Development of power line communication systems with low electromagnetic radiation

Kamarul Azran Ashari

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Development of Power Line Communications System with Low Electromagnetic Radiation

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A thesis submitted to the Nanyang Technological University
in fulfilment of the requirement for the degree of
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SUMMARY

One of the main issues raised by the widespread use of broadband power line communications (PLC) systems is the risk of interference to radio communication services caused by the generation of electromagnetic interference (EMI) emissions from the power lines over which PLC operate. Experiences across the world have led many to believe that the deliberate placement of high frequency signals on power lines by broadband PLC systems will lead to large-scale interference problems particularly in the radio frequency (RF) spectrum of 3 to 30 MHz. Current radio communication users of this spectrum include amateur radio, aeronautical and maritime communications and navigation services, broadcasting, and fixed and land mobile operations. RF communications are particularly important in most European and Asian countries for long distance communications, and some of these services have defense or safety-of-life implications.

Since the dominant radiation mode of the PLC network is common mode (CM) by nature, a feasibility study of implementing an optical coupler within the Analog Front End (AFE) module of a PLC modem to reduce CM noise coupling onto the power lines is conducted. This particular EMI mitigation technique is implemented within the PLC modem itself and not as a separate external component like the CM chokes or EMI filters. It involves changing the CM source impedance characteristic of the PLC modem by providing better galvanic isolation compared to the one provided by the isolation transformer, which is currently being used in many broadband PLC modems. In this thesis, an optically-coupled evaluation board, which is based on the conceptual idea of an optical-coupled AFE module, is designed, developed and tested in the actual power line network in a controlled environment in the laboratory. The measurement results collected show encouraging signs of reduction of CM noise coupling onto the power lines.

However, the presence of parasitic capacitances could affect the isolation capability of the optical coupler and this would in turn affect the overall effective CM source impedance of the PLC modem as seen by the power network. Therefore, characterization of the evaluation board and the respective power line network is done

using the two-current-probes approach. Based on these characterizations, the RF equivalent circuit models of the PLC system the CM suppression performance of the optical coupler as well as the effects of the parasitic capacitances between the electronic modules and the metal chassis of the modem on CM suppression are analyzed and discussed.

Finally, a full equivalent circuit model is developed to help the designer to gauge the performance of the optical-coupled evaluation board in the actual AC mains even at its earliest stage of design. Therefore, measurement and analysis of the characteristics of the indoor low-voltage distribution network (LVDN) in two residential units in Singapore are conducted to develop a reliable model to simulate the harsh conditions of the actual power line network and to estimate the CM conducted emissions generated by the optical-coupled evaluation board at the point of injection, removing the need for expensive, redundant on-site testing. The characteristics of interest include the power line unbalance level, and the variations in both the differential mode (DM or transverse) and CM (or longitudinal) impedances of the power line and the evaluation board. The two-current-probes approach is used to extract these important parameters. The equivalent circuit model is then derived by fitting equivalent resonant circuits to the measured parameters. It is shown that the equivalent circuit model developed using this technique is able to predict the CM current generated by their equipment on the power line with reasonable accuracy.

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LIST OF ABBREVIATIONS

ADC	Analog-to-Digital Converter
AFE	Analog Front End
AGC	Automatic Gain Control
BPL	Broadband Power Line
CISPR	<i>Comité International Spécial des Perturbations Radioélectriques</i> (in French) or Special International Committee on Radio Interference
CM	Common Mode
DAC	Digital-to-Analog Converter
DM	Differential Mode
DSL	Digital Subscriber Line
EMA	Energy Market Authority of Singapore
EMC	Electromagnetic Compatibility
EMI	Electromagnetic Interference
FCC	Federal Communication Commission
FTTH	Fiber-to-the-Home
HDB	Housing Development Board of Singapore
HDTV	High Definition Television
IDA	Infocommunication Development Authority of Singapore
ISP	Internet Service Provider
ITU	Telecommunication Standardization Sector
LAN	Local Area Network
LCL	Longitudinal Conversion Loss
LISN	Line Impedance Stabilization Network
LVDN	Low Voltage Distribution Network
OFDM	Orthogonal Frequency Division Multiplexing
PBTS	Public Basic Telecommunication Services
PLC	Power Line Communications
PVC	Polyvinyl Chloride
RF	Radio Frequency
TAS	Telecommunication Authority of Singapore
VoIP	Voice over Internet Protocol

Chapter 1

INTRODUCTION

1.1 MOTIVATION

In recent times, Internet access is a paramount concern for customers in the commercial, residential and educational community. According to the annual surveys conducted by the Infocomm Development Authority (IDA) of Singapore on Infocomm usage in household in Singapore between 2000 and 2005 [1], it showed that broadband adoption within the residential community alone in Singapore is widespread. The annual surveys indicate a growth by 13% in the personal computer (PC) ownership since 2000. This means that in the year 2005 nearly four in five households (73.7%) in Singapore own at least a desktop or laptop computer compared to only three in five households in 2000, with nearly two out of three of these households (64.6%) are connected to the Internet. Fig. 1.1 highlights these trends.

Fig. 1.1. Household infocomm penetration in Singapore between 2000 and 2005 [1]

The surveys further reflected a growth in broadband subscription among the household users in Singapore. Broadband access had almost doubled from 8% in 2000 to 17.7% in 2001. The number of households with broadband connectivity further experienced a significant 32% growth in 2003 with two out of five households having broadband connection compared to only about one in four households in 2002. The surveys also indicate that between the year 2003 and 2005 more than half of the nation's populations are connected to the Internet in their daily lives for work,

entertainment and educational purposes, and this figure is expected to increase over the next few years. With the increase in broadband users in Singapore, Internet applications such as online information retrieval, online banking, online job searches and applications, online entertainment and online education are fast becoming popular. Table 1.1 highlights the growing popularity of these applications among Singaporeans between the year 2002 and 2004:

Types of Internet Applications and Services	2002	2003	2004
E-Communications:			
E-Mail	95.50%	91.90%	95.20%
Chat/Instant Messaging	52.50%	47.50%	53.90%
Video Conferencing	8.30%	11.30%	18.80%
Internet SMS	30.80%	38.50%	37.30%
Internet Telephone Telephony	8.80%	12.80%	17.00%
Discussion Groups	22.80%	23.00%	28.50%
E-Transactions:			
Online Banking	27.40%	32.60%	39.50%
Online Investments/Securities	9.70%	8.40%	12.80%
Online Insurance Services	6.40%	5.00%	9.60%
Online Government Related Transactions	42.10%	42.20%	55.50%
Online Shopping	21.20%	20.00%	29.90%
Non-monetary E-Transactions			
Uploading and downloading of documents from office servers	28.40%	36.80%	52.90%
Online job searches and Applications	26.90%	31.20%	43.70%
Online Information Searches and Retrievals	59.30%	64.10%	82.30%
Online Library services	22.10%	29.90%	36.20%
Downloading of Application Software	44.90%	43.10%	61.50%
Creating of Online Personal Home Pages	13.30%	15.50%	21.40%
E-Entertainment:			
Downloading and playing online games	44.00%	48.40%	43.00%
Audio Streaming/Download and/or Upload Music	41.80%	37.10%	46.10%
Video Streaming/Download and/or Upload Video	31.10%	27.50%	32.40%
Download and/or Upload Digital photos	32.20%	34.10%	48.20%
Online Contest/Competitions	19.80%	20.90%	24.10%
E-Learning:			
E-Learning or Online Education	15.70%	29.80%	32.00%
Children Educational Content (e.g. Online Assessment)	11.30%	17.00%	21.10%
Total Internet/Broadband Users (Aged 15 years old & above)	1,374,000	1,463,370	1,723,300

Table 1.1. Usage of Internet by Singaporeans aged 15 years old and above

Within the business community, according to annual surveys conducted by IDA between 2002 and 2005 [1], the percentage of companies adopting Infocomm in year 2004 was estimated to be at a high percentage of 83.1%, which is consistent to the year 2002 and 2003, although there is a slight drop to a percentage of 79.7% for the

year 2005. Broadband usage of these companies has also increased from about 41.2% in 2002 to approximately 55% in 2005. Fig. 1.2 highlights these trends.

Fig. 1.2. Infocomm usage by Companies in Singapore between 2002 and 2005 [1]

Interestingly, the same surveys also indicate that the percentage of companies using narrowband Internet has declined from 40% in 2003 to 31% in 2005. This could be due to more and more companies shifting from narrowband to broadband Internet. The surveys also reveal that more and more companies are turning to broadband Internet as a platform for delivery of contents and services, marketing and promotion of their business activities, finding information about employment opportunities for recruitment, and for information searching and monitoring purposes. These companies generally rated “greater productivity”, “faster turnaround and time-to-market”, and “lower costs and overheads” as the top three benefits of using broadband Internet.

Thus, as global competition intensifies with the massive free flow of trade and information, the need to have access to instantaneous, up-to-date information within the residential and business communities in Singapore becomes more critical than ever. With the trends of deregulation and liberalization of the telecommunication industry, new players entering the market are beginning to explore how they can compete vigorously and effectively with the existing telecommunication companies in providing a full range of basic telecommunication services nation-wide especially in broadband services, which do not fall under the monopoly rights granted to the dominant Public Basic Telecommunication Services (PBTS) operator, the Singapore Telecom (SingTel). With the introduction of facilities-based competition before service-based competition by the Telecommunication Authority of Singapore (TAS)

in 2000 [1], service providers are expected to invest in the construction and development of high-quality and advanced telecommunication infrastructure, which would in turn provide the necessary foundation to generate a nationwide broadband-ready network and thus, enhancing Singapore international broadband connectivity.

1.1.1 Broadband Power Line Communication Systems

Broadband Power Lines (BPL), also known as Power Line Communication (PLC), is the use of the technology to provide broadband Internet access through ordinary power lines. The system can transmit voice and data by superimposing modulated analog signals over the standard 50 or 60 Hz alternating current. A computer (or any other communication device) only needs to plug a PLC modem into any AC power outlet in an equipped building to have access to high-speed Internet. BPL offers obvious benefits over regular coaxial cable or digital subscriber line (DSL) connections in such a way that the availability of extensive power line infrastructure allows more people in more locations that have access to the power supply to have access to the Internet as well. Also, such ubiquitous availability makes it much easier for other consumer electronics equipment, such as video or audio systems, to be connected to the network readily.

Therefore, with the introduction of the PLC system into the telecommunication industry, the local distribution will inevitably be a very competitive arena, where many utilities could take advantage of their vast, in-place wiring infrastructure to gain competitive advantage by offering new broadband Internet services in addition to electric power and to build brand recognition. With the proper technology, the infrastructure can also serve as a high-speed (more than 10 Mbps) communications medium over the 'first hundred feet' to the residence. With the PLC system, electrical utilities can supply both electric power and a pipe for high-speed, reliable communications traffic, including Internet access. Utility companies can further enhance their utility services in providing high-speed 'last mile' connectivity to Internet Service Providers (ISPs), which do not own the necessary infrastructure to reach their customers. For new non-dominant telecommunication (telcos) entrants, the issue of local access - the 'last mile' connectivity to customers' homes or offices - remains the key limitation. For example, the largest telecommunication service

provider in Singapore, SingTel, which owns most of the last-mile infrastructure to end-users, currently dominates the local broadband market. Therefore, all conventional ISPs have to rely on SingTel for that 'last mile' access to the customer's premises. The PLC technology provides new telcos a 'last mile' alternative infrastructure without having to incur heavy expenses for digging and laying additional cables. They will find PLC a more convenient and less expensive option as compared to the traditional laying of telecommunication wires to their customers' premises.

1.1.2 Broadband Power Line Communication Trial in Singapore

PLC made its debut in Singapore in the form of a one-year technology trial conducted by SP Telecom, a newly formed subsidiary of the Singapore Power Ltd, at the Singapore Polytechnic campus in November 2000 [2]. Initially, an access speed of 800 kbps was achieved and this was subsequently upgraded to 2.25 Mbps. After the successful trial at the Singapore Polytechnic campus, the Infocomm Development Authority (IDA) of Singapore launched a further 6-month commercial trial for PLC technology [3]. In the trial, SP Telecom, in partnership with two established ISPs provided 500 users an internet access speeds of up to 45 Mbps data rate (27 Mbps downstream and 18 Mbps upstream), which is more than 50 times faster than the current broadband capability. Fig. 1.3 shows a typical distribution system in Singapore where the outdoor PLC system is to be implemented.

Fig. 1.3. Simplified structure of the Low Voltage Distribution Network in Singapore

The outdoor low voltage distribution network (LVDN) system supplies about 5 to 6 high-rise buildings, each having around 100 households, in a ring configuration. The

proposed PLC system will be implemented at the low-voltage (LV) output of the distribution transformer in the LVDN, thus transforming the LVDN into a large-scale local area network (LAN). In the proposed system, the local distribution transformer substation will be connected to the telecommunication network or the Internet Protocol (IP) backbone via conventional telecommunications infrastructure as shown in Fig. 1.4. A PLC master modem in the distribution transformer substation provides the high-speed connection from the telecommunication backbone onto the LVDN. Necessary filtering devices will be added at the substation to prevent the interference from the power network from coupling into the station communication system, and to provide isolation to ensure that distant faults do not affect the PLC system.

Fig. 1.4. The proposed outdoor PLC system

However, PLC is currently facing stiff competition from other existing telecommunication services. For example, SingTel is currently pushing for 'Fiber-to-the-Home' (FTTH) project, which aims at providing a complete island-wide network of fiber optic cables that will connect residential and commercial buildings where residents of these buildings would be able to have Internet access with broadband speeds of up to 80 Mbps for advanced voice, video and data services [4], [5]. With more residential and business communities in Singapore turning to fiber optic cables for their primary system communication needs, the PLC could still be useful as a backup channel or for very simple low-cost installations that do not warrant a fiber drop. Thus, the PLC 'last mile' could be redefined to be within the intra-building scope, where the PLC range would be limited to the indoor LVDN from the main distribution board of any residential or industrial building as shown in Fig. 1.5. All a consumer would then require is a PLC modem, to connect to the power socket for easy access of the broadband service, 24-hours a day.

Fig. 1.5. The proposed intra-building PLC system

1.1.3 Electromagnetic Interference Issues in Broadband PLC System

As PLC technology is gaining lots of attention in Europe, Asia and the United States, utilities and telcos are seeking cost-effective alternatives to bring broadband Internet services to consumers. It has been touted as one of the best alternatives to conventional broadband, and Singapore has been keen to introduce it into the local market for some time. The one-year field trial conducted by SP Telecom within the Singapore Polytechnic campus in 2000 had demonstrated the capability of PLC to function effectively and safely without compromising the power supply and telecommunications service standards. For many ISPs, SP Telecom's entry into the Internet access business means that they can be less reliant on rivals SingTel and a second licensed PBTS operator StarHub, both of which own their own last-mile infrastructure by way of telephone lines, cables and even to a certain extent wireless technology. In fact, the survey conducted by IDA on infocomm usage in households for the year 2005 [1] showed that 29% of Singaporeans who have access to a computer at home but without Internet access have cited "high access cost" as the main reason for not having Internet access at home. Table 1.2 lists the top five barriers

to the home Internet access, as ranked by the IDA survey respondents based on the significance of each of the barriers listed:

Barriers to Home Internet Access	%	Rank
High access costs	29	1
Have access elsewhere	17	2
No necessity	16	3
Lack of interests	11	4
Concern about exposure to inappropriate/harmful content	10	5

Table 1.2. Top five reasons for not having Internet access at home

The survey conducted by IDA on infocomm usage by companies for the year 2005 [1] also showed that most companies are also most concerned about the cost of implementation apart from security issues when deciding to use the Internet. Table 1.3 lists their top five concerns:

Barriers to Usage of Internet	Rank
Security concerns (e.g. effects of hacking and viruses)	1
High expenses of development and maintenance of company web sites	2
Technology is too complicated	3
High data communication expenses	4
Loss in amount of working time due to time spent on irrelevant surfing	5

Table 1.3. Top five barriers to Internet usage by companies

The key advantage of PLC is that all the power lines are already laid, with power sockets in every room to function as a communication point for the users. It is thus a cost-effective solution. However, the power line network is not intentionally designed as a communication medium. It is an unbalanced signal line with mismatches due to stubs, switches, junctions, outlets, etc. Because of the unbalanced nature of the power line network, the high frequency broadband PLC signal transmitted through the power line results in significant amount of common mode (CM) current being generated and propagates throughout the power line network over which they operate, which causes significant level of electromagnetic field radiation. Currently, both IDA and another regulator, the Energy Market Authority (EMA) of Singapore, are concerned with the electromagnetic compatibility (EMC) issues of PLC and the risk of interference to the radio communications and wireless services. The measurement data collected from

many field trials conducted [6]–[8] in the United States, Europe and other Asian countries to quantify the interference potential in the frequency range of 3 to 30 MHz, have shown significant level of radiated emissions from the PLC network. As a result, commercial deployment of broadband PLC has been delayed [9] ever since, even though SP Telecom had expressed commitment to the launching of the PLC service by the end of 2002.

1.1.4 EMC Standards and Regulations for PLC Systems

As power line cables are generally untwisted and unshielded, they could act as efficient antennas and radiate significant amounts of radio energy. On the other hand, the PLC systems are also at risk of being interfered with external radio signals. Electromagnetic compatibility of wireless services and PLC is thus a bi-directional problem where radiation from one network affects the other. While isolating PLC system signals from the background or surrounding noises can be achieved by appropriate filtering techniques at the receiver side, it will be necessary to limit the level of radiations from the PLC system itself so that the primary users of wireless services remains unaffected. Many international standards are developed to control radiated emissions below 30 MHz in the short-wave user spectrum. These standards, however, vary from one country to another since the power distribution networks where the PLC system operates have different configurations in different countries especially between the United States and Europe. Thus, the EMC regulations developed for one country may not be suitable for the other. The regulating administration for telecommunications and posts (RegTP) in Germany has adopted the NB30 [10] for the radiation of telecommunications services in the frequency range 9 kHz to 3 GHz. The Radio Communications Agency in the United Kingdom has drafted the MPT1570 Part C [11] as its future standard for telecommunication systems operating in the frequency range of 1.6 MHz to 30 MHz. If one were to make a comparison between the standards in these two countries, as shown in Fig. 1.6, it is clearly observed that the limit proposed by UK's MPT1570 is generally about 20 dB below the current German RegTP's recommendation. Under such a strict regulation, it will not be justifiable for utilities owners to invest money and resources in upgrading their facilities to provide Internet services nationwide through PLC technology.

Fig. 1.6. Standards limits comparison for radiated measurements

Further comparison with the FCC Part 15 [12] regulations in Fig. 1.6, which is currently adopted by the United States for its PLC system, further highlight the disadvantages imposed by the regulations adopted by both Germany and UK. Therefore, it is hardly surprising that the development of high speed PLC equipment is enormously more progressive in the United States than anywhere else in Europe. With major standardization organizations for PLC system such as the HomePlug Powerline Alliance adopting the FCC Part 15 as their EMI regulatory standards, development of PLC systems capable of processing and handling data rates as high as 10 Mbps as specified in the Homeplug Specification version 1.0 are no longer impossible in the United States. In fact, the HomePlug Powerline Alliance has recently announced in 2005 that its specification for the next version, HomePlug AV (Audio/Video) [13], which is designed for transmission of the High Definition Television (HDTV) data and the Voice over Internet Protocol (VoIP) has finally been approved unanimously by the board of directors of the alliance. If successfully implemented, the HomePlug AV would accelerate the development of PLC systems capable of processing and handling data rates as high as 100 Mbps or more. On the other hand, with the current stringent regulations imposed in both Germany and UK, such a rapid progress for high-speed PLC networking seems doubtful even for PLC systems with data rates of only 1-2 Mbps. Since the broadcast, military, and commercial community, as well as licensed amateur radio services, in Germany and UK seriously objected to a nationwide implementation of PLC, efforts to raise the limits towards the US values will not happen - at least not for now. Hence, network

conditioning towards a more symmetrical network and EMI rejection will be an important issue for further exploitation of the PLC channel capacity if the regulatory standards in respective countries are to be maintained.

1.1.5 EMC Research Activities related to Broadband PLC System

Many research activities have been carried out to characterize the radiated emissions associated with the usage of PLC in power line networks, ranging from theoretical characterizations to experimental assessments. The objectives of these research activities not only sought to understand the behavior of radiated emissions from the PLC network but also aim to contribute to the regulatory harmonization of the electromagnetic interference and compatibility (EMI/EMC) standards in broadband PLC usage. They are also driven largely by the desire to reduce the testing burden for the affected PLC products by removing the need for expensive, redundant testing. PLC modems are often designed to transmit signals at reasonably high voltage in the frequency range of 3 to 30 MHz in order to compensate transmission losses in the power lines but this incur the risk of radiating large amounts of radio energy. Thus, for most PLC modem manufacturers, it would be important to have clear and easy-to-measure limits of voltage to be fed into the power lines. Most regulations, however, do not care for this injected voltage, but the possible radiation emitted. In order to bridge this gap, field tests have been performed [14]–[17] to investigate the relationship between the voltage levels of data signals fed into the power line network and the resulting levels of radiated electric and magnetic fields. A local transfer function, frequently referred to as the “mains decoupling factor” [18], has been proposed to define such a relationship. Unfortunately, this method of characterization of electromagnetic fields could only provide estimation for the level of EMI radiation emitted by any arbitrary symmetrical input signal injected into the power line network but is unable to provide further useful information, such as the effects of the different network configurations, cabling arrangements, method of signal coupling, and the varying environmental conditions on radiated electromagnetic fields. Although it is well known that the radiation mechanism from the PLC network originates from the undesired high-frequency signals or noise current generated by the system and propagating throughout the network, no in-depth comprehensive study has been carried out so far to understand the mechanism behind the generation of these

emissions from the PLC network. Hence, such study is crucial as it provides essential information for effective EMI reduction from the power line network.

In many field tests conducted, the radiated emissions measured are often dominantly caused by the common mode (CM) current rather than the differential mode (DM) current. Several contributions have been brought forward to explain the behavior of the CM currents from the power networks, ranging from the theoretical characterization to experimental assessment. It has been suggested that the electrical unbalance of cables or equipment in the low-voltage distribution network (LV DN), which provides the ‘last mile’ connectivity from the ISPs to their customers, plays an important role in the conversion from the DM signals injected into the power network into CM currents [19] (also described as longitudinal or asymmetrical signal). Therefore, two limits would have to be imposed on the design of any PLC modem: a limit to the DM input signal voltage and a limit to any additional unbalance introduced by the LV DN itself. To derive a suitable limit for the DM output of the PLC modem, the level of conversion from the DM voltage to CM voltage by the LV DN should be determined. The Longitudinal Conversion Loss (LCL) value, which is measured as the ratio between the DM signals injected into a wire pair network and the portion of these signals that are converted to CM signals, has been proposed by the Telecommunication Standardization Sector (ITU-T) specifications [20], [21] to describe how unbalanced a network is. A special probe by MacFarlane [22], [23] has been designed for measuring the LCL value of two-wire telecommunication networks and the associated devices connected to them. This probe is widely used and has been recommended by several PLC research groups [24]–[26] who have found it useful for measuring the electrical unbalance of LV DN. In fact, the International Radio Committee on Radio Interference of the International Electrotechnical Commission (CISPR-IEC) has proposed for the usage of the LCL values for the development of common standard guidelines in the EMI/EMC measurements in LV DN [27]–[29] for PLC usage. However, experimental data presented at a recent CISPR Subcommittee I Working Group 3 [30] meeting left serious doubt about whether the LCL measurements can be used to predict the level of CM current generated in LV DN by the usage of PLC modems. Hence, due to reliability issues, the probe has not been fully adopted for LCL measurement of AC mains line. There were some attempts to analyze the unbalance characteristics of an in-house power line network based on

numerical methods [31]–[35] in order to establish a relationship between the network balance performance of the network and the generation of CM currents. However, unlike DM signals in the power lines network, which have well-defined signal propagation paths, the propagation path of CM signals in power lines are far more complex and do not have well-defined signal paths. Hence, these attempts have not yet given reliable results and there is not much significant step forward in introducing effective solutions to minimize EMI emissions from broadband PLC systems.

1.2 OBJECTIVES

The project is initiated to investigate and develop new CM current suppression techniques, which are to be implemented in PLC modems, to limit the generation of CM currents at the point of interface between the PLC modem and the LVDN. This would reduce the level of the CM currents that propagates through the power line cables and thus reducing the level of radiated emissions from the LVDN.

There are two objectives in this research project:

The first objective focuses on the feasibility study of providing high CM attenuation within the Analog Front End (AFE) module of the PLC modem. In the feasibility study, two CM suppression methods, one using an optical coupler and another using a CM choke, are investigated.

The second objective looks into the characterization and modeling of the CM noise propagation path of the indoor power line network for residential buildings in Singapore so that CM currents generated by the PLC modem designed can be estimated. Design guidelines can then be imposed on the PLC modem developed.

1.3 MAJOR CONTRIBUTIONS OF THE THESIS

The major contributions of the thesis are as follows:

1. Using a two-current-probes measurement methodology, the CM and DM impedances of the power line networks and the PLC modems under high-

voltage conditions are measured. The measured results enable the development of the DM and CM propagation models of the PLC system while it is in operational mode to assist in the analysis of CM current in the power line network. The measurement methodology can also be configured to measure the level of electrical unbalance of power line networks accurately.

2. A working prototype of an Analog Front End (AFE) evaluation board that is capable of emulating power line interface functions similar to the AFE module in the PLC modem, is designed and developed. With the help of the evaluation board, a feasibility study of implementing an optical coupler within the AFE module as a potential candidate for providing good isolation against CM currents is conducted.
3. The full equivalent circuit model, which describes the relationship between the output signal voltage of the PLC modem, the unbalance of the power line network and the resultant CM current generated, is derived to estimate the level of CM current generated by the modems at the signal interface. This model is useful for evaluating the effectiveness of the CM suppression technique implemented in a PLC modem, and also to determine the suitable limit for the level of the DM output signal.

1.4 ORGANISATION OF THE THESIS

The thesis is organized into six chapters.

Chapter 1 introduces the research work, its motivation and objectives.

Chapter 2 details the theoretical background and the validation of the two-current-probes approach.

Chapter 3 describes the feasibility study of implementing a CM choke in the AFE module of a PLC modem for the reduction of CM current from the power lines without compromising the DM transmission signal in the RF range of 3 to 30 MHz. The development of the CM and DM equivalent circuit models using the two-current-

probe measurement approach is described. These models provide insight into the effects of CM and DM impedances of a PLC modem on the CM noise suppression performance of the CM choke.

Chapter 4 investigates the effectiveness of the optical coupler, implemented in an AFE evaluation board, on the CM current suppression. It also addresses the proper design of the evaluation board to ensure that the optical coupler provides effective isolation against CM currents. The effects of the existence of parasitic capacitances between the electronic modules and the metal chassis of the modem on the CM suppression capability of the optical isolator are also studied. Based on the characterization of the evaluation board and the power line network, the high frequency CM equivalent circuit models are established to allow such studies to be conducted.

The development of a full equivalent circuit model, which describes the relationship between the injected output voltage of the PLC modem, the unbalance of the power line network and the resultant CM current generated, is described in details in Chapter 5. A mathematical expression is also derived to explain the mechanism behind the generation of CM current generated by a PLC modem. The full equivalent circuit model is used to predict the level of the CM current generated at the signal interface of the optically-coupled evaluation board developed in Chapter 4 when it is plugged onto the power outlets in the residential buildings in Singapore. The measurement and characterization of the unbalance properties of the indoor power line network in typical residential units in Singapore for the development of the full equivalent circuit model is also described.

Finally, Chapter 6 concludes the work done and recommends possible future work. The information will be useful for other researchers who wish to conduct further researches in the areas EMI/EMC design for PLC system.

Chapter 2

TWO-CURRENT-PROBES MEASUREMENT APPROACH

Due to high voltage (230Vac) of the power line, direct impedance measurement on the power line using the standard measuring instrument is prohibited. To overcome this problem, a two-current-probes measurement approach is adopted. Using one current probe as an injection probe and the other current probe as a receiving probe, the common mode (CM) and the differential mode (DM) impedances, in the radio frequency (RF) range of 3 to 30 MHz, for any power line networks with connected loads could be measured with ease. With careful calibration of the measurement technique, good accuracy can be achieved for measuring both CM and DM impedances. This chapter describes the two-current-probes approach measurement methodology.

2.1 BASIC TWO-CURRENT-PROBES METHODOLOGY

The two clamp-on current probes measurement technique was originally employed by Nicholson and Malack [36] for DM impedance measurement of the low-voltage power distribution network at the power outlet in the frequency range of 20 kHz to 30 MHz. Kwasnoik *et al* [37] later extended the frequency range of the measurement methodology up to 500 MHz with great accuracy. See and Deng [38] adopted similar measurement methodology to characterize the CM noise source impedance of the switched mode power supply (SMPS) in the frequency range of 150 kHz to 30 MHz for optimum design of electromagnetic interference (EMI) filters.

Fig. 2.1 illustrates the basic setup of the two-current-probes approach to measure any unknown impedance Z_X . The measurement system consists of an injecting current probe, a monitoring current probe and a network analyzer. The two current probes and a decoupling capacitor C form a high-frequency coupling circuit to avoid direct connection between the network analyzer and the high voltage power mains. A pair of short wires with length d from the reference plane b' to the reference plane b connects

the coupler to Z_X . Port 1 of the network analyzer induces a continuous wave (CW) signal in the closed loop through the injecting current probe. Port 2 of the network analyzer measures the generated current propagating in the closed loop through the monitoring current probe.

Fig. 2.1. Two-current-probes measurement methodology

Fig. 2.2. Equivalent circuit of the injecting current probe

The injecting current probe can be modeled as a transformer equivalent circuit as shown in Fig. 2.2 where V_{p1} and Z_p are the source voltage and the source impedance of the injecting signal source from port 1 of the network analyzer, respectively. With the injected signal, the injecting probe induces a voltage V_w , which results in a current I_w circulating in the closed loop. L_p , L_w , and M are the primary self-inductance of the probe, the self-inductance of the wire in the probe volume, and the mutual inductance between the probe and the wire, respectively. V_{p1} and V_w can be expressed as follows:

$$V_{p1} = (Z_p + j\omega L_p)I_p - j\omega M I_w \quad (2.1)$$

$$V_w = -j\omega M I_p + j\omega L_w I_w \quad (2.2)$$

By combining (2.1) and (2.2) and eliminating I_p , the induced voltage V_w can be expressed as:

$$V_w = Z_{M1} I_w - V_{M1} \quad (2.3)$$

where

$$Z_{M1} = j\omega L_w + \left(\frac{(\omega M)^2}{Z_p + j\omega L_p} \right) \quad (2.4)$$

$$V_{M1} = V_{p1} \cdot \left(\frac{j\omega M}{Z_p + j\omega L_p} \right) \quad (2.5)$$

Equations (2.4) and (2.5) suggest that the injecting current probe can be represented at the reference plane a by a Thevenin equivalent circuit as shown in Fig. 2.3, where Z_{M2} is the reflected impedance seen by the loop due to the existence of the monitoring current probe. The equation for Z_{M2} is similar to (2.4) with the assumption that the electrical properties of the monitoring probe such as L_p , L_w , and M are equal to that of the injecting probe, and that the source impedance of the injecting signal source from port 1 of the network analyzer (Z_p) is equal to the load impedance of the receiving circuit at port 2 of the same network analyzer.

Fig. 2.3 Equivalent circuit for the two-current-probes setup

From (2.5), the ratio V_{M1} / V_{p1} , which is dependant on the properties of the injecting probe and its operating frequency ω , is given by

$$K_R = \frac{V_{M1}}{V_{p1}} = \frac{j\omega M}{Z_p + j\omega L_p} \quad (2.6)$$

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

$$V_{M1} = (Z_{ip} + Z_X) I_w \quad (2.7)$$

where $Z_{ip} = Z_{M1} + Z_{M2} + Z_C$. Substituting V_{M1} from (2.7) into (2.6), Z_X can be evaluated by

$$Z_X = (K_R Z_{T2}) \cdot \left(\frac{V_{p1}}{V_{p2}} \right) - Z_{ip} \quad (2.8)$$

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

$$\frac{V_{p1}}{V_{p2}} = \frac{S_{11} + 1}{S_{21}} \quad (2.9)$$

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

$$K_R Z_{T2} = \frac{Z_{ip} + Z_{std}}{\left(\frac{V_{p1}}{V_{p2}} \right) \Big|_{Z_X = Z_{std}}} \quad (2.10)$$

With $K_R Z_{T2}$ and Z_{ip} determined, the two-current-probe setup is ready to determine any unknown impedance (Z_X). After measuring V_{p1} and V_{p2} using the network analyzer, Z_X can be found based on (2.8).

2.2 VALIDATION OF TWO-CURRENT-PROBES METHODOLOGY

2.2.1 Measurement of Known Standard Precision Resistors

In the validation experiment that follows, the Tektronic CT-1 and CT-2 current probes have been chosen as the injecting and the receiving current probes, respectively. Both current probes, the Tektronic CT-1 and CT-2 are broadband types with the effective bandwidth of about 700 MHz, which is more than sufficient to cover the frequency of interest in this thesis. The Agilent 4395A Network Analyzer is employed for the S-parameters measurements. To check the validation of the two-current-probes measurement method, the following procedure is adopted:

1. Measure the impedance Z_{ip} of the two-current-probes coupling circuit setup using an impedance analyzer.
2. Connect a known standard precision resistor at the output of the coupling circuit in place of the unknown impedance Z_X , obtain V_{p2}/V_{p1} using the S-parameter measurement function of the network analyzer and (2.9), and evaluate the coefficient $K_{RZ_{T2}}$ using (2.10). To obtain the coefficient $K_{RZ_{T2}}$ for the measurement setup, a precision metal film resistor R_{sid} ($100\ \Omega \pm 1\%$) is inserted into the coupling circuit.
3. Several standard precision carbon film resistors with different values are then selected and inserted at the output of the coupling circuit individually in place of Z_X . With each of them treated as unknown impedance Z_X and with the value of the coefficient $K_{RZ_{T2}}$ known, their impedance values are evaluated using (2.8) and (2.9). Comparisons are then made with their respective stated values to determine the accuracy of the measurement methodology.

Fig. 2.4. Measurement results of high resistive value resistors

Fig. 2.4 shows that for large-resistance resistors ($120\ \Omega$, $560\ \Omega$, $1.5\ \text{k}\Omega$ and $3.3\ \text{k}\Omega$), the measured resistances are in close agreement with the stated resistive values of these resistors. For $1.5\ \text{k}\Omega$ and $3.3\ \text{k}\Omega$ resistors, the roll off above $10\ \text{MHz}$ is expected due to the parasitic capacitance that is inherent to resistors with very large resistances. In Fig. 2.4, Z_{ip} is also plotted to show its relative magnitude with respect to the measured

resistances. Based on the trend of impedance variation with frequency, the Z_{ip} of the coupling circuit can be modeled as an equivalent RLC series circuit having a $1.12\ \Omega$ resistor, a $240\ \text{nH}$ inductor and a $2.2\ \mu\text{F}$ capacitor connected in series. It can be seen that for large-resistance resistors, Z_{ip} is insignificantly small enough to be ignored from (2.8) without affecting the accuracy of the measurement.

Fig. 2.5. Measurement results of low resistive value resistors

For the resistors with small resistances, such as $2.7\ \Omega$ and $5.6\ \Omega$, the contribution of error due to impedance Z_{ip} of coupling circuit is rather significant. Hence, proper care must be taken in the measurement to ensure that Z_{ip} be subtracted from the measured impedance. Fig. 2.5 shows that, after eliminating Z_{ip} from the measurement, the measured resistances of the resistors have agreed very well with their respective stated values. Hence, Fig. 2.5 indicates clearly that Z_{ip} cannot be neglected and must be taken into account especially for low impedance values, which are comparable to Z_{ip} . Hence, by subtracting Z_{ip} , the error due to the RF coupling circuit on the measurement results can be eliminated.

2.2.2 Measurement of Impedances of the Impedance Stabilization Network

To further validate the accuracy of the two-current-probes approach, the CM and DM impedances ($Z_{in, CM}$ and $Z_{in, DM}$, respectively) of the Line Impedance Stabilizing Network (LISN) are first measured using the Agilent 4395A Impedance Analyzer and

the using the two-current-probes approach. The measurement results using the two methods are then compared. For the CM impedance measurement, the phase and the neutral wires are treated as one single outgoing conductor as shown in Fig. 2.6, and the safety ground wire is treated as the returning conductor, while for the DM input impedance measurement, the phase wire is treated as one single outgoing conductor and the neutral wire is treated as the returning conductor. For the measurement of the CM impedance and DM impedance of the LISN using the impedance analyzer, measurements are conducted with no AC power supply to prevent any damage to the measuring equipment.

Fig. 2.6. The Line Impedance Stabilizing Network

Fig. 2.7. Two-current-probes setup for the measurement of the CM input impedance (a) and the DM input impedance (b) of the Line Impedance Stabilizing Network.

Fig. 2.7 shows the setup for the measurement of the CM and DM impedances of the LISN under a high voltage condition, using the two-current-probes approach. For the measurement of the CM impedance, the two Tektronic current probes and two

coupling ‘Y’ class 0.1 μF capacitors (one for phase to ground and the other for neutral to ground) are used to form the CM coupling circuit to avoid any direct connection between the measuring equipment and the AC mains. For DM impedance measurement, the two current probes and two coupling ‘X’ class 0.1 μF capacitors (both connected between phase and neutral) are used to form the DM coupling circuit. To ensure that the setup impedances $Z_{ip,cm}$ and $Z_{ip,dm}$ of the CM and DM coupling circuit remains stable and repeatable, the current probes, coupling capacitors and connecting wires are fix-mounted on two different printed circuit boards as shown in Fig. 2.8. Both the CM and the DM coupling circuits have a rectangular loop area of $4.2\text{ cm} \times 5.5\text{ cm}$. The given loop size is still very much smaller than the wavelength of 30 MHz, the maximum frequency of interest. Therefore, the transmission line effect of the loop can be safely neglected.

Fig. 2.8. Implementation of the RF coupling circuit for impedance measurement

Both $Z_{ip,cm}$ and $Z_{ip,dm}$ are separately measured with the 4395A Impedance Analyzer. Based on the measured characteristics of $Z_{ip,cm}$, it can be modeled as an RLC series circuit with a resistor of $1.34\ \Omega$, an inductor of 143 nH and a capacitor of $0.169\mu\text{F}$. For the measured characteristics of $Z_{ip,dm}$, it can be modeled as an RLC series circuit with a $1.12\ \Omega$ resistor, a 240 nH inductor and a $2.2\ \mu\text{F}$ capacitor. With the known $Z_{ip,cm}$ and $Z_{ip,dm}$ of the coupling circuits and the coefficient $K_R Z_{T2}$, which is derived earlier in the pre-measurement calibration process, the two-current-probes setups are now ready to measure the unknown CM and DM input impedances of the LISN. The comparisons of measured results for DM and CM input impedances of the LISN using the two-current-probes approach and the Agilent 4395A Impedance Analyzer are shown in Fig. 2.9 and Fig. 2.10, respectively. The measured results, which consist of

the magnitude and phase components of the measured impedance value, show very good agreement with those obtained from the Agilent 4395A Impedance Analyzer.

Fig. 2.9. Measurement results of the DM input impedance of the LISN

Fig. 2.10. Measurement results of the CM input impedance of the LISN

The measurement results for the DM and CM input impedances of the LISN using the two-current-probes approach (where there is AC power supply) and the Agilent 4395A Impedance Analyzer (where there is no AC power supply) in Fig. 2.9 and Fig. 2.10, further show that the impedance characteristics of the LISN under these two conditions differ slightly from one another especially in the higher frequencies of 10 MHz and above. This could be due to the saturation effects of the inductive components of the LISN in high-voltage environment. High currents, which are most likely to occur due to the low impedance nature of the power lines in low frequencies, can cause significant magnetizing force that affect the effective inductance of the

LISN causing the overall impedance to decrease or even exhibit nonlinear characteristics. Hence, for accurate characterization of the PLC system, impedance measurements have to be done on the system under the actual ‘power on’ condition so as to take into account any changes in the overall impedance of the system due to the effects of high currents.

2.3 CONCLUDING REMARKS

The basic theory of the two-current-probes methodology has been described and its validity has been experimentally verified. The two-current-probes method with the model described in this chapter is proven to be accurate in measuring impedances from 0.1 MHz to 30 MHz. With the pre-measurement calibration process, the setup impedance of the RF coupling circuit, Z_{ip} , can be determined and accounted for. This would eliminate possible error contributed by the RF coupling circuit, if the impedance under measurement is small. Using the two-current-probes method, both the CM and the DM impedance measurements of any PLC modem or power line network can be performed while it is connected to the AC mains supply voltage (230Vac). This would enable us to measure and characterize the CM and DM input impedances of various power line networks and PLC modems under the actual ‘power on’ condition. Knowledge of these parameters allows the development of conducted noise propagation models in terms of conduction modes, which not only identifies the dominant mode of the noise source (DM or CM) in the power lines but also permits the design and analysis of signal coupling and EMI noise suppression techniques in PLC modems. The two-current-probes measurement methodology could also be extended to measure and characterize the levels of electrical unbalances in power line networks reliably and accurately as defined by the standards.

Chapter 3

COMMON MODE CHOKES IN PLC MODEMS

While the power line network has been viewed as a convenient and inexpensive communication medium for data and voice transmissions, its unbalanced nature leads to the generation of CM noise current in the power line network, which in turn generates EMI radiation, and interfere with existing wireless communications users. In reality, one has no control over the unbalances of the power lines caused by the electrical loads connected to them. Hence, the only possible way to control the EMI radiation is to ensure that the RF source (the PLC modem) presents very high CM impedance so that the CM current propagating through the power line network is kept to as low a level as possible. With the aim of reducing the generation of CM currents in the power line network, the author has initiated a feasibility study of inserting an inexpensive CM choke at the output of the analog front end (AFE) module of the PLC modem. CM chokes, which are sometimes frequently referred to as “CM inductors”, have been used in many EMI filter designs for both DC-DC and AC-DC switching power supplies and other power electronics applications to reduce CM emissions. Although the use of CM chokes are reported in some PLC trials [8] to reduce the EMI radiations, no formal in-depth study and systematic analysis has been carried out to study the effectiveness of a CM choke on the CM noise levels, which are mainly generated at the signal interface between the PLC modem and the power line network, and its side-effects on the transmitted DM signals.

Thus, this chapter aims to carry out a systematic in-depth study on the effectiveness of a built-in CM choke in attenuating CM noise current in the power line network without affecting the useful differential mode (DM) signals, which are transmitted over the power lines for communication purposes. The effectiveness of the CM choke in suppressing the CM currents in the power line network depends not only on the CM impedance characteristics of the choke itself but also on the CM impedances of the RF source (the PLC modem) and the terminating load (the power line network). Ideally, a CM choke would not affect the DM signal generated by the PLC modem. However, in practice, the imperfect cancellation of high frequency magnetic fluxes of the two windings of the CM choke often results in the existence of finite DM

impedance, which will have an impact on the communication signals transmitted by the PLC modem in the RF range (3 to 30 MHz). To study these effects, equivalent CM and DM electrical circuit models for the PLC modem connected to any power line network are established based on the measurement data collected from the PLC system using the two-current-probe measurement approach. These models would allow both the evaluation of the CM noise suppression performance of the PLC modem (with the built-in CM choke) and the assessment of the effects of the parasitic components of the choke on the DM signals.

3.1 COMMON MODE CHOKES

3.1.1 CM Current Propagation Path in a PLC System

While PLC modems are dedicated devices responsible for transferring data over the power lines, they are also known to generate significant levels of CM conducted emissions. Fig. 3.1 illustrates all possible CM current paths throughout the PLC system. As the parasitic elements (C_3 , C_4 and C_5) in the modem provides low impedances at high frequencies, the high frequency CM current could flow into the ground through these low impedance paths. Some of the CM current could even leak through the Local Area Network (LAN) interface connection between the modem and the PC.

Fig. 3.1. Common mode current propagation paths in a PLC system

Fig. 3.2 shows the equivalent circuit model of the PLC modem connected to the power line network at the power outlet. As with most telecommunication devices, PLC modems are normally designed and engineered to be balanced symmetrical RF

sources. Thus, the PLC modem can be represented as a balanced DM signal source of E_s with an equally balanced DM source impedance of $Z_{s,dm}$ and the CM impedance of $Z_{s,cm}$.

Fig. 3.2. The PLC modem at the power outlet (a) and its equivalent circuit (b) model

The power line network, as seen by the modem at the power outlet, can be modeled by a T-network equivalent circuit, which consists of the unbalance DM impedances of Z_{lv1} and Z_{lv2} , and the CM impedance of Z_{lv3} . It can be seen that the factors that affect the generation of the CM current are the cable unbalance of the power line network ($Z_{lv1} \neq Z_{lv2}$), which is responsible for the conversion of DM transmission signals E_s into CM voltages, and the CM impedances of the PLC modem ($Z_{s,cm}$) and the power line network (Z_{lv3}), which are responsible for the conversion of the CM voltages generated into CM currents. Since the unbalance characteristics of the power line network cannot be manipulated or controlled, increasing the impedance of the CM propagation path without disturbing the impedance of the DM signal path could reduce CM current generated at the signal interface, and that can be achieved by the implementation of a CM choke at the output of the PLC modem.

3.1.2 Implementation of a CM Choke in a PLC Modem

A CM choke, as shown in Fig. 3.3(a), is normally added to increase the CM impedance so as to effectively block and reduce the unwanted CM noises along the propagation path.

Fig. 3.3. An ideal toroidal common mode choke (a) and its implementation (b) in a PLC system

Fig. 3.3(b) illustrates the implementation of a CM choke at the output of the PLC modem. A CM choke is often produced by winding the two signal wires around a single ferrite core with equal number of turns. The choke shown in Fig. 3.3(a) has two windings on each side of the core where each of the windings is inserted between the output of the modem and the input of the power outlet into the power line network. The windings are wound such that the magnetic fluxes within the core due to DM current on both windings cancel, while magnetic fluxes due to CM current in both windings add. Hence, the CM choke would effectively functions as an inductor against the CM current and has no effect to DM current. The CM choke usually wound on ferrite core, which has a high permeability and therefore could achieve a very high inductance on a relatively small core.

The DM current, which propagates in each winding of the CM choke with equal magnitude but opposite in phase, would create fluxes within the core that are also

equal in magnitude but opposite in phase. These two fluxes would cancel each other, leaving the core in an unbiased state. Ideally, the flux cancellation would result in a zero inductance and no restriction would be imposed on the DM current propagating towards the load. Therefore, the DM current would not be affected by the existence of the CM choke. Hence, theoretically, the CM choke would be suitable for CM noise suppression on PLC networks where the DM signal remains unaffected.

Fig. 3.4. Magnetic fluxes within (a) an ideal and (b) a practical CM choke.

In reality, not all magnetic fluxes within the core of the CM choke cancel. This is due to the imperfect coupling of the windings around the core. The creation of the leakage fluxes, which do not link with all the turns, causes the windings on both sides of the core to have some form of self-inductance. This effectively acts as an inductor in series with the mutually-coupled transformer windings. In addition, the windings also have some winding resistances, which would have some effect on the transmitted DM signals. Fig. 3.4(a) and Fig. 3.4(b) compare the working principles of an ideal CM choke and a practical one, respectively.

3.2 MODELING OF POWER LINE COMMUNICATION SYSTEM

3.2.1 Measurement Setup for the Characterization of PLC System

Currently, most of the PLC modems that are available in the market are HomePlug 1.0 certified. Some of the salient characteristics of these modems are that they processed a data rate of up to 14 Mbps with an effective data rate up to 6 Mbps, they use the Orthogonal Frequency Division Multiplexing (OFDM) that allocates 84 carriers - each with a bandwidth of 195.3 kHz - within a band of 4.49 MHz to 20.7 MHz for their physical layer, and they allow LAN connectivity anywhere a power

outlet exists. These common characteristics allow easy integration with all manufacturers that are compliant with the Homeplug 1.0 standard. Therefore, all HomePlug-certified PLC modems available in the market from different vendors, as shown in Fig. 3.5, are compatible with each other, thus confirming the benefits of the existence of a common specification.

Fig. 3.5. A selection of HomePlug-certified PLC modems available in the market

Thus, the HomePlug-certified PLC modems are specifically chosen for the feasibility study. Fig. 3.6 shows the measurement setup for the study. The PLC modem is modeled in Fig. 3.6 as a combination of equivalent CM and DM signal sources. The DM signal source represents the intended communication signal transmitted by the PLC modem and the CM signal source represents the unwanted CM noise generated at the signal interface due to the unbalance of the power line network. To ensure the repeatability of the feasibility study, the PLC modem is connected to the power mains through a line impedance stabilization network (LISN) [12], [27], [39], which acted as a stable terminating load for the modem. A CM-DM discrimination network is added to the output of the LISN so that both the CM and DM emissions from the PLC modem can be discriminated and measured independently. The design and construction of the discrimination network can be found in [40], [41] and thus will not be discussed in this chapter. The PLC modem is programmed to be able to transmit and receive signals continuously in a ‘loopback’ test mode without using a receiver modem.

Fig. 3.6. Measurement test setup for the study

With the measurement setup shown in Fig. 3.6, the effects of adding a CM choke on both CM and DM emissions can be analyzed systematically. By adding a CM choke in the PLC modem, the CM and DM equivalent circuits for the PLC system in Fig. 3.6 can be represented in Figs. 3.7 and 3.8, respectively.

Fig. 3.7. CM Equivalent circuit of the PLC modem terminated with the LISN

Fig. 3.8. DM equivalent circuit of the PLC modem terminated with the LISN

The DM conducted emission measured by the LISN is the intentional communication signal and is denoted as V_{signal} . The CM conducted emission measured by the LISN is the unwanted noise and is denoted as V_{noise} . From these equivalent circuits, the CM and DM attenuations due to the insertion of a CM choke are given by (3.1) and (3.2), respectively

$$A_{t,cm} = \frac{V_{noise}}{V'_{noise}} \quad (3.1)$$

$$A_{t,dm} = \frac{V_{signal}}{V'_{signal}} \quad (3.2)$$

where V'_{noise} and V'_{signal} are the CM noises and the DM signals after adding the CM choke, respectively. The CM and DM attenuations can be derived and expressed as

$$A_{t,cm}(dB) = 20 \log_{10} \left(1 + \frac{Z_{ind,cm}}{Z_{lisc,cm} + Z_{s,cm}} \right) \quad (3.3)$$

$$A_{t,dm}(dB) = 20 \log_{10} \left(1 + \frac{Z_{ind,dm}}{Z_{lisc,dm} + Z_{s,dm}} \right) \quad (3.4)$$

It can be seen from (3.3) that the CM choke can suppress the CM noise spectrum effectively if the CM impedance of the CM choke is significantly higher than the effective CM source impedance of the PLC modem or the CM input impedance of the load, i.e. $Z_{ind,cm} > Z_{lisc,cm}$ or $Z_{ind,cm} > Z_{lisc,cm}$. From (3.4), it is also worth noting that the DM signals transmitted will also be attenuated if the DM impedance of the CM choke is significantly higher than the effective DM source impedance of the PLC modem or the DM input impedance of the load, i.e. $Z_{ind,dm} > Z_{lisc,dm}$ or $Z_{ind,dm} > Z_{lisc,dm}$.

3.2.2 CM and DM Equivalent Circuits of a CM Choke

The existence of parasitic components within the CM choke often limits its attenuation capability on CM currents. These parasitic components are mainly contributed by the inter-windings and inter-layer capacitances, and the core losses, which are due to Eddy-currents and magnetic hysteresis. The inter-winding and inter-layer capacitances reduce the effective inductance especially at high frequencies as these parasitic components often appear in the form of a capacitor connected in parallel with the effective inductance of the CM choke. The core losses appear in the

form of a dissipative element or a resistor, which is also in parallel with the effective inductance. The effects of the equivalent resistor within the equivalent circuit are dependent on the magnetic material of the core, its operating frequency, and the voltage drop across the choke. CM chokes can be modeled with the equivalent circuit shown in Fig. 3.9 [42], in the high frequency range.

Fig. 3.9. CM lumped parameter equivalent circuit of the CM choke

In this model, L_{cm} is the self-inductance of the CM choke and will play the most important role in the CM attenuation power of the choke. L_{dm} represents the choke's leakage inductance and C_{t-t} represents the inter-winding capacitance. R_w represents the winding resistance, and R_c represents the core losses. The core losses R_c has always been considered as an equivalent resistance [43]–[45] connected in series with the windings resistance R_w , but it can also be regarded as an equivalent resistance across the windings [46], [47]. The core loss equivalent resistance R_c mainly acts as the damping factor to the CM noise. The intra-winding capacitance C_{t-t} together with L_{cm} will cause resonance and will significantly affect the performance of the choke. If the equivalent circuit is decomposed into both CM and DM, the inductance for each modal equivalent circuit can be estimated as:

$$L_{ind,cm} = j\omega(L_{cm} + M) + j\omega\frac{L_{dm}}{2} \quad (3.6)$$

$$L_{ind,dm} = j\omega(L_{cm} - M) + j\omega 2L_{dm} \quad (3.7)$$

where L and M are the self and mutual inductances of the choke, respectively. Neglecting the effects of the winding conductor resistance, the CM and the DM electrical models of the choke can be represented by equivalent RLC parallel circuits

as shown in Fig. 3.10 with the parasitic parameters for each circuit clearly defined.

Fig. 3.10. The (a) CM and (b) DM equivalent electrical models of the CM choke

For the feasibility study, a 3.2 mH CM choke (model CMV4.0) from Roxburgh, which is commonly used to attenuate the CM conducted emissions in power electronics applications, is chosen. The CM and DM impedances of the choke are measured in the frequency range of 100 kHz to 30 MHz using the Agilent 4395A Impedance Analyzer. Fig. 3.11 shows the measured CM and DM impedance magnitudes of the choke. For any CM choke with a ferrite core, it is impossible to achieve perfect cancellation of the magnetic fluxes of the two windings on the same magnetic core. Thus, the non-zero DM inductance, which is effectively the leakage inductance of the choke, is resulted from such imperfect cancellation that leads to the finite DM impedance shown in Fig. 3.11. For most CM chokes, this leakage inductance dominates the DM impedance and its inductance value is usually estimated to be only 1% of the CM inductance value.

Fig. 3.11. The CM and DM impedances of the 3.2 mH CM choke

Based on the trends of measured impedance magnitudes with respect to frequency, the CM and DM equivalent circuit models of the choke can be extracted. Neglecting the effects of the winding conductor resistance, the CM equivalent circuit model of the choke can be derived using a curve fitting technique: by matching a resonant circuit to the impedance values obtained. Hence, the CM component of the choke can be represented by a parallel RLC resonant circuit with the following parameters: $R_{ind,cm} = 11.10 \text{ k}\Omega$, $L_{ind,cm} = 3.24 \text{ mH}$ and $C_{ind,cm} = 7.82 \text{ pF}$. The DM equivalent circuit model of the choke can also be represented by similar parallel RLC resonant circuit with the following parameters: $R_{ind,dm} = 44.40 \text{ k}\Omega$, $L_{ind,dm} = 27.8 \text{ }\mu\text{H}$, and $C_{ind,dm} = 3.79 \text{ pF}$. From these measurements, it can be seen that the leakage inductance is 0.73% of the CM impedance. Hence, the traditional method of estimating the leakage inductance to be 1% of the CM inductance is valid.

3.2.3 CM and DM Source Impedances of the PLC Modem

A complete model is needed to estimate the attenuation power of the CM inductor on the CM noise and to predict the effect of leakage inductance on the DM transmission signal. To estimate effects of the CM choke on the CM noise and the DM signal, the CM noise source impedance and the DM signal source impedance of the PLC modem must be measured and defined first. The two-current-probe approach is employed to determine the CM noise and the DM signal source impedances of the PLC modem. Fig. 3.12 and Fig. 3.13 show the setups used to measure the CM noise source impedance ($Z_{s,cm}$) and the DM signal source impedance ($Z_{s,dm}$) of the PLC modem, respectively.

Fig. 3.12. CM impedance measurement of the PLC modem

Fig. 3.13. DM impedance measurement of the PLC modem

Fig. 3.14. Measurement of the modal impedances of the PLC system by the author

For CM impedance measurement, the two current probes and two ‘Y’ class $0.1 \mu\text{F}$ capacitors (for live-to-earth and neutral-to-earth connections) form the CM coupling circuit to avoid any direct connection to the power mains. For the DM impedance measurement, the two current probes and two ‘X’ class $0.1 \mu\text{F}$ capacitors (both connected between the live and neutral conductors) form the DM coupling circuit. The CM and DM impedances of the network - $Z_{l,cm}$ and $Z_{l,dm}$, respectively - are first measured without the PLC modem connected to the power line network, and then the CM and DM impedances of the network - $Z_{t,cm}$ and $Z_{t,dm}$, respectively - with the modem connected to the network (and in the active transmitting mode) are measured. Therefore, the CM and DM source impedances of the PLC modem can be determined, respectively, by

$$Z_{s,cm} = \frac{Z_{t,cm} Z_{l,cm}}{Z_{l,cm} - Z_{t,cm}} \quad (3.8)$$

$$Z_{s,dm} = \frac{Z_{t,dm} Z_{l,dm}}{Z_{l,dm} - Z_{t,dm}} \quad (3.9)$$

Fig. 3.15 shows that $Z_{s,cm}$ is capacitive in nature with an estimated capacitance of 23 pF, which is contributed primarily by the parasitic capacitances that exist between the primary and the secondary windings of the isolation and power transformers in the AFE module of the PLC modem. For $Z_{s,dm}$, it can be modeled as a series RLC resonant circuit having an equivalent series resistance of 2.3 Ω , a series inductance of 0.41 μH and a series capacitance of 4.13 nF. By comparing these two impedance curves, it can be seen that $Z_{s,cm}$ is high-impedance in nature whereas $Z_{s,dm}$ is low-impedance in nature.

Fig. 3.15. CM and DM source impedances of a PLC Modem

3.3 EFFECTS OF CM AND DM IMPEDANCES OF PLC MODEM

Based on the earlier measurements and characterizations, the CM and DM equivalent circuit models of the PLC modem terminated with the LISN are shown in Fig. 3.16 and Fig. 3.17, respectively. Given the parameter values of the equivalent circuits in Fig. 3.16 and Fig. 3.17, the estimated CM and DM attenuations due to the insertion of the CM choke are calculated and plotted in Fig. 3.18.

Fig. 3.16. CM circuit model of the PLC modem terminated with the LISN.

Fig. 3.17. DM circuit model of the PLC modem terminated with the LISN.

Fig. 3.18. Calculated CM and DM attenuations due to the CM choke

Interestingly, the CM choke seems to have a larger effect on the DM signal, since the DM attenuation is higher than the CM attenuation across the full frequency band. The CM attenuation values remain relatively constant at around 10 to 12 dB in the PLC operating frequency range of 5 MHz to 25 MHz, whereas the DM attenuation can achieve attenuation of as high as 50 dB at around 15 MHz, with at least 15 to 30 dB at the other frequencies. A closer look reveals that the lower CM attenuation is expected due to the high-impedance nature of the CM source impedance of the PLC modem.

Hence, the CM choke must provide much higher impedance than the source impedance in order to have a more significant impact on the CM noise suppression. On the other hand, the finite DM impedance due to the imperfect CM choke, though its magnitude is small, would cause larger impact on the signal integrity of the DM signal because of the relatively low-impedance nature of the DM signal source impedance.

Fig. 3.19. CM noises before and after the CM choke is inserted

To further confirm the findings, the CM and DM conducted emissions are measured at the output of the LISN, with the help of CM-DM discrimination network. Fig. 3.19 shows the measured frequency domain spectrum of the CM noise, which is generated at the signal interface of the PLC modem, before and after the CM choke is inserted at the output of the PLC modem. It can be seen from Fig. 3.19 that the CM noise spectrum exhibits almost similar characteristics to the DM signal spectrum, shown in Fig. 3.20, which uses the Orthogonal Frequency Division Multiplexing (OFDM) in a band from 4.49 MHz to 20.7 MHz as defined by the HomePlug specification [48]. The CM noise measured is basically the CM voltage, which is converted from DM signal due to the unbalance of the LISN. Thus, the measurement results highlight the effectiveness of the CM choke in reducing the level of CM noise generated at the power outlet between the minimum value of 8.49 dB and the maximum value of 18.04 dB at the carrier frequencies of 8.10 MHz and 13.28 MHz, respectively, as estimated by the CM equivalent circuit model. Fig. 3.20 shows the measured DM signal spectrums before and after the CM choke is inserted to the output of the AFE module of the PLC modem. The measurement results clearly indicate the attenuating effects

of the leakage inductance of the CM choke on the level of the DM signals transmitted into the impedance network. The CM choke is found to be highly effective in reducing the level of the DM signals between a minimum value of 14.31 dB and a maximum value of 41.25 dB (at the frequency of 20.71 MHz and 14.85 MHz, respectively). Comparisons between the estimated attenuation values obtained from the DM equivalent circuit (refer to Fig. 3.18) and the measurement results also demonstrate close agreement between the two results.

Fig. 3.20. DM signals before and after the CM choke is inserted

Hence, the measurement results obtained have demonstrated the dominant effects of the leakage inductance on the DM transmission signal of the PLC modem. Since the DM source impedance of the PLC modem is significantly small compared to the DM impedance of the CM choke, which is dominated by its leakage inductance, the CM choke has managed to attenuate the DM signal spectrum even more effectively than on the CM noise.

Fig. 3.21 further compares the DM impedance of the CM choke with the DM source impedances of two HomePlug-certified PLC modems from two different PLC modem developers that are commercially available in the market. It can be seen that within the PLC broadband signal range 4.49 MHz to 20.7 MHz defined by the HomePlug specifications, the values of the DM source impedances of these modems are significantly small compared to the DM impedance of the choke. Fig. 3.22 also compares the DM impedance of the CM choke with the DM input impedances of the network under test. The effects of the different lengths of the power line cable on the

DM input impedance of the network under test are also studied. It can also be seen that within the PLC broadband signal range, the values of the DM input impedances of the network under test are also significantly small compared to the DM impedance of the choke. Hence, the DM attenuation power of the CM choke can be expected to be significantly large throughout the PLC frequency range.

Fig. 3.21. Comparison of DM impedances of the CM choke and PLC modems

Fig. 3.22. Comparison of CM impedances of the CM choke and the power network

On the other hand, the PLC modems exhibit high CM impedances that are capacitive in nature with capacitance values of 23 pF and 20 pF, respectively. It can be seen from Fig. 3.23 that these values are comparably close to the CM impedance value of the choke, which exhibits a capacitance value of 7.8 pF. Even though, the values of the CM input impedances of the network under test are also significantly small compared to the CM impedance of the choke, as shown in Fig. 3.24, the high CM

impedance nature of the PLC modems dominates the overall CM impedance of the PLC system. Thus, the significantly large values of the CM noise impedances of these modems with respect to the CM impedance of the choke has caused the CM attenuation power of the CM choke to be smaller throughout the PLC frequency range. The low-impedance characteristics of the DM impedance and the high-impedance characteristics of CM impedance of the PLC modems will be analyzed in greater detail in Chapter 4.

Fig. 3.23. Comparison of CM impedances of the CM choke and PLC modems

Fig. 3.24. Comparison of CM impedances of the CM choke and the power network

Therefore, the performance of the CM choke is highly dependent on the source impedances of the PLC modem and also to a certain extent, the input impedances of power line network. Based on these observations, the author has concluded that CM chokes are not feasible to be implemented in PLC modems.

3.4 CONCLUDING REMARKS

The feasibility of adding a CM choke in a PLC modem for CM noise suppression has been studied. Both the theoretical models and the measured results have demonstrated that the equivalent CM and DM source impedances of the PLC modem have a strong influence on how well a CM choke performs in suppressing noise emissions in the power line network. Due to the high-impedance nature of the CM source impedance of the PLC modem, there is a certain limit on how much CM noise attenuation that a CM choke can achieve. On the other hand, the finite DM impedance, which is due to the imperfections of the CM choke, has a much larger impact on the intentional DM signal because of the low-impedance nature of the DM source impedance of the PLC modem. Therefore, adding a CM choke in the AFE module of a PLC modem is not a feasible solution to reduce the CM noise in the power line network. Further work will be carried out to consider other alternatives so that large isolation barrier against CM noise can be offered without affecting the intentional DM PLC signal, for example, by replacing the existing isolation transformer in the AFE module of the PLC modem with an optical coupler unit. This will be studied in greater details in Chapter 4.

Chapter 4

REDUCTION OF CM NOISE EMISSIONS USING OPTICALLY COUPLED SIGNALING

Magnetic transformers have been commonly used in the Analog Front End (AFE) modules of PLC modems to transmit RF broadband signals onto the power line network. They are normally adequate devices for isolating the lower frequency CM noises in the power lines and at the same time isolating users from the high voltage hazards. However, magnetic transformers often have an adverse effect of coupling CM currents from the modem onto the power lines at higher frequencies. Thus, in the radio frequency (RF) range, the isolation barrier provided by the magnetic transformer is degraded due to the primary-to-secondary parasitic capacitance C_i , as shown in Fig. 4.1, which ranges from 3 pF for high-quality miniature transformers to a value as high as 100 pF for ordinary audio transformers [49].

Fig. 4.1. Equivalent circuit of a PLC modem with a magnetic isolating transformer

Thus, the isolation barrier against CM currents provided by these transformers is usually effective up to a few kHz. Therefore, alternative solutions are sought to lower the levels of CM currents sufficiently in the MHz range in order to meet the EMI regulatory limits. The author has initiated another feasibility study of implementing an optical coupler within the AFE module of a PLC modem as a potential replacement for the existing isolation transformers. If properly designed and implemented, the optical coupler could achieve a very good isolation between the user side and the power line side of the modem with the value of the bridging parasitic capacitance

reduced to as low as 1 pF [50], [51] to 0.07 pF [52]. Because of the very low value of parasitic capacitance C_i , as shown in Fig. 4.2, between the LED and transistor, very good CM isolation can be achieved between the two sides of the modem.

Fig. 4.2. Equivalent circuit of a PLC modem with an optical coupler

Some PLC modem developers have recently conducted their own feasibility study of implementing an optical coupler [53] in the AFE module of their PLC modem for broadband usage. The measurement results obtained have demonstrated the potential effectiveness of the optical coupler in reducing CM currents on the power line network by as much as 20 dB up to the frequency of 15 MHz without affecting the integrity of the injected DM signals. Beyond this frequency, however, its isolation capability seemed to have deteriorated. Although the existence of alternative CM propagation paths within the modem is suspected to be the main cause, no further comprehensive study has been conducted to study and to understand the effects of the bridging capacitance of the optical coupler and other parasitic capacitances on these paths. Thus, this chapter aims to carry out an in-depth study on the effectiveness of implementing an optically-coupled AFE module in PLC modem to reduce the CM current in the power line network. To allow such study, an optically-coupled evaluation board based on the working principles of the AFE module in actual PLC modems is designed and implemented. Since implementation of an optical coupler within the AFE module is expected to increase the CM source impedance of the evaluation board by means of a low bridging capacitance, the equivalent CM source impedances of the evaluation board are measured and characterized using the two-current-probe measurement approach. CM equivalent circuit models of the evaluation board can then be established to study the CM noise suppression performance of the

optical isolator as well as to allow further analysis of the effects of other parasitic capacitances on its CM noise suppression capability. The measurement results collected have shown encouraging signs of reduction of CM currents in the power lines by including an optical coupler within the AFE module.

4.1 OPTICALLY-COUPLED ANALOG FRONT END MODULE

4.1.1 PLC Modem Architectural Overview

The PLC modem adopts a mixed solution of analog and digital circuits for the processing and transmission of data and voice signals along the power line networks over the RF band of 3 to 30 MHz. While the digital circuits are used for control, digital processing and signal modulation purposes, the analog circuits are more focused on the coupling, filtering and signal amplification of the emitted and received modulated signals. In this section, the analog and digital circuits in the PLC modem are described. The PLC modem's basic system architecture is shown in Fig. 4.3.

Fig. 4.3. Block diagram of a transformer isolated PLC modem design

The PLC modem's system architecture can be divided into three main blocks, which consist of:

1. The Analog Front End (AFE) module that is made up of:
 - a. Transceiver pre-amplifiers, power amplifiers and line drivers.
 - b. Power Line Coupling and Transient Suppression Unit
2. The Analog-to-Digital (ADC) and Digital-to-Analog (DAC) Converters
3. The Coder-decoder (Codec) circuitry, which can be further subdivided into two main blocks:

- a. The Upper Layer Communication Processor for digital signal processing and control
- b. The Physical Layer Communication Processor for modulation and demodulation operations

4.1.1.1 Coder-Decoder Circuitry

The Host System Micro-controller within the central processing unit (CPU) of any personal computer (PC) interprets commands for user applications and provides the upper layers (Application, Network, Data Link, etc.) of the system communication protocol. The Upper Layer Communication Processor provides micro-controller peripherals (e.g. communication interface) functions - mostly through LAN cables - to communicate with the Host System Micro-controller. The Physical Layer Communication Processor provides physical layer services (modulation, demodulation, etc.) of the communication protocol. The Physical Layer for all HomePlug-certified PLC modems uses the Orthogonal Frequency Division Multiplexing (OFDM) technique for transmission of data along the power line network in the frequency band of 4.49 MHz to 20.7 MHz.

4.1.1.2 Analog Front End Module

The Analog Front End module provides the interface functions between the controller and the low-voltage distribution network, which include galvanic isolation, signal filtering and signal amplification for broadband communication purposes. It also filters modulated signals containing high out-of-band harmonics before injecting them onto the power line network. At the receiver circuit, the in-built ADC would convert the analog signals received by the AFE module into the digital domain before being demodulated by the Physical Layer processor. The automatic gain control (AGC) system implemented at the input of the ADC would then scale the range of the received signal to match the dynamic amplitude range of the ADC. This AGC system is often implemented in a closed loop configuration, which consists of a proportional-integral (PI) controller that guarantees a stable operation, and a gain control amplifier that exhibits exponential characteristics.

To transmit the modulated analog signals over the noisy and extremely low-impedance power lines, some manufacturers incorporate a high performance line driver circuit or a power amplifier within the AFE module of their modem that can deliver high peak-to-peak amplitude analog signals with high output currents [54], [55] to boost the modulated signal before it is coupled onto the power line.

4.1.1.3 Power Line Coupling Circuit

In PLC systems, where the transmission medium is the high-voltage power line networks, it would be necessary to have a coupling unit, which provides galvanic insulation between the processing circuits and the power line in order to block the high-voltage AC mains frequencies (50/60 Hz) from entering the transceiver circuit, and to isolate users from high voltage hazards. Injecting a communication signal operating within the RF range is generally realized by capacitively coupling the transceiver's output to the input of the power line network by means of an isolation transformer having a bandwidth of 1 to 30 MHz and a bandpass filter. For safety reasons, the isolation transformer is always present to protect users from the dangers of high voltage shock hazards. The power line coupling circuit is thus an apparatus that act as an interface between PLC modem and AC mains. The power line interface unit has thus four main tasks:

1. Providing adaptation between the transmitted signals and the power line channel to minimize signal losses into the distribution network.
2. Providing galvanic isolation between the power line side and the user side to protect the users from the potentials of high voltage hazards.
3. Filtering of modulated signals with high out-of-band harmonics by means of the isolation transformer's bandwidth and the bandpass filter's cut off frequencies.
4. Protection of processing circuits by limiting the maximum signal excursion.

4.1.2 Development of an Optically-Coupled Evaluation Board

If properly implemented, the optical coupler within the AFE module would be able to provide galvanic isolation to protect both the user and equipment from the AC line

voltage without the complication of bulky isolation transformers, thus minimizing circuit complexity and reducing cost. Fig. 4.4 illustrates the typical implementation of the optically-coupled AFE module in a PLC modem as recommended by one of the leading PLC modem manufacturers [56].

Fig. 4.4. Block diagram of an optically-coupled PLC modem

Fig. 4.5. Block diagram of the Optically-Coupled Evaluation Board

An optically-coupled power line evaluation board based on the conceptual idea of the AFE module shown in Fig. 4.4 is fabricated. Fig. 4.5 and Fig. 4.6 show the block diagram and the practical implementation of the evaluation board, respectively. The evaluation board performs power line interface functions similar to the AFE module in the PLC modem, which include electrical isolation and signal conditioning for communication purposes. The isolation transformer is retained in the evaluation board for performance comparison purposes.

4.1.2.1 Fiber Optic Link as a Broadband Optical Coupler

One of the disadvantages of using the existing off-the-shelves optical couplers is its low operational bandwidth. For transmission of high-frequency signals especially in

the RF range, these optical couplers would impose severe restrictions on the transmission capability of broadband PLC modems. Due to the bandwidth limitation of the existing optical couplers, a fiber optic link is used in the experiment to simulate an optical coupler to provide the galvanic isolation for the evaluation board and to allow the transmission of broadband DM signals. The fiber optic link consists of a light emitting diode (LED) driver, a fiber optic transmitter and receiver, and a short length of a fiber optic cable as shown in Fig. 4.7.

Fig. 4.6. The optically-coupled evaluation board

4.7. A circuit schematic of the fiber optic link

The Agilent HFBR-1312T Transmitter and the HFBR-2316T Receiver are chosen to provide the fiber optic link. Both the transmitter and receiver are able to process high frequency signals up to 125 MHz and are compatible with the 50/125 μm and 62.5/125 μm optical fibers, which are commonly used in many data communication applications.

4.1.2.2 DC Power Supplies

To ensure proper isolation between the user and the power line and to preserve the inherent galvanic isolation of the AFE module, separate DC power supplies (VCC1 and VCC2) must be provided to the optical coupler to power up the transmitter and receiver circuits, respectively, as shown in Fig. 4.7. Some PLC modem manufacturers [56] galvanically isolate their VCC1 supply from the VCC2 supply by simply referencing the VCC2 voltage rail directly from the live and neutral terminals, which can be tapped from the primary windings of the power transformer as shown in Fig. 4.4. However, this method could prove to be dangerous to be implemented on the evaluation board since the live and neutral leads would be connected directly to the evaluation board's supply line and signal ground, respectively. The tester could be exposed to a hazardous voltage especially if the live and neutral connections are accidentally reversed. Therefore, for safety reasons, VCC1 is supplied by a 9V DC battery through a DC voltage regulator while VCC2 is supplied by the AC mains through a power transformer, as shown in Fig. 4.5. Both configurations would enable the optical coupler to provide effective galvanic isolation between the user side and the power line side of the modem.

4.1.2.3 Square Waveforms as Broadband Signal Source

The coding and decoding circuitry for the generation of the OFDM-modulated signal is not implemented in the evaluation board due to its high level of complexity and small relevance for the testing of the optically-coupled evaluation board. Therefore, another simpler method of transmission method is adopted: a voltage-controlled multivibrator chip, as shown in Fig. 4.8, that generates periodic pulses of trapezoidal waveform with the fundamental frequency of 1.063 MHz and the peak-to-peak amplitude of 4.46V_{pp} to produce harmonics that sweep across the desired frequency range of 3 to 30 MHz. Multivibrators [57] are basically regenerative electronic circuits that are used to generate square waves for implementation of a variety of two-state systems for oscillation and timing applications. The most common form is the astable or free-running type, which toggles spontaneously between one state and the other, without any external command. Using the Motorola MC1568 multivibrator chip, the fundamental frequency of 1.06 MHz of the generated waveform can be

controlled by means of an external capacitor connected between the I/O pins. The typical rise and fall time for the MC1658 are controlled to be at approximately 1.6ns each. Hence, the transmitted signal is able to produce harmonics that attenuate by 20dB per decade up to the maximum frequency of 19.89 MHz, as shown in Fig. 4.9.

Fig. 4.8. Motorola MC1568 as a voltage controlled multistable multivibrator

Fig. 4.9. Spectral envelope of the periodic pulse signal generated by the multivibrator

4.1.2.4 Design of Power Line Coupling Circuit

The coupling to the mains is accomplished by means of two essential elements: a transformer with a bandwidth of 1 MHz to 30 MHz for galvanic isolation, and an ‘X’ class coupling capacitor to block the 50/60 Hz AC sine wave and to avoid transformer core saturation. The inductance of the transformer is designed such that the inductive-capacitive (LC) circuit formed with the coupling capacitor acts as a high-pass filter for the incoming signal. The cutoff frequency f_c of the coupling unit is estimated as:

$$f_c = \frac{1}{2\pi\sqrt{LC}} \quad (4.1)$$

The insertion loss of the designed filter is less than -2 dB at the PLC operating frequencies (4.49 MHz to 20.7 MHz) and is approaching more than -11 dB at lower frequencies as shown in Fig. 4.11. The value of the capacitor is chosen to be large enough - between 47nF to 100nF - so that its DM source impedance at the PLC operating frequency is low, yet small enough to ensure high impedance value at the frequency of the AC mains. Since the impedance of the capacitor can be considered as part of the evaluation board's source impedance, ensuring the low impedance value of the capacitor minimizes the loss in the output voltage of the injected signal, which is caused by the varying loading effects of the power distribution network.

The DM source impedance of the evaluation board with the in-built coupling circuit in the frequency range of 100 kHz and 30 MHz is shown in Fig. 4.12, which is measured using the two-current-probe technique with the evaluation board connected to the network and set in the ‘active’ mode. It can be modeled as a series RLC resonant circuit with an equivalent series resistance of 10 Ω and an equivalent series inductance of 1 μH .

Fig. 4.12. Measured input impedance of the designed power-line interface

By providing a low impedance path for high frequency signals, the coupling unit is also exposed to high frequency transients that may have large amplitudes and fast rise times. Thus, a metal oxide varistor (MOV) is normally fitted at the input to the coupling unit as shown in Fig. 4.10 to provide protection against high transient voltages in the power line network by shunting the large currents generated away from the sensitive front-end analog circuits in PLC modems. In practice, however, the MOV is often subjected to a sustained abnormal over-voltage, which exceeds its maximum rating. The MOV can be overwhelmed by the surge and "explode", leaving the AFE module unprotected without any indication of this failure. Thus, the zener (or avalanche) diodes D_1 and D_2 are often implemented to act as a ‘last line of defence’ against the high voltage transients by clamping the voltage level to a lower value although they are not able to handle heavy surge currents as well as the MOVs.

4.1.2.5 Grounding Layout Considerations for Electromagnetic Compatibility

Careful PCB layout is important to ensure that the optical coupler in the PLC modem will be effective in isolating and suppressing the CM currents. In a PLC modem, the

PCB layout is often designed such that the analog and digital circuit signal grounds are connected to the metal chassis through a single or multiple points (although a single point connection is most commonly used). Most PLC modem systems require at least three separate ground returns, as shown in Fig. 4.13, where the signal ground used for low-level analog processing circuits is normally separated from the noisy digital ground used for digital processing circuits. A third ground is normally reserved for the mechanical enclosures or chassis connections.

Fig. 4.13. Different classes of grounding connections

Fig. 4.14. Parasitic leakage capacitances to the chassis ground and the earth ground

In practice, for safety reasons, the metal chassis is often connected to the earth ground at the chassis' power-input socket. If the digital and analog circuitry of the PLC modem is housed inside a metal chassis, especially with the metal chassis connected to the safety-earth ground, any leakage capacitances from the high-speed signal lines and grounds to the chassis would contribute to the existence of low impedance paths to the chassis ground and this would provide a complete close-loop low-impedance propagation path for the CM current to couple onto the power line and propagates. The CM currents, which flow between chassis ground and earth conductor in the

power line network due to this low-impedance path, can produce a potential difference between the earth and the chassis ground that causes radiated emissions to occur. Thus, it is very important to minimize the influence of these parasitic capacitances on the CM impedance of the PLC modem. Fig. 4.14 shows the possible parasitic leakage capacitances between the circuit traces to the chassis and earth ground in the optically-coupled evaluation board.

For the study and analysis of the CM noise propagation paths in the evaluation board and their effects on the CM noise suppression capability of the optical coupler, three jumpers: JP1, JP2, and JP3, shown in Fig. 4.14, are implemented in the evaluation board. Jumper JP1 is basically implemented to provide direct connection between the signal grounds of the circuits on the user side (GND1) and the metal chassis. Jumper JP2 is to provide direct connection between the signal grounds of the circuits on the power line side (GND2) and the metal chassis. Jumper JP3 would provide connection between the two signal grounds. These three jumpers would enable the testing personnel to create different low impedance CM paths to the power line cables via the ground chassis for the propagation of CM currents.

Fig. 4.15. CM equivalent circuit of an optically-coupled evaluation board

Fig. 4.15 represents the approximate CM equivalent circuit of the evaluation board based on the preliminary analysis of the evaluation board where V_{cm} represents the CM voltage source induced in the power line by the unbalance nature of the power distribution network. C_{opt} represents the bridging parasitic capacitance across the optical coupler. The capacitances C_{t1} and C_{t2} represent the parasitic capacitances across the power transformers and the isolation transformer, respectively while the capacitances C_1 , C_2 , C_3 , C_4 and C_5 represent the parasitic capacitances introduced by the analog and digital circuits in the evaluation board.

4.2 FEASIBILITY STUDY OF OPTICAL COUPLER IN PLC MODEM

4.2.1 Measurement Setup and Procedures

Since implementation of an optical coupler within the AFE module is expected to increase the CM source impedance of the evaluation board by means of a low bridging capacitance (C_{opt}), CM impedance measurements of the board are conducted using the two-current-probes setup shown in Fig. 4.16. The evaluation board is set to be able to transmit the signals continuously. Thus, the CM source impedance would be measured while the evaluation board is in the 'active' mode. To ensure the repeatability of the measurement, the coupling signal from the evaluation board is terminated with a line impedance stabilization network (LISN) as defined by the standards in [12] and [27].

Fig. 4.16. Measurement of the CM impedance of the evaluation board

To facilitate the measurement of the CM impedance of the evaluation board, a CM choke with an inductance of 3.2 mH, which provides sufficiently high impedance across the measured frequency range, is inserted between the LISN and the two-current-probe coupling circuit to provide sufficient CM isolation between the coupling circuit and the LISN so that only the CM impedance of the evaluation board is being measured by the two-current-probe setup. As the rated current of the evaluation board is measured to be 680 mA, a CM choke with current rating of 3A is chosen to avoid possible core saturation. The impedance of the LISN ($Z_{t,cm}$) is first obtained with the evaluation board disconnected from the AC mains. The impedance of the LISN with the evaluation board connected ($Z_{t,cm}$) is then measured in a

separated measurement. Hence, the actual CM source impedance ($Z_{s,cm}$) of the evaluation board is given by

$$Z_{s,cm} = \frac{Z_{t,cm} Z_{l,cm}}{Z_{l,cm} - Z_{t,cm}} \quad (4.1)$$

The CM choke is then removed for the measurement of the CM current generated by the evaluation board using another clamp-on type RF current probe.

4.2.2 Effects of the Isolation and Power Transformers on CM Currents

This study is conducted with jumper JP1 connected to the chassis ground as shown in Fig. 4.17. The measurements are also conducted in two different phases: with jumper JP3 connected and disconnected to disable and activate the optical coupler, respectively. The initial measurement is conducted with jumper JP3 connected, as shown in Fig. 4.17, to determine only the effects of the bridging parasitic capacitances across the windings of the power transformer and the isolation transformer (C_{l2} and C_{l1} , respectively) on both the effective the CM source impedance of the evaluation board and the CM current generated at the signal interface.

Fig. 4.17. CM equivalent circuit of the evaluation board with JP1 and JP3 connected

Jumper JP3 is connected between the signal grounds of the circuits on the power line side and the user side (GND 2 and GND1, respectively). By connecting these two signal grounds, CM currents are allowed to bypass the parasitic capacitance across the optical coupler and hence, the galvanic isolation capability of the evaluation board is effectively disabled. Jumper JP1 is connected between the signal ground of the circuits on the user side (GND1) and the metal chassis to simulate the common practice of grounding the signal grounds of the analog and digital processing circuits at a single point to the metal chassis in PLC modems. Based on these configurations,

it can be seen from Fig. 4.18 that the CM impedance measured is capacitive in nature with an effective capacitance of 48 pF. Since the jumpers JP1 and JP3 nullify the effects of the parasitic capacitances contributed by the circuits on the user side (C_3 , C_4 and C_5) and the power line side (C_1 and C_2), respectively, on the effective CM impedance of the evaluation board, the capacitive nature of the CM impedance measured is contributed mainly by the bridging parasitic capacitances provided by the isolation transformer and the power transformer.

Fig. 4.18. CM input impedance before and after the optical coupler is enabled

Fig. 4.19. CM equivalent circuit of the evaluation board with JP3 disconnected

With JP1 remains connected, JP3 is then disconnected in the second phase of measurement to activate the galvanic isolation of the evaluation board, which is provided by the bridging parasitic capacitance across the optical coupler (C_{opt}), by isolating GND1 from GND2 as shown in Fig. 4.19. It can be also seen from the impedance measurement results in Fig 4.18 that the effective CM source impedance of the evaluation board has increased with its effective capacitance reduced to a lower capacitive value of 17 pF.

Fig. 4.20. CM current before and after the optical coupler is enabled

Fig. 4.20 shows the measured frequency domain spectrum of the CM current using the clamp-on current probe. The measurement results clearly indicate the effectiveness of the optical coupler in reducing the level of CM current generated at the power outlet between a minimum value of 10 dB at the frequency of 13.77 MHz and a maximum value of 28.94 dB at the frequency of 24.37 MHz. The increase in the CM source impedance of the evaluation board, which reflects the increase in the impedance of the CM propagation path within the board, has brought the CM current below the CISPR 22 limit [26] between the minimum margin of 0.71 dB (at 13.77 MHz) and the maximum margin of 29.26 dB (at 24.37 MHz). Comparisons with the initial measurement results also show that the optical coupler has reduced the CM current level by nearly 10 to 30 dB up to 30 MHz, except for an increase of 8.79 dB at the frequency of 15 MHz. This shows that the CM source impedance of 17 pF provided by the evaluation board is not large enough to reduce the influence of the unbalance of the power line network on its CM noise attenuation power. The effects of the unbalance characteristics of the power line network on the CM current generated by PLC modems will be analyzed in greater detail in Chapter 5.

4.2.3 Effects of Floating Ground on CM Current

The next analysis is conducted with jumper JP1 disconnected from the chassis ground. The measurements are also conducted in two different phases: with the optical coupler disabled and enabled. The first phase of the measurement is conducted with jumper

JP3 connected to deactivate the galvanic isolation of the evaluation board as shown in Fig. 4.21. This is to determine the effects of introducing additional parasitic capacitances contributed by the circuits on the user side (C_3 , C_4 and C_5) of the evaluation board on the effective CM source impedance of the board, and the resultant CM current generated.

Fig. 4.21. CM equivalent circuit of the evaluation board with JP1 disconnected

Fig. 4.22. CM input impedance before and after the optical coupler is enabled

From the impedance measurement results in Fig. 4.22, it can be seen that even with the optical coupler disabled, the additional effects of parasitic capacitances C_3 , C_4 and C_5 contributed by the circuits in the user side have created an additional barrier within the CM propagation path, and this results in an increase in the CM source impedance measured, which is capacitive in nature with a capacitance value of 28 pF, as compared to the capacitive value of 48 pF when JP1 was connected to the chassis.

In the second phase of the measurement, with JP1 remained disconnected, JP3 is then disconnected, as shown in Fig. 4.23, to allow the optical barrier to come into effect.

Interestingly, as seen from the impedance measurement results in Fig. 4.22, the CM impedance is measured to be having a capacitance of 17 pF, which is similar to the capacitance value obtained when JP1 is connected to the chassis ground. If comparisons are to be made between these two configurations (compare Fig. 4.23 with Fig. 4.19), it can be seen that the parasitic capacitance across the optical coupler (C_{opt}) is significantly small enough to effectively isolate the effects of the parasitic capacitances C_3 , C_4 and C_5 from the power line side once the optical coupler is activated. Thus, only the parasitic capacitances C_1 and C_2 , which are contributed by the circuits on the power line side, are reflected in the CM source impedance value. Therefore, the galvanic isolation provided by the optical coupler is significantly large enough to reduce the effects of the parasitic capacitances C_3 , C_4 and C_5 on the propagation paths of the CM currents and thus, reducing the level of the CM current generated. However, the parasitic capacitances C_1 and C_2 remain un-isolated from the power mains. Hence, the values of capacitances C_1 and C_2 would continue to influence the CM source impedance of the evaluation board.

Fig. 4.23. CM equivalent circuit of the evaluation board with JP3 disconnected

Fig. 4.24 shows the measured frequency domain spectrum of the CM current using the clamp-on current probe. The measurement results demonstrate the significant effects of the parasitic capacitances C_3 , C_4 and C_5 on the CM current level when the optical coupler of the evaluation board is disabled. Hence, it can be seen that the additional effects of parasitic capacitances contributed by the circuitry in the user side of the evaluation board have created an additional barrier within the CM propagation path, which results in the reduction of the CM current. Comparison with the CISPR 22 limits indicates that this additional barrier has reduced the CM current down to a minimum margin of -2.59 dB (at 9.52 MHz) and a maximum margin of 29.26 (at 26.49 MHz) below the limits throughout the PLC frequency range. This is a

significant improvement compared to the minimum margin of -11.94 dB (at 9.52 MHz) and a maximum margin of 0.54 (at 26.49 MHz) below the limits, which is achieved by the evaluation board when the signal grounds of the circuits in the user side is connected to the chassis ground through jumper JP1.

Fig. 4.24. CM current before and after the optical coupler is enabled

Fig. 4.24 also shows the measured frequency domain spectrum of the CM current generated when the optical coupler of the evaluation board is activated (by connecting the grounds of the user side circuit and power line side circuit through jumper JP3). The measurement results show the effectiveness of the optical coupler in reducing the level of CM current generated at the power outlet between a minimum value of 1.5 dB (at the frequency of 20.11 MHz) and a maximum value of 16.98 dB (at the frequency of 3.17 MHz). It can also be seen from Fig. 4.24 that the level of CM current measured at the output of the evaluation board with jumper JP1 disconnected from the chassis ground and the signal ground of the user side is almost similar to the level of CM current generated by when jumper JP1 is connected. In both configurations, the CM impedance of the evaluation board is measured to be having a capacitance of 17 pF.

4.2.4 Effects of Metal Chassis on CM Currents

Metal housings are often implemented for shielding electronic units to prevent both the emission and absorption of electromagnetic waves, which could lead to malfunctioning of the unit and other electronic equipment. However, some PLC

modems available in the market have a plastic body rather than a metal housing. This could be because plastic injection molded parts are believed to be more economical to produce and also they offer maximum design diversity. Thus, the third set of measurements is conducted to determine the effects of removing the metal plane on the effective CM source impedance of the evaluation board, and the resultant CM current generated. The measurements are also conducted in two different phases: with the optical coupler disabled and enabled.

Fig. 4.25. CM impedances of the Evaluation Board and PLC modems of other brands

The first phase of the measurement is conducted with jumper JP3 connected to deactivate the optical coupler to measure the CM source impedance of the evaluation board. Based on this configuration, it can be seen from Fig. 4.25 that the CM impedance measured is capacitive in nature with an effective capacitance of 27 pF. Fig. 4.25 also compares the measured CM impedance of the non-isolated evaluation board with the measured CM impedances of two HomePlug-compliant PLC modems from two different manufacturers that are available in the market. The PLC modems exhibit CM impedances with capacitive values of 23 pF and 20 pF that are comparatively close to the CM impedance value of the evaluation board since the electronic units of the two modems are not embedded in a metal housing but are embedded in plastic housings instead. From the impedance measurement results in Fig. 4.25, it can be seen that even with the removal of the metal plane, the measured CM source impedance, which is capacitive in nature with a capacitance of 27 pF, is significantly close to the measured CM impedance of the evaluation board (with the

galvanic isolation capability disabled) when the metal plane was present and the signal ground of the circuits in the user side was isolated from the metal plane. Even with the earth conductor physically disconnected from any part of the evaluation board, capacitive coupling still occurs between the signal and ground traces of the circuits and the earth conductor as highlighted in Fig. 4.26. Thus, the existence of these parasitic capacitances between the two lines contributes to the formation of a low-impedance propagation path for CM, and this resulted in the significant value of the CM impedance measured.

Fig. 4.26. Parasitic capacitances between signal ground terminals and earth conductor

Fig. 4.27. CM equivalent circuit of the evaluation board with JP3 connected

In the second phase of the measurement, JP3 is then disconnected to activate the galvanic isolation of the evaluation board by isolating GND1 from GND2. It can be seen from the source impedance measurement results in Fig 4.27 that the CM source impedance of the evaluation board has increased with its effective capacitance

reduced further to a lower value of 11 pF. Thus, the measurement results have demonstrated the effectiveness of the optical coupler in reducing the coupling capacitances between the signal and ground traces of the processing circuits on the evaluation board and the earth ground. However, the capacitive coupling between signal and ground traces of the circuits in the power line side of the evaluation board and the earth conductor still occurs although with lower capacitive value.

Fig. 4.28. CM current before and after the optical coupler is enabled

Fig. 4.28 shows the measured frequency domain spectrum of the CM current using the clamp-on current probe. Fig. 4.28 also shows the measured frequency domain spectrum of the CM current generated when the optical coupler of the evaluation board is enabled. The measurement results demonstrate that the galvanic isolation provided by the optical coupler is significantly large enough to reduce effects of the parasitic capacitances between signal and ground traces of the circuits in the evaluation board and the earth conductor on the propagation paths of the CM currents. Thus, the level of the CM current generated is reduced. Comparison between the two sets of measurement results highlight the effectiveness of the optical coupler in reducing the level of CM current generated at the power outlet between a minimum value of 1.59 dB (at 9.52 MHz) and the maximum value of 23.16 dB (at 3.17 MHz). Comparison with the CISPR 22 limits further indicates that this additional barrier has reduced the CM current down to a minimum margin of 3.68 dB (at 13.77 MHz) and a maximum margin of 27.19 (at 24.52 MHz) below the limits throughout the PLC frequency range. This is a significant improvement compared to the minimum margin

of -4.21 dB (at 9.52 MHz) and a maximum margin of 18.65 dB (at 26.49 MHz) below the limits, which is achieved by the evaluation board without any galvanic isolation.

4.3 CONCLUDING REMARKS

Based on the study of the optically-coupled evaluation board, the author has shown that the inherent low-parasitic bridging capacitance of the optical coupler provides much higher CM source impedance to the evaluation board over the conventional transformer coupler where the winding-to-winding capacitance is generally higher. Experiments conducted have yielded promising results, which indicate the capability of the optical coupler to provide an alternative way of transmitting high frequency signals and providing galvanic isolation against CM currents. Such inherent feature is desirable because it results in superior CM current attenuation and therefore lowers the level of the CM current generated in the power line network. If the bandwidth limitations of the optical coupler are successfully overcome, the optical coupler could be manufactured into an integrated circuit to provide optical isolation that gets rid of the bulky isolation transformer and it could greatly reduce the total number of discrete components in the AFE module in the PLC modem.

There shall always be some amount of CM currents being generated into the power line network, which could be attributed to the existence of alternative CM propagation paths within the AFE module. The study presented has also identified the existence of alternative CM propagation paths through the parasitic capacitances introduced by the digital and analog processing circuit that would affect the performance of optical coupler. For PCB designers, special attention has to be given to the design of the signal and grounding layouts for the digital and analog-processing circuits so as to reduce their parasitic capacitances as much as possible as these capacitances could contribute to the interference in the power line cable channels. Therefore, controlling these effects is important when designing the signal and grounding layouts in the PCB circuit. For the optical coupler to be able to provide effective galvanic isolation against CM currents, separate groundings should also be provided to the optical coupler. Thus, JP3 have to be disconnected to isolate the signal ground of the power line side (GND2) from the signal ground of the user side (GND1). For more effective control over the CM currents, JP2 should also remain open to isolate GND2 from the

chassis ground. This is to prevent the CM currents from bypassing the isolation barrier provided by the optical coupler and creating an alternative CM propagation path between GND2 and the chassis ground. Once the two conditions are met, any signal ground-to-chassis connections (through JP1) at the user side of the modem would not affect the CM suppression performance of the optical coupler. This would give the PCB designer more flexibility in designing suitable grounding layouts according to the circuit configurations in the user side of the modem. For example, analog circuits with signal frequencies up to 300 kHz may be candidates for single-point grounding while digital circuits with signal frequencies in the high frequency range should use multipoint grounding. This is important to further reduce the overall electromagnetic emissions from the PLC modem since the bulk of the analog and digital circuits are located in the user side.

However, even with well-designed PCB circuits, the electrical unbalances in the power transmission lines would still give rise to unbalanced currents. Hence, characterization of the unbalance properties of the in-house power line networks is needed to understand their contribution to the generation of the CM current in the power line network. Since the unbalance characteristic of the power line network varies from places to places, a full equivalent circuit model is needed to reflect such variation. The full equivalent circuit model, which emulates the connection between PLC modem and the power line network, would allow one to quantify the CM suppression performance of a built-in optical coupler. Also, more measurements could be conducted at different residential buildings to obtain a good statistical power line network model in Singapore. These additional efforts would allow more conclusive findings on using optical coupling technique in PLC modem to suppress CM currents in the power line network.

Chapter 5

MODELING OF UNBALANCE PROPERTIES OF POWER LINE NETWORK

In order to limit the radiation from the low voltage distribution network (LVDN) due to the usage of Power Line Communication (PLC) system, design guidelines for the PLC modem on the level of the differential mode (DM) output voltage to be injected into the LVDN, and the level of common mode (CM) source impedance introduced by the optical coupler, have to be established. Although the PLC modems can be designed to be reasonably balanced with well-designed optical coupler and proper PCB layout designs, the effects of the electrical unbalances that are present in the LVDN due to the geometry of the wires, the effects of the surrounding environment and the electrical equipment connected to them, would still be significant enough to give rise to the CM currents. If left unchecked, these CM currents would create magnetic fields, which could interfere with the surrounding electronic equipment and other wireless communication systems.

Therefore, the mechanism behind the modal conversion of the injected DM signals needs to be studied. To allow such study, a full equivalent circuit model of any PLC modem connected to any power outlet has to be established to predict the level of the CM current generated at the signal interface due to the unbalance nature of the LVDN. Based on a two-current-probe impedance measurement approach, the parameters for the equivalent circuit model of the PLC modem and the indoor LVDN are characterized. Measurements are conducted at different residential buildings to obtain good statistical parameters for the power line network model in Singapore. Since these parameters are obtained through measurement, the model derived would take into account the characteristics of the LVDN in terms of the DM and CM impedances, and the unbalance levels of the power line due to the effects of different layout geometries and electrical loading conditions. The DM and CM impedances of the PLC modems and the output levels of the DM signals are also included. This equivalent circuit model would be useful for evaluating the CM current suppression performance of a PLC modem with a built-in optical coupler at the earliest design stage without having to conduct expensive field measurements. The derived data

would also be useful for future EMI regulatory and PLC modem design considerations.

5.1 MODELING OF POWER LINE COMMUNICATION SYSTEM

5.1.1 CM Current Generation Mechanism

In a PLC network, the causes of the CM current are the DM signal source of the modem and the unbalance of the LVDN. The mechanism of the generation of the CM current can be explained further with Fig. 5.1, which shows the equivalent circuit of the PLC modem connected to the LVDN through a power outlet. In this figure, the PLC modem, as seen by the LVDN, can be represented as a balanced DM RF signal source E_s with source impedances (Z_1 and Z_2) and CM impedance (Z_3).

Fig. 5.1. Equivalent circuit of a PLC modem and the power line network

The power line network equivalent circuit, as seen by the modem, can be modeled by a T-network consisting of discrete impedances Z_4 , Z_5 and Z_6 . With this T-network, any unbalances in the power line network can be easily represented and taken into account. From the equivalent circuit, the following equations can be obtained:

$$E_s = I_1(Z_1 + Z_4 + Z_3 + Z_6) - I_2(Z_3 + Z_6) \quad (5.1)$$

$$E_s = -I_1(Z_3 + Z_6) + I_2(Z_2 + Z_5 + Z_3 + Z_6) \quad (5.2)$$

By combining (5.1) and (5.2), we have

$$(I_1 - I_2) = \frac{(Z_2 - Z_1) + (Z_5 - Z_4)}{(Z_3 + Z_6)(Z_1 + Z_2 + Z_4 + Z_5) + (Z_4 + Z_1)(Z_5 + Z_2)} E_s \quad (5.3)$$

where $I_{cm} = I_1 - I_2$ is the CM current caused by the unbalance of the power line network. Most telecommunication devices, including PLC modems, are normally designed to be balanced symmetrical sources with $Z_1 = Z_2$. Therefore, under such a condition, (5.3) can be simplified to

$$I_{cm} = \frac{4\Delta Z}{4(Z_{lt})(Z_t + Z_0) + Z_0^2 - \Delta Z^2 + 2Z_t Z_0 + Z_t^2} E_s \quad (5.4)$$

where $Z_t = Z_1 + Z_2$ and $Z_0 = Z_4 + Z_5$ are the DM impedances of the PLC modem and the LVDN, respectively. $Z_{lt} = Z_3 + Z_6$ is the total CM impedance of the PLC modem and the LVDN. $\Delta Z = |Z_5 - Z_4|$ indicates the unbalance level of the LVDN. It can thus be seen that the factors that affect the magnitude of the CM current generated at the output of the PLC modem are the electrical unbalance of the LVDN ($Z_4 \neq Z_5$), which is responsible for the conversion of the transmitted DM signal E_s into CM voltage, and the CM impedances of the PLC modem and the LVDN (Z_3 and Z_6 , respectively), which are responsible for the conversion of the CM voltage into CM current. If the DM source impedance (Z_t) of the PLC modem is somehow designed to be insignificantly small compared to the DM impedance Z_0 of the LVDN, the CM current can be easily simplified to

$$I_{cm} = \frac{4\Delta Z}{4(Z_{lt})(Z_0) + Z_0^2 - \Delta Z^2} E_s \quad (5.5)$$

Since the effects of the unbalance level of the power line network is beyond one's control, the CM current can only be reduced either by increasing the CM source impedance Z_3 of the PLC modem by means of a good isolation unit (e.g. an optical coupler as shown in Chapter 4) or by reducing the level of the injected DM signal E_s .

5.1.2 Development of the Full Equivalent Circuit Model

To be able to quantify the performance of the optical coupler in an actual power line environment, an equivalent circuit that represents the power line network is needed so

that the CM current generated by a PLC modem with a built-in optical coupler can be estimated with reasonable accuracy. Using the T-network equivalent circuit for the PLC system as shown in Fig. 5.1, the CM current generated at the signal interface between the PLC modem and the LVDN through the power outlet can be estimated based on the knowledge of the values of the T-network equivalent parameters of the PLC modem (Z_1 , Z_2 and Z_3) and the LVDN (Z_4 , Z_5 and Z_6). The parameters for the LVDN can be obtained from the input impedances: Z_{m1} (between live and neutral), Z_{m2} (between live and earth) and Z_{m3} (between neutral and earth), which can be measured using the two-current-probe measurement approach at any power outlet, as shown in Fig. 5.2, where the PLC modem is expected to be connected. With the impedances Z_{m1} , Z_{m2} and Z_{m3} measured the T-network parameters Z_4 , Z_5 and Z_6 can be derived as follows:

$$Z_4 = \frac{Z_{m1} + Z_{m2} - Z_{m3}}{2} \quad (5.6)$$

$$Z_5 = \frac{Z_{m1} - Z_{m2} + Z_{m3}}{2} \quad (5.7)$$

$$Z_6 = \frac{-Z_{m1} + Z_{m2} + Z_{m3}}{2} \quad (5.8)$$

Fig. 5.2. Two-current-probe setup for measurement of input impedances of AC mains

The relations between the measured input impedances (Z_{m1} , Z_{m2} and Z_{m3}) and the unbalance level and the DM and CM impedance parameters (ΔZ , Z_0 and Z_{cm} , respectively) of the power line network then can be given by

$$\Delta Z = |Z_4 - Z_5| = |Z_{m2} - Z_{m3}| \quad (5.9)$$

$$Z_0 = Z_4 + Z_5 = Z_{m1} \quad (5.10)$$

$$Z_{cm} = Z_4 // Z_5 + Z_6 \quad (5.11)$$

For validation purposes, the input impedances of the LISN, which is used for the study and analysis of the optically-coupled evaluation board in Chapter 4, are measured and converted to the equivalent T-network impedance parameters. The measurements of the network impedance parameters are also carried out at the power outlet of an indoor LVDN in a research laboratory located within the premises of the university. Fig. 5.3 and 5.4 show the values of the impedance parameters measured at these two locations.

Fig. 5.3. T-network impedance parameters for the LISN

From Fig. 5.3 and 5.4, the values of Z_4 and Z_5 , which represents the components of the DM impedance ($Z_0 = Z_4 + Z_5$) of the network under test, and the values of Z_6 , which represents the CM impedance of the network under test, vary greatly with frequency in the range of a few ohms and a few hundred ohms. At lower frequencies, the impedance curves at these two measurement points exhibit inductive behaviors. At higher frequencies above 15 MHz, the impedance curves become more resistive and exhibit asymptotical behaviors. The positions of the maximum peaks vary with location and network topology.

Fig. 5.4. T-network impedance parameters for in-house LVND in Residential Unit

It can also be seen from the two figures that the unbalance levels ($\Delta Z = |Z_4 + Z_5|$) also increase with frequency with the highest value almost comparable to the DM and CM impedance values. Thus, these observations contradict the assumptions made by some researchers [19], [22] that the unbalance levels of the LVND can be assumed to be significantly small compared to the DM impedance values of the LVND. The T-equivalent circuit model for the LVND is simply derived by fitting appropriate equivalent resonant circuits [58], [59] to the measured T-network impedance parameters (Z_4 , Z_5 , and Z_6). Parallel RLC resonant circuits, shown in Fig. 5.5, can be constructed in series configuration to match the curves of the measured impedance parameters where each of these circuits would be assigned to each maximum peak of the impedance parameters. At the resonant frequency f_{0_max} of each peak, the impedance of each circuit is equal to the equivalent resistance R_p , which can be determined from the measured curves in Fig. 5.3 and Fig. 5.4. With the values of R_p and f_{0_max} known, the suitable equivalent inductance L_p and capacitance C_p of the parallel RLC resonant circuit assigned for each peak can be determined by

$$C_p = \frac{1}{2\pi f_{0_max} R_p (BW)} \quad (5.12)$$

$$L_p = \frac{1}{(2\pi f_{0_max})^2 \cdot C_p} \quad (5.13)$$

where BW is the frequency spread between two frequencies at which the overall effective impedance of the parallel RLC resonant circuit decreases to 0.707 times its maximum value of R_p .

Fig. 5.5. Parallel RLC resonators and their respective frequency responses

Alternative, series RLC resonant circuits can be constructed in parallel configuration as shown in Fig. 5.6 to match the curves of the measured impedance parameters where each of these circuits is assigned to each minimum peak of the parameter curves. At the resonant frequency f_{0_min} of each peak, the impedance of each circuit is equal to the equivalent resistance R_s , which can also be determined from the measured impedance parameters in Fig. 5.3 and 5.4. With the values of R_s and f_{0_min} known, the suitable equivalent inductance L_s and capacitance C_s for the series RLC resonant circuit assigned for each peak can be determined by

$$C_s = \frac{BW}{2\pi f_{0_min} R_s} \quad (5.14)$$

$$L_s = \frac{1}{(2\pi f_{0_min})^2 \cdot C_s} \quad (5.15)$$

where BW is the frequency spread between two frequencies at which the overall effective impedance of the series RLC resonant circuit increases to 0.707 times its

minimum value of R_s . The impedance characteristics of these models show quite a close fit to the measured characteristics.

Fig. 5.6. Series RLC resonators and their respective frequency responses

5.1.3 Validation of the Full Equivalent Circuit Model

The same evaluation board with optical coupler developed in Chapter 4 is employed to emulate a PLC modem. With the known transmitted signal (E_s), DM source impedance (Z_t) and CM source impedance (Z_3) of the evaluation board, the CM current generated can be calculated. Based on the previous characterization of the evaluation board in Chapter 4, the CM source impedance of the evaluation board with the optical coupler activated can be represented by a capacitance of 17 pF. The DM impedance of the PLC modem can be represented with a 10 Ω resistor connected in series with a 1 μ H inductor. For the transmitted signal (E_s), periodic pulses of trapezoidal waveform with a fundamental frequency of 1.063 MHz and peak-to-peak amplitude of 4.46 V_{pp} can be used to approximate the transmission signal injected into the network under test. For a periodic waveform, the Fourier spectrum is composed of series of discrete frequency components, which comprises the fundamental frequency f_0 and all integer n multiples of f_0 . The amplitude of the spectrum of such a periodic pulse train is given as

$$|C_n| = 2V_{pp} \frac{\tau}{T} \left| \frac{\sin(n\pi\tau/T)}{n\pi\tau/T} \right| \left| \frac{\sin(n\pi t_r/T)}{n\pi t_r/T} \right| \quad (5.16)$$

where V_{pp} is the peak-to-peak transmission signal amplitude, while T , τ , and t_r are the repetition time, pulse width, rise time, respectively. Preliminary measurements conducted have determined T to be at $0.94 \mu\text{s}$, τ to be 470.02 ns , and t_r and t_f to be 3.11 ns .

Fig. 5.7. Equivalent circuit of the optically-coupled evaluation board and the LISN

The full equivalent circuit model of the optically-coupled evaluation board terminated by the LISN is shown in Fig. 5.7. Fig. 5.8 compares the measured CM current generated by the PLC evaluation board and the calculated CM current based on the equivalent electrical model given in Fig. 5.7. Close agreement between the measured and the calculated CM currents is demonstrated. Both the analytical and measurement results for the CM current in Fig. 5.8 show that the CM currents generated are generally low except in the frequency range of 13 to 18 MHz. From the impedance curves in Fig. 5.3, it can be seen that the unbalance level (ΔZ) of the LISN are highest in the frequency range of 13 to 18 MHz. These observations further strengthened the conclusion made during the design stage of optically-coupled evaluation board in Chapter 4 that the CM source impedance - with a capacitive value 17 pF - provided by the optical coupler is not large enough to isolate the effects of the unbalance of the LVDN on the CM current generated. Therefore, if the capacitive value of the CM source impedance of the evaluation board is reduced further, the level of CM currents

generated in this frequency range can also be expected to drop.

Fig. 5.8. Comparison of the measured and calculated CM currents

Fig. 5.9. Equivalent circuit of the optically-coupled evaluation board and the indoor LVND in a research laboratory

The full equivalent circuit model of the optically-coupled evaluation board connected to the power outlet in the research laboratory is shown in Fig. 5.9. Fig. 5.10 compares the CM current measured near the signal interface of the evaluation board and the indoor LVND in the research laboratory, and the CM current derived from the

equivalent circuit model. The results demonstrate close agreement between the measured and the estimated results. Hence, with the T-network equivalent circuit model, the performance of the optically-coupled evaluation board on the indoor power line network of typical Singapore household could be predicted and analyzed with reasonable accuracy without having to conduct expensive field measurements.

Fig. 5.10. Comparison of the measured and calculated CM currents

5.2 PERFORMANCE OF OPTICAL COUPLER IN ACTUAL AC MAINS

5.2.1 Characterization of the Indoor AC Mains in Residential Buildings

In this section, the levels of CM currents generated by the optically coupled evaluation board in the indoor LVDN of residential units in Singapore are estimated and evaluated. To obtain a reliable estimation of the CM current value for the indoor LVDN in Singapore, a statistical measurement and characterization must be performed on various indoor power outlets across the country to ensure that the full equivalent circuit developed is representative of the indoor LVDN conditions in Singapore. Thus, several measurement locations are chosen so that different types of installations and building usage are covered by the measurements. Fig. 5.11 and Fig. 5.12 illustrate the layouts of two four-room residential units that are selected for the measurement along with the location of the power outlets. Both units are developed by the Housing Development Board of Singapore (HDB), which are typical of the residential units in Singapore. In each residential unit, different power outlets at points A–E, F are selected and measured accordingly. All measurements are performed

during the day where the household electrical appliances are normally connected to the AC mains.

Fig. 5.11. Power outlets A–F in a four-room HDB residential unit in East Singapore

Fig. 5.12. Power outlets A–E in a four-room HDB residential unit in West Singapore

The power line network under study in the two residential units consist of stranded, copper core cables, which are commonly used for in-house wiring for residential power line networks in Singapore. The cross-sectional area of each conductor (live, neutral or earth) is 2.5 mm^2 and the diameter of a single copper strand in each conductor is 0.67 mm. The stranded copper-core cables are also covered with PVC insulation (not laid in metal conduit) as shown in Fig. 5.13. The structure of the

networks in both residential units is single-phased and each network is divided into rings or loops. The network is connected to the external 230Vac power supply through an AC distribution board and an electricity meter, which effectively filter off the high frequency signals and noises coming from the neighboring residential units and thus isolating the network from the effects of the varying external electrical loads from the neighboring units.

Fig. 5.13. Cross-sectional view of the in-house power line cable

Figures 5.14, 5.15 and 5.16 show the DM impedances (Z_0), the CM impedances (Z_{cm}) and the unbalance levels (ΔZ), respectively, of the indoor LVDN in the two residential units under different loading conditions. The unbalance levels and the modal impedances are obtained from the input impedances (Z_{m1} , Z_{m2} and Z_{m3}) measured at each power outlet using the two-current-probes approach and the equations defined in (5.6)–(5.11).

Fig. 5.14. Derived DM impedances (Z_0) of the indoor LVDN for the residential units

Results from Fig. 5.14 and Fig. 5.15 show that the both the CM and the DM impedances of the LVDN exhibit resonant behaviors as the frequency increases. This

is expected because as the length of the cable approaches the wavelength of the RF signal, the transmission line effects surface. The measurement results collected show a large variety of impedance values, which extends from a flat minimum of approximately $10\ \Omega$ to a large number of peaks up to approximately $2\ \text{k}\Omega$.

Fig. 5.15. Derived CM impedances (Z_{cm}) of the indoor LV DN for the residential units

Fig. 5.16. Derived unbalance levels (ΔZ) of the indoor LV DN for the residential units

Fig. 5.16 shows the unbalance levels of the indoor LV DN for the two residential units, which are obtained at different power outlets. The measurement results also show that the variations in the electrical unbalance levels between these power outlets are larger in the lower frequency range of 3 to 15 MHz. These large variations could be caused mainly by the effects of the different electrical appliances connected to the LV DN. At higher frequencies, as the length of the cable separating the electrical appliances and the power outlet approaches the wavelength of the RF signal, the transmission line

effects of the power cables would dominate, thus neglecting the unbalance effects of the electrical loads. Therefore, the variations in the electrical unbalance levels between these power outlets are smaller. However, the measured results also show that although the contribution of the power cables to the electrical unbalance of the LVDN is lower than the contribution made by electrical loads, small changes in the cable wiring and network configuration may still induce significant changes in the LCL values of the PLC network. Thus, the level of electrical unbalance is also dependent on the cabling layout and network configuration of the power line network to a certain extent.

5.2.2 Estimation of CM Current in the Actual AC Mains

The CM current values at the indoor power plugs in these units are calculated using the full equivalent circuit models constructed based on the T-network impedance parameters derived in the earlier measurement campaigns. Fig. 5.17(a) and Fig. 5.17(b) show the estimated CM current generated at the signal interface of the optically-isolated evaluation board and the indoor LVDN of the two four-room residential units under various loading conditions. From these figures, it can be seen that the CM currents generated at the power outlets under different loading and network layout conditions are estimated to exceed the limit imposed by CISPR 22 [27] by an average margin of 13 dB throughout the RF range of 3 to 30 MHz.

Fig. 5.17(a). Estimated levels of CM currents generated at the indoor power outlets in a four-room HDB residential unit in the eastern part of Singapore.

Fig. 5.17(b). Estimated levels of CM currents generated at the indoor power outlets in a four-room HDB residential unit in the western part of Singapore.

The level of the CM current can be reduced to below the limit imposed by CISPR by decreasing the value of the DM output voltage E_s injected into the LVDN. From the equivalent circuit models developed for the power outlets of the four room residential unit, it can be seen that the level of the CM currents generated at these power outlets can be estimated to reduce to a level below the limit with a minimum margin of 0 dB if DM output voltage injected by the optically-coupled evaluation board is reduced from the high peak-to-peak amplitude of 4.46 Vpp to lower values ranging from 2.95 Vpp to 86 mVpp depending on the unbalance properties of each power outlet. Fig. 5.18(a) and Fig. 5.18(b) highlight these estimations.

Fig. 5.18(a). Estimated levels of CM currents generated at the indoor power outlets in a four-room HDB residential unit in the eastern part of Singapore.

Fig. 5.18(b). Estimated levels of CM currents generated at the indoor power outlets in a four-room HDB residential unit in the western part of Singapore.

Alternatively, the level of CM currents generated at the power outlets can also be reduced to below the CISPR limit with the DM output voltage maintained at the peak-to-peak amplitude of 4.46 V_{pp} if the value of the CM source impedance of the evaluation board is reduced. Based on the full equivalent circuit model, the capacitive value of CM source impedance of the evaluation board need to be reduced to lower values ranging from 8 pF to at least 0.5 pF! For this to be made possible, proper PCB design has to be implemented on the optically-coupled evaluation board to eliminate the effects of parasitic capacitances, which are contributed by the analog and the digital processing circuits, on the CM source impedance value so as to achieve the optimum isolation power for the optical coupler.

5.3 CONCLUDING REMARKS

Based on the preliminary measurement results of the optically-coupled evaluation board in Chapter 4, the author has shown that the inherent low-parasitic capacitance of the optical coupler increases the CM source impedance of the PLC modem and therefore lowers the level CM current in the power network. However, the CM attenuation performance of the optical coupler is also dependent on the unbalance properties of the power line network. Since the unbalance characteristics of the in-house power line network varies from places to places, a T-network electrical model is derived based on a two-current-probe impedance measurement approach to reflect

such variations. With the two-current-probe approach, the source model of any PLC modem could also be developed. The full equivalent circuit model, which emulates the connection between PLC modem and the power line network, allows one to quantify the CM suppression performance of a built-in optical coupler. The full equivalent circuit model also serve as a useful input to quickly verify if a PLC modem developed in the laboratory is EMI compliant at the earliest stage of its design without having to conduct expensive field measurements. More measurements will be conducted at different residential and commercial buildings to obtain a good statistical equivalent circuit model that reflects the in-house LVDN conditions in Singapore.

Chapter 6

CONCLUSIONS AND RECOMMENDATIONS

6.1 CONCLUSIONS

As electromagnetic interference (EMI) is the major obstacle for the widespread use of broadband Power Line Communications (PLC) technology, PLC technology developers and vendors must overcome this barrier before their PLC home networking devices and products can be marketed internationally. This research is, therefore, an effort to implement common mode (CM) current mitigation methods, which include the implementation of optical couplers and CM chokes, in PLC modems to suppress the generation of CM currents at the signal interface of these modems and the indoor low voltage distribution network (LVDN). However, without an in-depth understanding of the mechanisms behind the generation of the CM currents, EMI issues involving these modems and their associated power line networks are usually resolved by trial-and-error approaches. Thus, the countermeasures developed are often expensive to implement especially when the PLC system developed is already operational in the field. Hence, reduction of EMI by identifying the root of the problem is a rational and economical approach and this can be achieved by the development of equivalent circuit models for the PLC system.

The possibility of implementing an inexpensive CM choke at the output of the analog front end (AFE) module of a commercially-available PLC modem for the suppression of CM current is studied. The CM and DM equivalent circuit models for the PLC modems are developed to allow the author to analyze and compare the CM and DM signal attenuation performance of the PLC modem with the in-built CM choke. Using the two-current-probe method, characterizations of the PLC system, which consist of the PLC modem and the associated power line network, in terms of the CM and the DM equivalent circuits while they are in operational mode can be performed safely and accurately. Both the equivalent circuit models and the measurement results obtained have demonstrated the dominant effects of the CM impedance of the PLC modem on the CM current suppression capability of the CM choke. Since the CM source impedance of the PLC modem is very large with respect to the CM impedance

of the CM choke, the CM current suppression power of the choke is smaller. On the other hand, the DM impedance of the CM choke, which is due to the existence of leakage inductance, is significantly larger than the DM source impedance of the PLC modem. Hence, the CM choke has attenuated the useful DM signal effectively. Hence, the author has concluded that CM chokes are not feasible to be implemented in a PLC modem since the performance of the CM choke is highly dependent on the DM and CM impedances of the PLC modem and to a certain extent the power line network.

Therefore, the possibility of implementing an optical coupler within the AFE module of a PLC modem for the suppression of CM current is also studied. The implementation of an optical coupler within the AFE module is expected to provide higher galvanic isolation against CM current by increasing the effective CM source impedance of the modem. An optically-coupled evaluation board, which is able to perform power line interface functions similar to the AFE module in the commercially-available PLC modems, is designed and developed to test this concept. To quantify the CM noise source impedance of the PLC modem with in-built optical coupler, the two-current-probe measurement methodology is employed. Comparative studies are also made between the optical coupler and the in-built isolation transformer. CM equivalent circuit models for the evaluation board are also established to allow the author to analyze the effects of parasitic capacitances between the electronic modules, the metal chassis and the ground conductor on the CM source impedance of the evaluation board. From these equivalent circuit models, the author has been able to identify various parasitic capacitances that are introduced by the digital and analog processing circuits of the evaluation board. These parasitic capacitances reduce the effective CM source impedance of the evaluation board and thus allowing the CM currents to propagate. These findings would definitely allow PCB designers to reduce the coupling effects of the parasitic capacitances in the early stages of the PLC modem design without employing trial-and-error approach. From the preliminary measurement results obtained, the author has shown that if properly designed, the inherent low parasitic capacitance of the optical coupler could provide better galvanic isolation against the parasitic capacitances over the conventional transformer coupler where the winding-to-winding capacitance is generally higher.

The proposed optical-based isolation technique is, therefore, a promising candidate for a new EMI mitigation methodology in the PLC system.

However, even with well-designed couplers in PLC modems, the high electrical unbalances in the power line network would still give rise to CM currents. Hence, a full equivalent circuit model, which reflects the unbalance properties of the in-house power line networks in Singapore, is developed to understand their contribution to the generation of the CM current. Since the unbalance characteristic of the power line network varies from places to places, the equivalent circuit model is developed based on the statistical characterization of the indoor power line network in different residential buildings across Singapore using the two-current-probe approach so that such variations are included. Measurement results have shown that the unbalance levels of the in-door power line network for the residential units in Singapore are generally very high and are highly dependent on the power line cabling layout, the power network geometry and the electrical loads connected. Based on these data, the full equivalent circuit model, which emulates the connection between a PLC modem and the actual power line network, is developed. The full equivalent circuit model also aims to establish a mathematical relation between the DM output voltage of the PLC modem, the characteristics of the power line network - in terms of the DM and CM impedances, and the unbalance levels - and the CM current generated at the signal interfaces. The full equivalent circuit model would allow the author to quantify the CM suppression performance of a built-in optical coupler of the evaluation board in the actual AC mains without having to conduct field measurements. Then, design guidelines such as a limit for the DM output voltage and the level of CM source impedance introduced by the optical coupler can be imposed on the PLC modem. From the equivalent circuits developed for the LVDN in the residential units, the CM currents generated by the evaluation board with the in-built optical coupler can be reduced further by reducing the coupling effects of the parasitic capacitances to improve the overall effective CM source impedance of the evaluation board, or by reducing the output voltage of the DM signal. The full equivalent circuit model thus provides a complete picture of the CM current generation mechanism of the PLC system in the indoor power line network.

6.2 RECOMMENDATIONS FOR FURTHER RESEARCH

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

1. An EMI mitigation method of introducing an auxiliary signal to complement the existing PLC signal on the power line network was recently proposed in [60]. The idea is to take advantage of the presence of the ground conductor to inject a 180° out-of-phase version of the useful PLC signal into the neutral-to-ground wires as shown in Fig. 6.1.

Fig. 6.1. Injection of PLC and auxiliary signals using constant voltage sources

If all the three wires (live, neutral and ground) run parallel and in close proximity to one another, the expectation is that the radiated fields from the phase-to-ground PLC signals V_{plc} and neutral-to-ground auxiliary signals V_{comp} would have similar amplitudes and that they will be approximately 180° out of phase with each other. Under these conditions, destructive interference should lead to partial or total cancellation of the radiated fields. Experimental tests carried out on simple network geometries, which consist of a single power line cable with a single load terminated at the far end, show that reductions in the emissions of more than 20 dB can be achieved [61] with such signal injection technique. However, measurements made on more realistic geometry networks showed less good results and even a worsening of the emissions level for certain frequencies at which the interference between the radiation from the wanted PLC signals V_{plc} and from the auxiliary signals V_{comp} appeared to be constructive. This is due to the fact that the generation of the current I_{cm2} needed to complement the current I_{cm1} for the complete cancellations of the radiated fields is too dependant on the impedance and the unbalance level of the power line networks.

An alternative way of injecting the PLC and auxiliary signals by means of RF constant current sources as shown in Fig. 6.2 is proposed to ensure that the generation of the currents: I_{cm2} and I_{cm1} remains constant and independent of the impedance and the unbalance level of the power line networks for the frequency of 3 to 30 MHz to ensure complete cancellations of the radiated fields.

Fig. 6.2. Injection of PLC and auxiliary signals using constant current sources

Thus a feasibility study of implementing such signal-injection techniques needs to be conducted to ensure the suppression of EMI radiations from PLC systems. Further work should also be carried out to extend this technique to a more complex power line network that involves multiple interconnecting power lines.

2. Further work should be carried out to extend the mathematical relation between the DM output voltage of the PLC modem, the unbalance characteristics of the power line network, the CM current generated and the resultant radiated fields to a more complex power line network that involves multiple interconnecting power lines for an accurate estimation of the CM current and the radiated emission. The in-depth analysis of the effects of long cable and different network configurations on the radiated emissions should, therefore, be the topic of a further set of studies. Further work also need to be carried out to establish CM noise propagation model of more complex PLC network that involves multiple interconnecting power lines, which are characterized by discontinuities, line bends, and non-horizontal sections. With the knowledge of the CM noise propagation model of the power line, the level of the CM current on the power line for different cable lengths and different loading conditions can be estimated with reasonable accuracy. The radiation model derived would provide a complete picture of the EMI radiation mechanism of the PLC network.

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PUBLICATIONS

Some of the research work presented in this thesis has been reported in the following international refereed journal and conference publications:

- [1] K.Y. See, A. Kamarul, and P.L. So, “Feasibility Study of Adding a Common-Mode Choke in PLC Modem for EMI Suppression,” accepted for future publication in *IEEE Transactions on Power Delivery*.
- [2] K.Y. See, A. Kamarul, and P.L. So, “Reduction of CM Noise Emissions from PLC Modem Using Optically Coupled Signaling,” in *IEEE Transactions on Electromagnetic Compatibility*, vol. 48, no. 4, pp. 648-653, Nov 2006.
- [3] A. Kamarul, K.Y. See, and P.L. So, “Feasibility Study of Optical Isolator in Reducing CM Conducted Emission from PLC Modem,” in *Proceedings of the 17th International Zurich Symposium on Electromagnetic Compatibility*, Singapore, pp. 424-427, Feb 2006.
- [4] K.Y. See, A. Kamarul, and P.L. So, “Longitudinal Conversion Loss of Power Line Network for Typical Singapore Household,” in *Proceedings of the 7th International Power Engineering Conference*, Singapore, pp. 1-6, Nov 2005.
- [5] K.Y. See, P.L. So, A. Kamarul, and E. Gunawan, “Radio Frequency Common Mode Noise Propagation Model for Power Line Cable,” in *IEEE Transactions on Power Delivery*, vol. 20, no. 4, pp. 2443-2449, Oct 2005.

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