. It can be seen that the portion of the network that is on the right hand side of the reference plane can just be regarded as a tree topology network N, for which a corresponding scattering matrix can be determined. Similar to the derivation of transfer function, the behavior of the waves inside this network can be analyzed using two directional waves E and F acting as incident waves from the two terminals respectively.

Figure 6.6. Equivalent Diagram of Ring-Topology Network for S_{11} Derivation

From Figure 6.6, the reflection coefficient S_{11} at port 1 is defined as

$$S_{11} = \left. \frac{b_1}{a_1} \right|_{a_2=0} \quad (6.11)$$

With the assumption of perfect matching at the receiver side, $a_2=0$, and assuming that $a_1=1$, resulting in $S_{11}=b_1$, the initial values for E and F are

$$\begin{cases} E_0 = g \\ F_0 = g \end{cases} \quad (6.12)$$

assuming that the joint at port 1 is a symmetrical power divider with a voltage gain g .

The final E and F waves are results of multiple stages of transmissions through the network N and reflections from the network N. For example, a portion of E_n wave at stage

n is reflected by N , forming wave F_{n+1} of stage $n+1$. At the same time, the other portion of E_n that transmits through N will contribute to E_{n+1} of stage $n+1$. Likewise, F wave will exhibit similar characteristics and effects on the subsequent stages of E wave and itself. The relationship between the waves at stage n and stage $n+1$ can be written as,

$$\begin{cases} E_{n+1} = g \cdot S_{21} \cdot E_n + g \cdot S_{22} \cdot F_n \\ F_{n+1} = g \cdot S_{11} \cdot E_n + g \cdot S_{12} \cdot F_n \end{cases} \quad (6.13)$$

From Figure 6.6, except for the initial stage, whenever there are E and F waves, b_1 wave will be formed such that

$$S_{11} \Big|_{a_1=1, a_2=0} = b_1 = \sum_{i=1}^{\infty} (E_i + F_i) \quad (6.14)$$

Again, to transform this infinite summation into a closed form, a vector $\mathbf{Y}_n$ at stage n is defined as,

$$\mathbf{Y}_n = [E_n \quad F_n]^T$$

and (6.12) can be rewritten into a matrix form,

$$\mathbf{Y}_{n+1} = \mathbf{V} \cdot \mathbf{Y}_n \quad (6.15)$$

where

$$\mathbf{V} = \begin{bmatrix} g \cdot S_{21} & g \cdot S_{22} \\ g \cdot S_{11} & g \cdot S_{12} \end{bmatrix}$$

Then the total summation of all stages of wave E can be expressed as

$$E_{tot} = \sum_{i=1}^{\infty} E_i = [1 \quad 0] \cdot \left[(\mathbf{V} + \mathbf{V}^2 + \mathbf{V}^3 + \dots) \cdot \mathbf{Y}_0 \right] \quad (6.16)$$

If the infinite sum of the geometric progression of matrix $\mathbf{V}$ converges as proven in Appendix A.2, (6.16) can be simplified to

$$E_{tot} = [1 \quad 0] \cdot \left\{ \left[(\mathbf{I} - \mathbf{V})^{-1} - \mathbf{I} \right] \cdot \mathbf{Y}_0 \right\} \quad (6.17)$$

Similarly, the total summation of all stages of wave F is

$$F_{tot} = [0 \quad 1] \cdot \left\{ \left[(\mathbf{I} - \mathbf{V})^{-1} - \mathbf{I} \right] \cdot \mathbf{Y}_0 \right\} \quad (6.18)$$

As a result, (6.14) can be expressed in a closed form as

$$S_{11} = E_{tot} + F_{tot} = [1 \quad 1] \cdot \left\{ \left[(\mathbf{I} - \mathbf{V})^{-1} - \mathbf{I} \right] \cdot \mathbf{Y}_0 \right\} \quad (6.19)$$

6.2.2 Verification of Input Impedance Model

Figure 6.7. Comparison of Derived and Measured Input Impedances

A similar verification measurement has been conducted for the input impedance model. The test network is the same as that in Figure 6.4 and the measurement setup is the same as that is shown in Figure 5.13. From the results shown in Figure 6.7, it can be seen that the model can clearly identify the amplitude peaks and the phase changes of the input impedance of the network under test. The results therefore confirm the applicability and accuracy of the proposed modeling approach.

In this chapter, although examples are given only for PLC network with one ring, the proposed ring topology channel modeling approaches described in Sections 6.1 and 6.2 can be easily applied to model the characteristics of more practical power networks with multiple connected rings as shown in Figure 6.1. Specifically, if the transceivers of interest are on the same ring, the model earlier in this research is directly applicable. On the other hand, if the transceivers are located on separate rings, their channel model can still be derived based on the equivalent channel diagram shown in Figure 6.8.

Figure 6.8. Simplified Channel Diagram for PLC Transceivers on Different Rings

In addition, the other portions of the network are regarded as a load impedance connected to one of the rings. The total transfer function can then be obtained in two steps: (1) Determine the scattering matrices for ring 1 and ring 2 separately using the method presented in this research. (2) Use the approach of cascaded chain scattering matrices to obtain the total transfer function. Hence, the channel modeling approach presented in this chapter is useful for analyzing the characteristics of practical PLC networks with generalized ring topology.

The modeling techniques described in Chapters 5 and 6 are proposed for single-phase power networks only. To extend the current technique to a three-phase system, some studies on the interferences and cross-talks in between the phases need to be considered.

Chapter 7

NOISE MODELING

In this chapter, the approaches of modelling the third channel parameter --- noises, are discussed. Although power line noise is categorized into background noise and impulsive noise, there should not be a rigid border or distinction between them. In other words, the background noise can also contain impulsive-like noise and the impulsive noise can also contribute to the background noise. They are classified into two categories just according to how they vary over time. The background noise is always there and changes with time slowly while the impulsive noise is due to the network loading conditions and it varies relatively rapidly with time. As a result, the background noise determines the long term or average system performance. However, the impulsive noise will dominate if some appliances, which can generate strong impulsive noise, are operating near the PLC transceivers.

In the following sections of this chapter, a frequency-domain approach is used to characterize both the background noise and the appliance noise from measurements. The final noise model serves to provide the total noise expected at the PLC receiver. In other words, this receiver noise is the sum of the background noise and the noises generated by all closely appliances.

7.1 BACKGROUND NOISES

To model the background noise, long-term measurements on noise spectrum from 1MHz to 30MHz have been conducted at two sites, a laboratory and a residential house. In each measurement, for every five minutes, one set of spectrum data was recorded and the whole measurement lasted for one week. This long-term measurement is controlled by the software HP-VEE, which can be referred to Appendix A.1. The measurement results have given us sufficient information on how the background noises at different sites change

with frequency and time. Figure 7.1 shows the measurement setup for the noise spectrum measurement.

Figure 7.1. Measurement Setup for Noise Spectrum Measurement

Figure 7.2 shows the measured noise amplitude spectrums in laboratory and residential house respectively. The unit of the noise amplitude spectrum ($\text{dB}\mu\text{V}/\text{Hz}^{1/2}$) is the measured noise format from the spectrum analyzer. According to [60], when the analyzer measures the noise, the measurement value is normalized by an equivalent bandwidth of the RBW (Resolution Bandwidth) filter, so that the unit ($\text{dB}\mu\text{V}/\text{Hz}^{1/2}$) is the normalized noise amplitude in the frequency domain. To illustrate the measured results in a more vivid way, the mean and standard deviation of the noise amplitudes at the two sites are plotted in Figure 7.3.

(a)

Figure 7.2. Noise Amplitude Spectrum in (a) Laboratory (b) Residential House

As seen from the figures, both the mean and standard deviation of the noise spectrum in time domain have relatively large values at lower frequencies (1MHz~10MHz), and relatively smaller values at higher frequencies (10MHz~30MHz). This is reasonable since there are more noise sources at lower frequencies in the power network and the short wave (SW) radios in Singapore are also concentrated in this frequency range [61]. After comparing the mean and the standard deviation values, it can be seen that the noise spectrum does not change a lot with time at both sites.

(a)

Figure 7.3. Time Domain Mean and Standard Deviation of Noise Spectrum in (a) Laboratory (b) Residential House

However, the mean and the standard deviation can only roughly describe the variations of the noise with respect to time. To further understand the exact random behaviour of the noise in the time domain, statistical analysis method needs to be used. Although as mentioned in Section 2.2.2 that several PDFs have been suggested in the literature, in this research, and after the extensive study of the measured noise amplitude spectrums, it is found that the probability distribution of the time-domain noise amplitudes resembles the Nakagami- m distribution. The Nakagami- m PDF can be written as

$$p(r) = \frac{2}{\Gamma(m)} \left(\frac{m}{\Omega} \right)^m r^{2m-1} e^{-\frac{mr^2}{\Omega}} \quad (7.1)$$

where r is the random variable, p is the probability of the corresponding random variable, $\Gamma(\bullet)$ is the Gamma function, m is defined as the ratio of moments, and Ω is the mean of the power of the random variable.

The Nakagami model [80] is often used to represent the fading in multi-path scattering with relatively large delay-time spreads, and with different clusters of reflected waves. This model can also be used to describe the interferences cumulated from multiple independent Rayleigh fading sources [62]. In the power network, although the average amplitude of the noise coming from a single source is roughly unchanged, but due to the complex scattering in the network, the resultant noise at the receiver can be roughly

regarded as Rayleigh distributed. Since the background noise is the combination of these Rayleigh distributed noises from multiple sources, it is reasonable to use the Nakagami PDF to represent the noise amplitude distribution. From (7.1), a Nakagami distribution is determined by two parameters, m and Ω . The value of m denotes the closeness between the Nakagami and Rayleigh distributions while Ω denotes the power of the random variable. As shown in Figure 7.4, the Nakagami PDF is exactly the same as Rayleigh PDF when $m=1$. When m is larger than 1, the amplitude of random variable is more concentrated to a certain value while when m is smaller than 1, the amplitude of random variable is more wide spread, as compared to the Rayleigh PDF. If the noise amplitude is Rayleigh distributed, both real and imaginary parts of the noise will be Gaussian distributed and in this condition the noise can be regarded as white Gaussian noise.

Figure 7.4. Comparison of Rayleigh and Nakagami Distributions

By using Least Mean Square (LMS) fitting, the frequency-domain m and Ω profiles of the measured noise at the two sites are obtained as shown in Figure 7.5. The m profiles of the two sites look quite different but they still show some similarities. They both imply that the noise is close to white Gaussian at high frequencies while at low frequencies, the probability of the noise amplitudes will be more wide spread over the amplitude range. For measurement done in the laboratory, the noise PDF approaches Gaussian distribution above 22MHz, while for residential house this happens above 8MHz.

(a)

(b)

Figure 7.5. m and Ω Profile of (a) Laboratory (b) Residential House

To illustrate the validity and accuracy of using Nakagami- m PDF to model the power line background noise, the modelled and measured PDFs of the two test sites are compared in Figure 7.6. Only 5MHz and 25MHz are chosen since at 5MHz, the m value for both test sites is less than 1, and at 25MHz, the m value for both sites is close to 1. The maximum m value for both sites is around 1.

Figure 7.6. Comparison of Measured and Modelled Noise PDF at 5MHz and 25MHz for (a) Laboratory (b) Residential House

The comparisons show that our modelled PDFs are well matched to the measured PDFs at both test sites although the m and Ω profiles of the two sites are different. This demonstrates that using Nakagami- m distribution to approximate the background noise is correct.

7.2 APPLIANCE NOISES

The noise spectrums from appliances are also obtained through measurements. To do this, the measurement setup shown in Figure 7.7 was used. The procedure can be divided into three steps. Step one is to first measure the background noise spectrum with the switch opened. Then in step two, the switch is closed to measure the summation of the background noise and the appliance noise. Finally, subtracting the spectrum of step two

from that of step one gives the noise spectrum generated by the appliance. It should be noticed that the noise spectrum used in this portion should be the power spectrum. This is because that the noises from different sources are non-coherent in power line, so that they can only be added or subtracted in power but not amplitude.

Figure 7.7. Measurement Setup for Appliance Noise Spectrum Measurement

Figure 7.8. Comparison between the Background Noises Measured with and without Isolation Transformer

Comparing to the previous setup in background noise measurement (Figure 7.1), a 1:1 isolation transformer is added in between the signal coupler and the power network. This setup was proposed in [36], and the main reason for adding an isolation transformer is to provide an isolation between the background noise coming from the external power

network and from the appliance under test. Although the actual measurement results show that this isolation is quite limited, as shown in Figure 7.8, it still can provide up to 10dB isolation at high frequencies.

Figure 7.9. Noise Amplitude Spectrums of Several Appliances

By using the measurement setup shown in Figure 7.7, the noise spectrums of several appliances were measured and the results are shown in Figure 7.9. Except for fluorescent light and vacuum cleaner, it is observed that other appliances do not contribute much to the noise in the frequency band of 1MHz~30MHz. Among these appliances, cooker and dimmer use linear power supplies with either center trap transformer input or full-wave bridge rectifier input. On the other hand, TV and computer use switching mode power supplies. Fan and vacuum cleaner are basically motors, but the power of the vacuum cleaner is much higher than that of the fan. Hence, the vacuum cleaner produces much stronger noise at the high frequencies. Light dimmer uses triac while fluorescent light is nothing but a bulb filled with inert gas. For the fluorescent light, the strong noise should be due to the impulsive current caused by the electron emission inside the inert gas.

7.3 NOISES AT THE RECEIVER

Figure 7.10. Illustration of Modeling for Total Receiver Noise

The noise at the receiver is always an important parameter in the design of any communication system. With the knowledge of this parameter, the channel SNR can be obtained to serve as the essential factor for designing the modulation and coding schemes. In this research, as mentioned earlier, the noise at the receiver is modeled as a combination of the background noise and the noises from nearby appliances. Figure 7.10 illustrates the

modeling approach for the total noise receiver. $S_i(f)$ is the noise power spectrum generated by i^{th} noise source (or appliance) and $H_i(f)$ is the transfer function from i^{th} noise source location to the receiver so that $S_i(f) \times |H_i(f)|^2$ is the contribution to the receiver noise spectrum from i^{th} noise source. Since they are from different sources, the noises from different appliance are un-correlated. As a result, the total appliance noise spectrum at the receiver ($S_{rx_appliance}$) can be determined by superposition through the summation of noise power spectrums from all the appliances,

$$S_{rx_appliance}(f) = \sum_{i=1}^N \left[S_i(f) \cdot |H_i(f)|^2 \right] \quad (7.2)$$

$H_i(f)$ can be obtained by the transfer function model as proposed in Chapters 5 and 7. Based on (7.2), the total noise at the receiver is the summation of the background noise and the total appliance noise at the receiver,

$$\begin{aligned} S_{rx}(f) &= S_{background}(f) + S_{rx_appliance}(f) \\ &= S_{background}(f) + \sum_{i=1}^N \left[S_i(f) \cdot |H_i(f)|^2 \right] \end{aligned} \quad (7.3)$$

To verify the derived noise model, a test network shown in Figure 7.11 was constructed. A fluorescent light and a vacuum cleaner were chosen as loads in the network mainly because their noise spectrums in the frequency band of interest are quite significant and hence are suitable for the verification purpose. Port 1 of the network is connected to the measurement setup shown in Figure 7.7.

Figure 7.11. Test Network for Noise Model Verification

Using the noise spectrum obtained from Figure 7.9, the noise spectrum at Port 1 can be determined from (7.2). The $|H_i(f)|$ in (7.2) is derived from tree topology channel transfer model described in Section 5.2.

Figure 7.12. Comparison of Measured and Modelled Noise Spectrum at the Receiver of the Network under Test

The measured and modeled receiver noise spectrums are compared in Figure 7.12. The overall trend of the theoretical model follows well with that of measurement except around 18MHz. From this verification, it is demonstrated that this approach to model the noise at the receiver is reasonably accurate.

Although here a test network with tree topology is used in the example, this noise model can be equally applied to the ring topology. To do it, just change the transfer function $|H_i(f)|$ in (7.2) to the channel transfer function between the appliance (noise source) and the receiver on the ring network. The modeled receiver noise in this chapter is used in the subsequent chapter as one of the parameters for designing and optimizing the performance of a communication system over the power line channel. Since the main purpose of the next chapter is to show readers how the proposed model can be applied in the system design, the focus is on the average system performance. Hence, it is only necessary to consider the background noise in the noise model. However, if some appliances, which are strong noise sources, are connected near to the PLC transceivers, the characteristic of the noise in the channel will be quite different from that of the background noise. In such cases, to study the system performance, the appliances' noise must also be considered. However, this will need a very accurate time domain model to account for the appliances'

noises. The time-domain analysis of the appliances' noises is not included in this research, but it is regarded as an important future work for this research as discussed in Section 9.2.

Chapter 8

PLC SYSTEM DESIGN

In the previous three chapters, the proposed modeling techniques for transfer function, input impedance, and noise, are discussed. The focus of this chapter is on the applications of the channel model on the PLC system design. Specifically, several usages of the channel models are studied which include how to use the channel transfer function model to determine the PLC system parameters, how to apply the noise model to predict the system performance, and how to optimize the system performance based on the transfer function and noise models.

Before studying all these model applications, the modulation scheme of the PLC system must be first determined. The selection of a proper modulation scheme for PLC must consider the following three factors:

1. The presence of the Nakagami-m distributed background noise and the impulse disturbances.
2. The multi-path nature of the channel resulting in the frequency selective characteristic of the channel transfer function.
3. Regulation constraints on the transmission power of the PLC signal due to the EMC consideration.

As a result, the choice for a proper modulation scheme is one that is both robust and adaptive. Currently, modulation schemes such as Frequency Shift Keying (FSK) [63], Direct Sequence Spread Spectrum (DSSS) [64, 65] or sometimes called Direct Sequence - Code Division Multiple Access (DS-CDMA) in the multi-user case, and OFDM [66-68] or sometimes called DMT in DSL standards, have either been suggested as appropriate choices or used for PLC. Depending on the target application, each modulation technique has its own advantages and disadvantages.

For a low cost and low data rate PLC system, FSK is a good solution. On the other hand, DSSS and OFDM are two strong competitors for the high data rate system. The advantages in DSSS technique is its remarkable robustness to narrow band interference

and the system's inherent processing gain can be utilized to meet radiation limits. On the other hand, OFDM technique can be optimized to greatly reduce the multi-path effect of the channel and to increase the resistance towards narrow band and impulsive noise. Furthermore, adaptive techniques can be applied to OFDM to increase the system capacity. In a recent research work, the performances of these two techniques are compared and it is concluded that the OFDM with bit loading performs much better than that of DSSS [69]. As a result, in this research, the OFDM system is adopted for studying the applications of the proposed channel model. These applications include determining the system parameters of OFDM, predicting the error performance of the system under the effect of power line noise, and optimizing the OFDM by using adaptive loading technique considering the multi-path fading environment of power network.

8.1 DETERMINATION OF SYSTEM PARAMETERS

In this section, the determination of the OFDM system parameters by applying the derived transfer function model is discussed. Using the ring topology model described in Chapter 6, the transfer function of the PLC network in a typical residential house (shown in Fig. 6.1) is determined. Several household appliances are incorporated into the model, including cooker, dimmer, fan, fridge, kettle, microwave oven, PC, TV, fluorescent light, and vacuum cleaner. This model allows complete freedom in freely modifying the locations of the PLC transceivers and the appliances on the network. The on and off states of the appliances can also be controlled.

With this model, a particular transfer function of the channel formed in the house can be simulated as shown in Figure 8.1 (a). By simply performing an inverse Fourier transform on this transfer function, the impulse response of this channel can be obtained as shown in Figure 8.1 (b). If the delay threshold amplitude is selected as 0.005 V, or 24dB lower than the maximum impulse amplitude, the maximum excess delay of the channel can be estimated to be 0.35 μ s.

Figure 8.1. Transfer Function and Impulse Response of a Particular PLC Channel in Residential House

To reduce the multi-path effects of this channel, the duration of the guard interval (T_g) in the OFDM symbol needs to be larger than the maximum excess delay. Let's select $T_g=0.4\mu\text{s}$. A common way of determining the OFDM symbol duration (T_s) is to make it

four times the duration of the guard interval, so that $T_s = 4 \times T_g = 1.6 \mu s$. Subsequently, the inter-carrier spacing Δf for this symbol duration can then be obtained,

$$\Delta f = \frac{1}{T_s} = 0.625(\text{MHz})$$

Since our bandwidth of interest (B) is from 1MHz to 30MHz, the number of sub-carriers can be determined by

$$N = \frac{B}{\Delta f} = \frac{30-1}{0.625} = 46.4$$

Because the number of carriers needs to be an integer and also normally it is chosen as the power of 2 for the convenience of FFT implementation, $N=64$ is selected for the number of sub-carriers used in the OFDM in-house PLC system.

By using the above method, which is based on our proposed channel model, the OFDM parameters for implementing PLC in power networks of various topologies and loading conditions can be easily determined as described in the following sections.

8.2 PREDICTION OF SYSTEM PERFORMANCE

Although the proposed noise model can model the receiver noise spectrum including both background noise and appliance noise, normally the system performance study is based on the long-term noise. Hence, only the background noise is considered when designing the PLC modulation system. As characterized in Chapter 7, the background noise at each single frequency follows the Nakagami-m distribution. To study the effects of this kind of noise to modulation systems, the axis-PDF of the noise needs to be determined. The axis-PDF stands for the probability distribution function of the noise on the real and imaginary axes. For example, the amplitude distribution of the well-known Additive White Gaussian Noise (AWGN) is the Rayleigh distribution. This type of noise has Gaussian distributions on the real and imaginary axis so that the axis-PDF for AWGN is Gaussian distribution. With the knowledge of the axis-PDF of the noise, the error performance of the modulation schemes under the noise can be easily studied [70].

The amplitude PDF of the power line background noise is shown in (7.1). If the phase of the noise is assumed to be uniformly distributed, and letting the random variable of noise amplitude to be x , and the random variable of noise phase to be θ , and the real portion of the noise to be y , then

$$p(x) = \frac{2}{\Gamma(m)} \left(\frac{m}{\Omega} \right)^m x^{2m-1} e^{-\frac{mx^2}{\Omega}} \quad (8.1)$$

and,

$$y = x \cdot \cos(\theta) \quad \Rightarrow \quad x = \frac{y}{\cos(\theta)} \quad (8.2)$$

so that

$$\frac{dy}{dx} = \cos(\theta) \quad (8.3)$$

From [71],

$$p(x) \cdot dx = p(y) \cdot dy \quad (8.4)$$

if the noise phase θ is regarded as constant, then the PDF of y can be determined as follows,

$$\begin{aligned} p(y)|_{\theta} &= \frac{p(x)}{dy/dx} = \frac{2}{\Gamma(m) \cdot \cos(\theta)} \left(\frac{m}{\Omega} \right)^m x^{2m-1} e^{-\frac{mx^2}{\Omega}} \\ &= \frac{2y^{2m-1}}{\Gamma(m) \cdot \cos^{2m}(\theta)} \left(\frac{m}{\Omega} \right)^m e^{-\frac{my^2}{\Omega \cos^2(\theta)}} \end{aligned} \quad (8.5)$$

However, since θ is a uniformly distributed random variable, the joint PDF for y and θ will be

$$\begin{aligned} p(y, \theta) &= \frac{2y^{2m-1}}{\Gamma(m) \cdot \cos^{2m}(\theta)} \left(\frac{m}{\Omega} \right)^m e^{-\frac{my^2}{\Omega \cos^2(\theta)}} \times \frac{1}{2\pi} \\ &= \frac{y^{2m-1}}{\pi \cdot \Gamma(m) \cdot \cos^{2m}(\theta)} \left(\frac{m}{\Omega} \right)^m e^{-\frac{my^2}{\Omega \cos^2(\theta)}} \end{aligned} \quad (8.6)$$

then the PDF of $y \geq 0$ can be obtained,

$$\begin{aligned}
p(y) &= \int_0^{\pi} p(y, \theta) \cdot d\theta \\
&= \int_{-\frac{\pi}{2}}^{\frac{\pi}{2}} \left(\frac{y^{2m-1}}{\pi \cdot \Gamma(m) \cdot \cos^{2m}(\theta)} \left(\frac{m}{\Omega} \right)^m e^{-\frac{my^2}{\Omega \cos^2(\theta)}} \right) \cdot d\theta \quad (8.7) \\
&= 2 \cdot \int_0^{\frac{\pi}{2}} \left(\frac{y^{2m-1}}{\pi \cdot \Gamma(m) \cdot \cos^{2m}(\theta)} \left(\frac{m}{\Omega} \right)^m e^{-\frac{my^2}{\Omega \cos^2(\theta)}} \right) \cdot d\theta
\end{aligned}$$

Since the phase is assumed to be uniformly distributed, the probability for $y < 0$ will be the same as that for $y > 0$. Unfortunately, the integration in (8.7) can not be solved easily. In this research, the numerical integration method is used. (8.7) is numerically integrated with respect to θ ,

$$p(y) = \frac{4}{N} \sum_{i=0}^{N-1} \frac{y^{2m-1}}{\Gamma(m) \cdot \cos^{2m}\left(\frac{i \cdot \pi}{2N}\right)} \left(\frac{m}{\Omega} \right)^m e^{-\frac{my^2}{\Omega \cos^2\left(\frac{i \cdot \pi}{2N}\right)}} \quad (8.8)$$

where N is the number of steps in the numerical integration. This result has been verified by simulation as shown in following figure.

Figure 8.2. Comparison of Simulated and Modelled axis-PDF for the Noise at Unit Power and with $m=0.7$

Figure 8.2 shows the comparison of the simulated axis-PDF and the modeled axis-PDF. The simulated PDF is obtained from a large number of simulated noise random variable, whose amplitude follows the Nakagami distribution while the phase follows the uniform

distribution. On the other hand, the modeled PDF is computed from (8.8). The two PDFs follow each other closely, which demonstrates that the derived axis-PDF of (8.8) is correct.

With the help of noise amplitude axis-PDF $p(y)$, the error performance under the power line noise can be studied. The procedure is similar to the analysis with AWGN. Supposing that Binary Phase Shift Keying (BPSK) is used and the signal amplitude is $\pm A$, the bit error rate can be estimated by

$$p_e|_{BPSK} = \frac{1}{2} \cdot [p(y \leq -A) + p(y \geq +A)] \quad (8.9)$$

Since $p(y)$ is symmetrical in the real axis, (8.9) can be simplified to

$$p_e|_{BPSK} = 1 - P(Y = A) \quad (8.10)$$

where $P(Y)$ is the Cumulated Distribution Function (CDF) of random variable y ,

$$P(Y) = \int_{-\infty}^y p(y) \cdot dy \quad (8.11)$$

$P(Y)$ can also be obtained through numerical integration.

Figure 8.3. Block Diagram for BPSK Modulation System

To verify the derived bit error rate in (8.10), a Monte Carlo simulation is performed on the BPSK system as shown in Figure 8.3. Here the input and output data are assumed to be digitized data, so that A/D and D/A converters are not included in the block diagram. Since our purpose is to study the noise effects, the fading effect of the channel is not included in this block diagram. Figure 8.4 compares the Bit Error Rate (BER) under different Nakagami distributed noise. It can be seen that the modeled BER matches the simulated one closely. It should be noticed that all the SNRs used in this chapter indicate the ratio between the power of the signal before optimizing and the power of the channel noise.

Figure 8.4. Comparison of Simulated and Modelled BER under Nakagami Noises of Different m Values in BPSK System

In BPSK system, the error performance becomes worse when the Nakagami's m value becomes small. When $m=1.0$, with which the noise is actually Gaussian distributed, the system has the smallest BER, which means that the lower frequency carriers perform worse under the power line noise than under the AWGN.

It is well known that under AWGN, the performance of the system with or without OFDM are actually the same. This is because OFDM demodulation is actually performing the Fast Fourier Transform (FFT), which is a linear process. The additive Gaussian noise, even after going through the linear FFT process alongside the OFDM signal during demodulation, is still Gaussian distributed. On the other hand, the power line noise is found to be Nakagami distributed. It is uncertain what the distribution of the power line noise is after it has been FFTed.

There are two methods which can be used to find this distribution. The first one is the conventional method, where the FFT is written as

$$A(k) = \sum_{n=1}^N a(i) \cdot e^{-j \frac{2\pi k i}{N}} \quad (8.12)$$

where $A(k)$ is the frequency domain samples, $a(i)$ is the time domain samples, and N is the FFT size. The random variable of the power line noise can be regarded as time domain samples, such that this FFT is actually adding a phase shift to the noise random variable and subsequently all N variables are summed together. Since the noise random variable has a uniformly distributed phase, the phase adding effect in (8.12) can be ignored, so that the FFT of the noise variable is just the summation of N noise variables. Letting the random variable after FFTing of the noise to be M , and the noise random variable to be n , then

$$M = n_1 + n_2 + \cdots + n_N$$

Letting $M_1 = n_1 + n_2$, and from [71],

$$\begin{aligned} p(M_1) &= \iint p_n(n_1) \cdot p_n(n_2) \cdot dn_1 dn_2 \\ &= \int p_n(M_1 - n_2) \cdot p_n(n_2) \cdot dn_2 \end{aligned}$$

Similarly, letting $M_2 = M_1 + n_3$,

$$p(M_2) = \int p_{M_1}(M_2 - n_3) \cdot p_n(n_3) \cdot dn_3$$

Continuing this iteration until $M = M_{N-1} = M_{N-2} + n_N$,

$$p(M) = \int p_{M_{N-2}}(M - n_N) \cdot p_n(n_N) \cdot dn_N$$

As shown above, this method involves many integrations and it is really difficult to obtain a closed form for the distribution of M .

In this research, a statistical approximation method is used. According to the Central Limit theorem, for a large population, if a sample mean $\bar{x}$ is obtained from samples taken from this population and the samples are with sufficiently large size, then the distribution of $\bar{x}$ is well approximated by a normal distribution, even when the population is not normally distributed itself. This normal distribution can be characterized by a mean $\mu_{\bar{x}} = \mu$ and a

standard deviation $\sigma_{\bar{x}} = \frac{\sigma}{\sqrt{K}}$ where μ and σ are mean and standard deviation of the population and K is the size of the sample. From (8.12), it is known that FFT introduces phase rotations and summing of N random variables. If the random variable of the real part of the FFT output is F , and the random variable of the real part of the noise variable is y , then F can be regarded as a sample in the population,

$$F = \sum_{i=1}^N y_i = N \cdot \frac{1}{N} \cdot \sum_{i=1}^N y_i = N \cdot \bar{y} \quad (8.13)$$

where $\bar{y}$ is the random variable for the sample mean and here the population is the real part of the noise variable y . Assume the y is characterized by its mean μ_y and standard deviation σ_y . According to the Central Limit theorem, the distribution of $\bar{y}$ follows the normal distribution with a mean $\mu_{\bar{y}} = \mu_y$ and a standard deviation $\sigma_{\bar{y}} = \frac{\sigma_y}{\sqrt{N}}$. Then, from (8.13), it can be easily derived that the random variable F will also follow the normal distribution with mean $\mu_F = N \cdot \mu_y$. Since the noise's phase is uniformly distributed, $\mu_y = \mu_F = 0$. The variance of F can be obtained by

$$\sigma_F^2 = E[(N \cdot \bar{y})^2] - E[(N \cdot \bar{y})]^2 \quad (8.14)$$

where $E[\bullet]$ is the expected value of the variable. The last term in (8.14), $E[(N \cdot \bar{y})]^2 = \mu_F^2 = 0$, so that

$$\sigma_F^2 = E[(N \cdot \bar{y})^2] = N^2 \cdot E[\bar{y}^2] = N^2 \cdot \sigma_{\bar{y}}^2 = N \cdot \sigma_y^2 \quad (8.15)$$

and the standard deviation of F

$$\sigma_F = \sqrt{N} \cdot \sigma_y \quad (8.16)$$

Because the signal's amplitude is also scaled up by a factor of $\sqrt{N}$ after passing through the FFT, the average SNR remains unchanged after the FFT operation.

As a result, the performance of the OFDM system under the power line noise will just be the same as that under AWGN. The BER of the OFDM-BPSK system can be given by [70]

$$p_e|_{\text{OFDM-BPSK}} = Q[\sqrt{2 \cdot \text{SNR}}] \quad (8.17)$$

where $Q[\bullet]$ is the Marcum-Q function.

To verify the above conclusion, the performance of an OFDM-BPSK system, whose block diagram is shown in Figure 8.5, is simulated. The BPSK modulator and IFFT make up the

receiver. The comparison of the error performances under different power line noises against that under AWGN confirms the above deduction. In the OFDM system, the power line noise is further randomized by IFFT and its axis-PDF will be close to Gaussian distribution. Because of this, the performance of the OFDM system is much better than those non-OFDM systems as shown by the comparison between Figures 8.4 and 8.6. The OFDM system will not be affected by the different noise PDFs (according to m value) on the sub-carriers as it only depends on the noise power encountered by the sub-carriers.

Figure 8.5. Block Diagram for OFDM-BPSK Modulation System

Figure 8.6. Comparison of Simulated and Modelled BER under Nakagami Noises of Different m Values in OFDM-BPSK System

8.3 SYSTEM OPTIMIZATION

Based on the derivations and analysis from the previous section, the error performance of the OFDM-QPSK system under different m and Ω profiles in residential house (shown in Figure 7.5 (b)) can be studied as shown in Figure 8.7. Due to the non-uniform profiles of m and Ω at different sub-carriers, the performance of the system degrades. To achieve 10^{-5} error rate, the system with Nakagami-m power line noise requires about 2.2 dB more than that under AWGN.

Figure 8.7. Comparison of BERs under Residential House Noise (Nakagami Noise) and AWGN for OFDM-QPSK System in Non-fading Channel

However, this performance is only for system in a non-fading channel. If the frequency selective fading is considered in the channel, the performance will be even worse. In this case, the multi-path effects of the channel will cause strong Inter-Symbol Interference (ISI). To avoid ISI, the Guard Interval (GI) is needed. The GI allows time for multi-path signals from the previous symbol to die away before signal of the following symbol is gathered. The most effective guard interval is a cyclic extension of the symbol or Cyclic Prefix (CP), which is the one used in the following analysis. A mirror image in time, of the end of the symbol waveform is put at the start of the symbol as the guard interval. This effectively extends the length of the symbol, while maintaining the orthogonality of the waveform. Using this cyclic extended symbol the samples required for performing the

FFT (to decode the symbol), can be taken anywhere over the length of the symbol. This provides multi-path immunity as well as higher tolerance against symbol time synchronization error.

Figure 8.8. Block Diagram for OFDM-MQAM Modulation System in the Frequency Selective Power Line Channel

Figure 8.9. Comparison of Error Performances of OFDM Systems with Different GIs

Figure 8.8 illustrates the block diagram for studying an OFDM system that uses M-Ary Quadrature Amplitude Modulation (MQAM) for each sub-carrier. The channel noise model used is the modeled Nakagami-m distributed noise with m and Ω profiles shown in Figure 7.5 obtained from measurement conducted in a residential house. Note that BPSK

can be regarded as 2QAM and QPSK can be regarded as 4QAM. The CP is added to the transmitted signal in order to compensate the channel phase distortion. A frequency domain equalizer is also used to reduce the ISI. By using the power line channel as shown in Figure 8.1, the performances of the OFDM system under different GIs are plotted and compared in Figure 8.9. Without using GI, the strong ISI will cause an interference floor and even when the SNR is increased, the bit error rate will still remain large. When the GI is used, the ISI floor is reduced and the error rate will drop with the increasing of SNR. When the GI is made equal to $0.4\mu\text{s}$, which is the preferred channel length as described in Section 8.1, the system achieves its best performance. However, comparing to the performance of the non-fading channel, the system requires about another 25.5dB to achieve a BER of 10^{-5} .

To further improve the performance of the system, adaptive modulation method needs to be used. The basic concept here is to assign those carriers operating at frequencies where the channel has low attenuation and low noise with more power and more bits. At the same time, those carriers that have strong attenuation and strong noise are loaded with less power and less bits. Currently, there are several ways to achieve the aforementioned adaptive modulation [73-77]. For the sake of simplicity and efficiency, the bit removing method [77] is applied in this research.

From [78], the relationship between the number of bits in a carrier and the power needed to transmit that number of bits by uncoded QAM for a specified BER is given by the following expression,

$$b_k = \log_2 \left(\frac{3B_k \cdot g_k}{U \cdot N_k} + 1 \right) \quad (8.18)$$

where b_k is the number of bits in carrier k , B_k is the power required in carrier k to transmit b_k bits, and $\frac{g_k}{N_k}$ is the attenuation to noise ratio in carrier k , which in this case can be provided by the channel model. U is given by

$$U = \left[Q^{-1} \left(\frac{P_e}{N_e} \right) \right]^2 \quad (8.19)$$

where $Q^{-1}(\bullet)$ is the inverse of the Marcum-Q function, P_e is the desired BER, which is chosen as 10^{-5} in the following simulation, N_e is the number of nearest neighbor constellation point which can be calculated as

$$N_e = 4 \left(1 - \frac{1}{\sqrt{2^{b_k}}} \right) \quad (8.20)$$

(8.18) can be inverted to express the power needed to transmit b_k bits,

$$\begin{aligned} B_k [(b_k) \text{ bits}] &= (2^{b_k} - 1) \cdot \left(\frac{U \cdot N_k}{3g_k} \right) \\ &= (2^{b_k} - 1) \cdot \alpha_k \end{aligned} \quad (8.21)$$

so that the power saved by removing one bit from carrier k is given by

$$\begin{aligned} \Delta B_k &= B_k [(b_k) \text{ bits}] - B_k [(b_k - 1) \text{ bits}] \\ &= \left[(2^{b_k} - 1) - (2^{b_k - 1} - 1) \right] \cdot \alpha_k \\ &= 2^{b_k - 1} \cdot \alpha_k \end{aligned} \quad (8.22)$$

Figure 8.10. Bit and Energy Distribution under the Adaptive Modulation

Based on (8.22), adaptive modulation can be performed according to the characteristics of the power line channel. Firstly, all the sub-carriers in the system are filled with 64QAM.

Then according to (8.22) and the modeled $\frac{g_k}{N_k}$ ratios for all the sub-carriers, one bit is removed from the sub-carrier that has the highest saved power. This bit removing process continues until the total power is less than the power budget or the total bit number is equal to the bit number target. By using the constraint that the total bit is equal to the case when OFDM-QPSK is used, which is $64 \times 4 = 256$ (bits), the bit distributions along the sub-carriers for the channel shown in Figure 8.1 is illustrated in Figure 8.10.

By using the bit and energy distribution described above, the performance of the adaptive OFDM system in a fading power line channel is again simulated and the BERs of the system with and without adaptive loading are compared in Figure 8.11. As shown, the adaptive loading improves the error performance greatly. In this case, about 10dB less SNR is required by the system to achieve a BER at 10^{-5} .

Figure 8.11. Comparison of Error Performances of OFDM Systems with and without Adaptive Loading

Chapter 9

CONCLUSIONS AND RECOMMENDATIONS

9.1 CONCLUSIONS

As PLC becomes a candidate in the family of broadband communication systems, it has attracted more and more interests from the market. To provide an overview of the current status of the broadband PLC, different organizations, standards, and various technologies used in the current PLC industry have been introduced. In addition, the three major PLC applications namely, control and automation, access, and home networking, are discussed and evaluated with consideration of other competing technologies. Furthermore, based on the current PLC technological constraints and the structure of the power distribution network in Singapore, a suitable PLC application for Singapore residential buildings, intro-building access and networking, is suggested. With the understanding that a well adapted communication system can only be achieved with good understanding of the communication channel, the objective of this research is focused on the modeling and characterizing of the LV power line channel at the high frequency band.

The characteristics of LV power line channel have been introduced in terms of its transfer function, input impedance, and noise interference. Specifically, the classifications of these channel parameters have been described and the corresponding causes have been discussed. Moreover, detailed literature reviews on the methods and techniques for modeling these channel parameters are discussed. These include the network modeling techniques for transfer function and input impedance, and various techniques for modeling the background and impulsive noise. Based on these reviews, new modeling approaches for the three channel parameters have been proposed. The analytical approach of network modeling, which is based on a frequency-domain algorithm is used to model the channel transfer function and input impedance. The background noise and the impulsive noise are also modeled by frequency-domain approach.

The first phase in the network modeling of transfer function and input impedance is the line modeling. Here the power line is analyzed as a MTL and based on the fundamental Maxwell's equations, the two MTL equations that describe the relationships among line voltage, line current, and per-unit-length parameters, are obtained. With the help of the MTL equations, the two intrinsic line parameters, the characteristic impedance (Z_0) and the propagation constant (γ), are derived. These two line parameters serve as the basis of the proposed model and they are used as the model parameters throughout this research.

It is known that to determine Z_0 and γ , the four per-unit-length parameter matrices of $\mathbf{R}$, $\mathbf{L}$, $\mathbf{G}$, and $\mathbf{C}$, need to be determined first. Detailed derivation of these four primary line parameters have been described and measurements have been conducted to verify and confirm the derived line parameters.

By making the two intrinsic line parameters as the model parameters, a network model for tree topology network is later derived. Specifically, the time-domain approach based on echo model and the frequency-domain approach based on S-parameters are described. The impedances of some typical household appliances (without power cords) over the PLC operating frequency band are obtained through measurements. With these measured impedances, the derived channel model is verified using a simulated test network. It is found that the frequency-domain model can provide a more accurate prediction of the transfer functions and the input impedances than the time-domain model, especially when the network configuration is complex. Due to this advantage, the frequency-domain model is adopted throughout the entire research. To show the applicability of the model, the model error sensitivities towards various network configuration parameters have been studied. These parameters include size of cable conductor, distance between cable conductors, length of direct signal path, and length of load branches. It is concluded that: (1) Both transfer function model and input impedance model have relatively low sensitivities to the conductor size. (2) The length of direct signal path does not cause significant change to the transfer function but it does have strong influence on the input impedance. (3) Both models show relatively high sensitivities to the spacing between cables and the length of the cable branches.

Since many of the electrical power socket wirings in residential houses, commercial and industrial buildings are laid in ring circuits, this research also proposed a technique to model the transfer function and the input impedance for the ring topology power network. Based on the S-parameters of the paralleled tree circuits that form the ring network, the proposed modeling technique takes into account all of the wave attenuations, reflections and interactions in the network. Verification measurements have been conducted and the results show that the proposed approach is accurate. In addition, the idea of extending the proposed ring model to model a more practical PLC channel whose transceivers are not located on the same ring circuits is also described.

A noise model is proposed to determine the noise spectrum at the receiver so that its effects can be considered in the design of PLC transceivers. Both the background noise and the impulsive noise are analyzed in the frequency domain. Based on two long-term measurements conducted in a residential house and in a laboratory, the amplitude spectrum of the power line background noise at each frequency is found to follow the Nakagami-m distribution. The comparison of measured PDF and the modeled PDF validates this finding. On the other hand, to model the total appliance noise at the receiver, the noise amplitude spectra generated by different house appliances are measured. Then the previously derived transfer function of the channel model is used to obtain the total appliance noise reaching the receiver. In total, the noise at the receiver is determined by the summation of the background noise and the total appliance noise reaching the receiver. The derived receiver noise model has also been verified by measurements on test network.

According to other researchers' evaluation, OFDM seems to be the best choice for the PLC system. The derived channel model is later applied to help in designing an OFDM-based PLC system. Firstly, with the transfer function model, the impulse responses of different power networks are obtained, so that the OFDM system parameters such as symbol duration, GI, and total number of sub-carriers can be determined. Then, using the derived noise model, analytical predictions of error performance under the influence of power line noise for non-OFDM-based system and OFDM-based system are undertaken and verified. Finally, the transfer function and noise models are applied to design an adaptive modulation scheme. Simulation results show that by using the proposed adaptive modulation method, the error performance of the OFDM-based system can be improved significantly.

9.2 RECOMMENDATIONS

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

1. The EMI/EMC issue is often regarded as the largest obstacle on the road for PLC. In this thesis, to study the signal transmission in the power line channel, the differential model MTL model has been applied. On the other hand, it is known that due to the electrical unbalance, the conversion from differential to common mode, which are the LCL (Longitudinal Conversion Loss) or the LCTL (Longitudinal Conversion Transfer Loss), are regarded as measures for estimating the radiation. By expanding the proposed MTL model, it is possible to predict the radiation level for different power networks.
2. The scope of current developed model is still limited within the power network of a residential house. In a typical high-rise building as in Singapore, as shown in Figure 9.1, there exist many branches from the cable risers going to different floors.

Figure 9.1. Typical Power Network for High-rise Building

At each floor, from the main distribution board, it is again branched and connected to several houses. This whole network may be made of different types of cables and devices and with very complex topologies. To account for all variants, a more comprehensive model is needed. The modeling technique would

need to be extended to account for rectangular bus bars and other cable structures and laying methods. On the other hand, interferences and cross-talks will be introduced in between the phases when PLC signals are transmitted on all the three phases. Further works are needed to account for all these effects.

3. Another possible future work is to study the time behavior of the power line noise. Although in this research, the noise has been categorically characterized in the frequency domain, but if the noise, especially the impulsive noise, can also be characterized in the time domain, the noise model will be much more complete. For example, some time-domain noise parameters such as pulse duration and inter-arrival-time can be very important to the design of the PLC MAC layer protocol. Furthermore, the time-domain model will allow users to study not only the average but also the instantaneous performance of the system. As a result, certain way of combining both frequency-domain and time-domain noise models may need to be developed.
4. During the MTL analysis of the power line, there are some factors that have not been considered, such as the skin effect on the self inductance of the cable conductor, the imperfectness in the assumption that the metal conduit is a perfect conductor, and the assumption that power line matches the Heaviside's condition. From the verification results, these factors do not appear to be the prominent ones that affect the accuracy of the model. However, to be precise, the error sensitivities of these factors to the model should be studied in the future.
5. The quarter wavelength problem is encountered when measuring the characteristic impedance and the propagation constant of the power line. In this research to reduce the effect of this problem to the measurement results, only those measurement points, which are affected by the quarter wave effect, are ignored. To cover the entire frequency range, several lengths of cables are used. One good suggestion, which is useful to the readers, is to show the erroneous percentage in the measurement and list the corresponding frequency-length pairs of the measured cables.

APPENDIX

In this appendix chapter, the hardware and software designs for conducting practical measurements on power lines will be introduced. On the other hand, the complementarity for the mathematical derivation of the closed form progressive matrices summation will also be given in this chapter.

A.1 MEASUREMENTS OF PLC CHANNEL

Throughout this research, measurement plays an important role, since the measurement results not only provide an accurate description of the channel behaviors but also give a convincing reference to verify our proposed channel model. Measurement on the PLC channel is quite different from those on other communication channels. The problem of using the standard measuring instrument to perform direct measurement on the power line is that the mains supply voltage (230V) in the power line is too high for the instrument's probe to withstand, resulting in electrical damage to the instrument. This chapter explains how the abovementioned problem is overcome. Furthermore, the software tools used to acquire the required information from the instrument is also described.

A.1.1 Design of Measurement Devices

A.1.1.1 *Signal Coupler*

The electronic components in any test equipment and PLC transceiver are not able to withstand voltage level of 230V in the power line. Therefore, a signal coupler is required to interface the test equipment or PLC transceiver to the power line. The signal coupler is a coupling circuit used to inject any generated high frequency signal from signal generator or PLC transmitter onto the power line, and to extract any high frequency signal from the power line to the spectrum analyzer or PLC receiver. It blocks the 50Hz power line voltage from reaching the test equipment and PLC transceiver.

The methodology used for coupling the signal onto and from the power line is based on Differential Mode Coupling that involves the live wire as one terminal and the neutral

wire as the second terminal. Secondly, the signal coupler's operation is based on capacitive coupling. The advantage of capacitive coupling is that it offers low impedance path to high frequency communication signal but high impedance to the 50Hz power line voltage. On the other hand, to achieve the isolation between mains network and measurement equipments, a 1:1 coupling transformer is incorporated into the design. The schematic design of the signal coupler is shown in Figure A.1.

Figure A.1. Power Line Signal Coupler Circuit

The capacitive coupling here is provided by the class X2 100nF capacitor which is rated at 275V. A 1MΩ resistor is in connected parallel with the coupling capacitor to discharge the capacitor in order to minimize the hazard of high voltage shock caused by the charge stored in capacitor.

Figure A.2. Coupling Transformer Made by Toroidal Wiring

The performance of the signal coupler is dominated by the design of the coupling transformer. In our original design of the signal coupler, the transformer was made in the laboratory by using toroidal wirings around a ring shape ferrite core as shown in Figure

A.2. However, a toroidal wiring has an effective capacitance, which may be considered to be in parallel with the inductance. This is the result of the summation of capacitances from turn to turn, layer to layer, and from parts of the wiring to the core. Consequently, a parallel resonant circuit will generate a high impedance and block the signal at the resonant frequency. The solution to this resonance is to reduce the Q factor of the resonant circuit or shift resonant frequency out of the frequency range of interest. In this design, the coupling transformer (ADT1-1WT) from Mini Circuit has been used to replace the handmade one. The improvement can be seen from the Figure A.3.

Figure A.3. Comparison of the Designed and the Commercial Coupling Circuits

Figure A.3 has compared the performance of the original designed, current designed and the commercial coupling circuits. It can be seen that the original design has a deep attenuation notch of about -7.5dB at around 25MHz. Although the current design still has a slightly strong attenuation at this frequency range, it has improved by 4dB. Comparing to the commercial one from a Spanish company DS2, which manufactures the present fastest PLC modem in the world, our design has better performance at most of the frequencies. From the transfer function of the commercial coupling circuit, it can be seen that it still has the resonant effects, but the resonant frequency is almost out of 30MHz, which makes its transfer function almost flat in 1~30MHz.

A.1.1.2 Line Trap

In PLC measurements, one important thing is to isolate some power devices from the communication signals if they are easily affected by the high frequency signals. This should be done without interrupting the supply of the electrical energy. To achieve these requirements, the line trap is needed.

In contrast with signal coupler, the main function of the line trap is to restrict the carrier transmission over the power line to a given line section, and to prevent the carrier signals from spilling over to adjacent sections of the network. Moreover, the line trap with high impedance termination at its both ports stabilizes line impedance as it is able to reduce the loading effect caused by low impedance loads. The line trap can also provide the noise isolation for some particular portions of the network.

The line trap will be in series with the power line and presents high impedance to the carrier band of 1~30MHz. From the circuit point of view, it is just simply a low pass filter. But it must be able to withstand high voltage (230V) and high current (up to 10A) so that the normal power supply to the house appliances is not affected. By following the steps for designing the Chebyshev filter, the line trap can be easily achieved as shown in Figure A.4.

Figure A.4. Power Line Trap Circuit

The class X2 capacitor is also used in this circuit. The inductor is specially ordered and has a current rating of 10A, since the normal inductor does not have such high current rating.

(a)

(b)

Figure A.5. Characteristics of Power Line Trap: (a) Attenuation Level (b) Input Impedance

The characteristics of line trap are shown in Figure A.5. The isolation level that it can provides is about 60dB and above, while the input impedance looking into the line trap is quite large since its curve concentrates on the right hand side of the Smith's chart. The characteristics of the line trap indicate that it can be used to provide isolation to communication signal and noise and impedance stabilization in the frequency band of 1~30MHz.

A.1.2 Software Tools

During the measurement, some software tools are used for data acquisition and analysis. The whole measurement procedure is illustrated in Figure A.6.

Figure A.6. Block Diagram of Measurement Procedure

Firstly, the data is measured by the measurement equipment. Then, the software HP-VEE (Herlett Packard Visual Engineering Environment) is used to acquire the data into the computer through General Purpose Interface Board (GPIB). Finally, the data file generated by HP-VEE will be processed by MATLAB [81].

HP-VEE is a virtual programming language, which is suitable for developing programs for instrument control and data acquisition. Its main advantage is its simplicity in instrument control and automation. The functions of the equipment are illustrated as blocks in the HP-VEE, so that complicated controls to the equipment can be easily formed into a flow diagram in the program. Because of this, it is very easy to conduct long-term or intermittent measurement by using HP-VEE. However, HP-VEE is lack of mathematic functions. A simple math process can become very complex in HP-VEE.

To solve this problem, MATLAB is used in the measurement to process the data acquired by HP-VEE. MATLAB is powerful in math processing and displays. It can easily load the data file generated by HP-VEE into its workspace and process. Although MATLAB itself also has the data acquisition toolbox, but comparing with HP-VEE, it is lack of many instrument control functions and it is much more expensive.

Combining the advantages of data acquisition and equipment control from HP-VEE and data processing from MATLAB, the whole measurement was made much faster and more efficient.

A.2 DERIVATION OF CLOSED FORM PROGRESSIVE MATRICES SUMMATION

To ensure that statements (6.9) and (6.19) are true, suppose matrix $\mathbf{P}$ is diagonalizable (i.e. $\mathbf{P} = \mathbf{S} \cdot \mathbf{Q} \cdot \mathbf{S}^{-1}$ where $\mathbf{Q}$ is a diagonal matrix), then $\mathbf{P}^k = (\mathbf{S} \cdot \mathbf{Q} \cdot \mathbf{S}^{-1})^k = \mathbf{S} \cdot \mathbf{Q}^k \cdot \mathbf{S}^{-1}$. So, a polynomial of matrix $\mathbf{P}$,

$$\begin{aligned} f(\mathbf{P}) &= \mathbf{I} + \mathbf{P} + \mathbf{P}^2 + \mathbf{P}^3 + \dots \\ &= \mathbf{S} \cdot \mathbf{S}^{-1} + \mathbf{S} \cdot \mathbf{Q} \cdot \mathbf{S}^{-1} + \mathbf{S} \cdot \mathbf{Q}^2 \cdot \mathbf{S}^{-1} + \mathbf{S} \cdot \mathbf{Q}^3 \cdot \mathbf{S}^{-1} + \dots \\ &= \mathbf{S} \cdot (\mathbf{I} + \mathbf{Q} + \mathbf{Q}^2 + \mathbf{Q}^3 + \dots) \cdot \mathbf{S}^{-1} \end{aligned} \quad (\text{A.1})$$

with

$$\mathbf{Q} = \begin{bmatrix} q_1 & 0 & 0 & \dots & 0 \\ 0 & q_2 & 0 & \dots & 0 \\ 0 & 0 & q_3 & \dots & 0 \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & 0 & \dots & q_n \end{bmatrix}$$

will converge when the summation $\mathbf{I} + \mathbf{Q} + \mathbf{Q}^2 + \mathbf{Q}^3 + \dots$ converges. This summation $\mathbf{I} + \mathbf{Q} + \mathbf{Q}^2 + \mathbf{Q}^3 + \dots$ and hence $f(\mathbf{P})$ will converge if and only if $|q_i| < 1$ for all i .

From the Cayley-Hamilton Theorem [78], $q_i = f(\lambda_i)$, where λ_i is the i^{th} eigenvalue of matrix $\mathbf{P}$. In other words, a necessary condition for the above convergence is that all q_i and the eigenvalues of $\mathbf{P}$ must have a modulus of less than unity. With this condition, (A.2) can be obtained. This polynomial $f(\mathbf{P})$ will converge if all eigen values of $\mathbf{P}$ have modulus of less than 1.

$$\mathbf{I} + \mathbf{Q} + \mathbf{Q}^2 + \mathbf{Q}^3 + \dots = \begin{bmatrix} \frac{1}{1-q_1} & 0 & 0 & \dots & 0 \\ 0 & \frac{1}{1-q_2} & 0 & \dots & 0 \\ 0 & 0 & \frac{1}{1-q_3} & \dots & 0 \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & 0 & \dots & \frac{1}{1-q_n} \end{bmatrix} \quad (\text{A.2})$$

$$= (\mathbf{I} - \mathbf{Q})^{-1}$$

and thus,

$$f(\mathbf{P}) = \mathbf{S} \cdot (\mathbf{I} - \mathbf{Q})^{-1} \cdot \mathbf{S}^{-1} = [\mathbf{S} \cdot (\mathbf{I} - \mathbf{Q}) \cdot \mathbf{S}^{-1}]^{-1} = (\mathbf{I} - \mathbf{P})^{-1}$$

Based on the above derivation, statements (6.9) and (6.19) will be true if all the eigen values of the matrix $\mathbf{U}$ and $\mathbf{V}$ in section IV are less than 1. This is always true because the electrical power network is primarily passive in the PLC frequency band. This means that the absolute values of all of the scattering parameters are always less than unity. As a result, the closed form of S_{21} in (6.9) and S_{11} in (6.19) will always be true.

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