Western University Scholarship@Western

Electronic Thesis and Dissertation Repository

7-8-2013 12:00 AM

Modeling and Optimization of Next-Generation Wireless Access Networks

Maysam Mirahmadi The University of Western Ontario

Supervisor Dr. Abdalllah Shami *The University of Western Ontario*

Graduate Program in Electrical and Computer Engineering A thesis submitted in partial fulfillment of the requirements for the degree in Doctor of Philosophy © Maysam Mirahmadi 2013

Follow this and additional works at: https://ir.lib.uwo.ca/etd

Part of the Systems and Communications Commons

Recommended Citation

Mirahmadi, Maysam, "Modeling and Optimization of Next-Generation Wireless Access Networks" (2013). *Electronic Thesis and Dissertation Repository*. 1349. https://ir.lib.uwo.ca/etd/1349

This Dissertation/Thesis is brought to you for free and open access by Scholarship@Western. It has been accepted for inclusion in Electronic Thesis and Dissertation Repository by an authorized administrator of Scholarship@Western. For more information, please contact wlswadmin@uwo.ca.

Modeling and Optimization of Next-Generation Wireless Access Networks

(Thesis format: Integrated-Article)

by

Maysam Mirahmadi

Graduate Program in Electrical and Computer Engineering Dept. of Electrical and Computer Engineering

A thesis submitted in partial fulfillment of the requirements for the degree of Doctor of Philosophy

School of Graduate and Postdoctoral Studies The University of Western Ontario London, Ontario, Canada

© Maysam Mirahmadi 2013

Abstract

The ultimate goal of the next generation access networks is to provide all network users, whether they are fixed or mobile, indoor or outdoor, with high data rate connectivity, while ensuring a high quality of service. In order to realize this ambitious goal, delay, jitter, error rate and packet loss should be minimized: a goal that can only be achieved through integrating different technologies, including passive optical networks, 4th generation wireless networks, and femtocells, among others.

This thesis focuses on medium access control and physical layers of future networks. In this regard, the first part of this thesis discusses techniques to improve the end-to-end quality of service in hybrid optical-wireless networks. In these hybrid networks, users are connected to a wireless base station that relays their data to the core network through an optical connection. Hence, by integrating wireless and optical parts of these networks, a smart scheduler can predict the incoming traffic to the optical network. The prediction data generated herein is then used to propose a traffic-aware dynamic bandwidth assignment algorithm for reducing the end-to-end delay.

The second part of this thesis addresses the challenging problem of interference management in a two-tier macrocell/femtocell network. A high quality, high speed connection for indoor users is ensured only if the network has a high signal to noise ratio. A requirement that can be fulfilled with using femtocells in cellular networks. However, since femtocells generate harmful interference to macrocell users in proximity of them, careful analysis and realistic models should be developed to manage the introduced interference. Thus, a realistic model for femtocell interference outside suburban houses is proposed and several performance measures, e.g., signal to interference and noise ratio and outage probability are derived mathematically for further analysis.

The quality of service of cellular networks can be degraded by several factors. For example, in industrial environments, simultaneous fading and strong impulsive noise significantly deteriorate the error rate performance. In the third part of this thesis, a technique to improve the bit error rate of orthogonal frequency division multiplexing systems in industrial environments is presented. This system is the most widely used technology in next-generation networks, and is very susceptible to impulsive noise, especially in fading channels. Mathematical analysis proves that the proposed method can effectively mitigate the degradation caused by impulsive noise and significantly improve signal to interference and noise ratio and bit error rate, even in frequency-selective fading channels.

Keywords: Passive Optical Network, Hybrid Wireless-Optical, Dynamic Bandwidth Assignment, Orthogonal Frequency Multiplexing, Interference Management, Femtocell, Building Architecture Model, Impulsive Noise Mitigation, Time-Domain Interleaver.

Acknowledgements

First and foremost, I would like to express my sincere gratitude to my supervisors, Dr. Abdallah Shami and Dr. Arafat Al-Dweik, whom without their wise advice, consistent support and insightful guidance none of the works reported in this thesis would have been possible.

I would also like to thank my examiners: Dr. Serguei Primak, Dr. Vijay Parsa, Dr. Michael Bauer and Dr. Pin-Han Ho for taking the time to review and examine my thesis, and also for their valuable comments.

Special thanks goes to my dear friends at the networking and communications lab: Dr. Dan Dechene, Dr. Dan Wallace, Oscar Filio, Elena Uchiteleva, Karim Hammad, Mohamad Kalil, Mohammad Abusharkh, Manar Jamal, Aidin Rayhani Masouleh, Dr. Andrew Roberts, Dr. Abdessameud Abdelkader, Dr. Abdou Ahmed, Ayman Alghamdi, Mahadevan Balakrishnan, Elham Dolatabadi, Siamack Ghadimi, Sahar Javaher Haghighi, Jamil Hussain, Jason Liu, Dr. Marco Luccini, Khalim Meerja, Jay Nadeau, Abdulfattah Noorwali and Tomasz Rybak, among others.

I am grateful to the staff at the Department of Electrical and Computer Engineering, specially Sandra Vilovski and Chris Marriott who helped me along the way in numerous occasions.

My sincere thanks also belongs to my dear friends: Mohammad, Osam, Vahid, Omid, Seyed Mohammad, Ali, Hossein, Vali, Mahdi, Amirhossein, Ali, Iman, Maysam, Ghodsieh, Maliheh, Nasrin, Samar, Mahdieh and all the members of our gatherings. You have been an essential part of my life in the past few years.

I am thankful to my kind parents and sisters who mean a lot to me. My parents provided me unconditional care and support since my childhood, and I am deeply indebted to them and love them from the bottom of my heart.

Last but by no means least, I would like to thank my dearest wife, Mahta, for her continuous love, encouragement, and patience in all these years. Mahta, you were, are and will be my sole source of happiness and joy and I am grateful for that. To my dear parents,

Mohammad Reza and Masoumeh.

Table of Contents

Abstra	ict.		ii
Acknow	wledge	ments	iv
Dedica	tion .		v
Table o	of Con	tents	vi
List of	Tables	5	x
List of	Figure	$es \ldots \ldots$	xi
Acrony	yms .		ciii
1 Intr 1.1 1.2	roducti Thesis Contri 1.2.1 1.2.2 1.2.3	on Outline Outline Dutline butions of the Thesis Dutline Improving quality of service (QoS) of hybrid optical-wireless networks Dutline Modeling the interference caused by femtocells in pre-planned wireless access networks Dutline Improving the bit error rate performance of OFDM systems in frequency-selective fading channels in presence of the impulsive noise Dutline	1 4 5 5 5
I Im Netwo	iprovi ork	ng Quality of Service in Hybrid Optical Wireless	8
2 Imp wor	oroving ks via	the Quality of Service of Hybrid Optical Wireless Net- a Novel Traffic-Aware Dynamic Bandwidth Assignment	

	• •	
Alg	orithm	9
2.1	Introduction	9
2.2	System Model	11
	2.2.1 Quality of service in hybrid wireless-optical networks	12
2.3	Literature Review	15

2.4	Improving QoS via Anticipated Traffic Information		16
2.5	Impler	Implementing Prediction Mechanism in DBA Algorithms	
	2.5.1	Predicting incoming traffic	20
	2.5.2	Employing predictions in DBA algorithms	24
2.6	Traffic	-Prediction-Assisted Algorithms	26
	2.6.1	Inter-ONU scheduling	26
	2.6.2	Intra-ONU scheduling	28
2.7	Maximum Throughput Analysis		
2.8	Simulation Platform		31
	2.8.1	Traffic generation and types	34
2.9	Simulation Results		36
	2.9.1	Load analysis	37
	2.9.2	EPON cycle length analysis	41
	2.9.3	WiMAX frame length analysis	45
	2.9.4	A realistic scenario	48
2.10	Chapt	er Summary	50

II Managing Femtocell Interference in Preplanned Cellular Network 58

3	Predicting Femtocell Interference for Next-Generation Networks via			
	a N	ovel B	uilding Architecture Model	59
	3.1	Pream	\mathbf{ble}	59
	3.2	Signal	Propagation into Buildings	61
	3.3	Model	ing Methodology	65
	3.4	Const	ructing Femtocell Signal Propagation Model Based on Building	
		Archit	ectural Model (BAM)	67
		3.4.1	Estimating the attenuation with Gaussian-mixture distribution	70
		3.4.2	Recived power distribution	70
	3.5	An Ap	oplication: Optimal FBS Placement	72
	3.6	Chapt	er Summary	74
	.			
4	4 Novel Algorithm for Real-time Procedural Generation of Building			
	Flo	or Plai	ns	78
	4.1	Pream	ıble	78
	4.2	Litera	ture review	79
	4.3	Propo	sed Algorithm for Constructing Floor Plans	80
		4.3.1	Determining the outer shape of the house	82
		4.3.2	Rooms placement	82
		4.3.3	Corridor placement	86
		4.3.4	Placing windows and doors	89
	4.4	Result	ïs	89

	4.5	Chapt	er Summary	92
5	Ind	oor-Ou	utdoor Channel Characterization and its Application to In-	-
	terf	erence	Analyses of Femtocells	95
	5.1	Pream	ble	95
	5.2	Propos	sed Composite Shadowing/Multipath Fading in Mixed Indoor-	
		Outdo	or Environments	98
		5.2.1	A simplified stochastic attenuation model	102
	5.3	Receiv	red Signal Model	105
		5.3.1	OFDM signal model	105
		5.3.2	Macrocell/Femtocell signal model	106
5.4 Performance Analysis of a Macro User in the Presence		mance Analysis of a Macro User in the Presence of Femtocells .	108	
		5.4.1	Femtocell signal approximation using Nakagami-m distribution	108
		5.4.2	PDF of the received signal envelope	109
		5.4.3	Signal to interference ratio (SIR)	111
		5.4.4	Outage probability	116
	5.5	Numer	rical Results	117
	5.6	Chapt	er Summary	121

III Mitigating Impulsive Noise in Frequency-Selective Chan-126

6	6 BER Reduction of OFDM Based Broadband Communication S		uction of OFDM Based Broadband Communication Sys-
tems over Multipath Channels with Impulsive Noise			Multipath Channels with Impulsive Noise
6.1 Preamble			ble
	6.2 OFDM System and Channel Models		I System and Channel Models
		6.2.1	OFDM system model
		6.2.2	Impulsive noise model
	6.3	Propos	sed System Model
	6.4	TDI S	ymbol Blanking $\ldots \ldots 134$
	6.5 System Performance in AWGN Channels		n Performance in AWGN Channels
	6.6 System Performance in Fading Channels using ZF Equalizer		
		6.6.1 No samples blanked, $\epsilon = 0$	
		6.6.2	Samples blanked, $\epsilon > 0$
	6.7	5.7 System Performance in Fading Channels using MMSE Equalizer	
		6.7.1	No samples blanked, $\epsilon = 0$
		6.7.2	Samples blanked, $\epsilon > 0$
	6.8	Numer	ical Results $\ldots \ldots 152$
		6.8.1	AWGN channels 154
		6.8.2	Fading channels
	6.9	Chapte	er Summary

7	Con	clusion and Future Works	169
	7.1	Improving Quality of Service in Hybrid Optical Wireless Network	169
	7.2	Managing Femtocell Interference in Pre-planned Cellular Network	170
	7.3	Mitigating Impulsive Noise in Frequency-Selective Fading Channels .	172
AĮ	open	dices	
Α	Blar	nking Distortion in OFDM Systems	175
В	Sam	ples of Automatically Generated Floor Plans	178
\mathbf{C}	Cur	riculum Vitae	180

List of Tables

2.1	Some example applications and their corresponding EPON QoS class	
	and WiMAX service type	14
2.2	Notations used in Algorithm 1	28
2.3	Notations used in Algorithm 2	29
2.4	Simulation parameters.	33
2.5	Traffic loads in realistic scenario.	34
3.1	Attenuation of walls in different frequencies	64
3.2	GMM paramters	71
4.1	Joint probability of the number of bedrooms (rows) and the number	
	of rooms (columns).	83
4.2	List of the parameters used in generating floor plans	90

List of Figures

Timing diagram of sending a marked packet (a) Normal scheduling (b)		
17		
19		
20		
22		
24		
32		
34		
35		
37		
38		
39		
40		
41		
42		
43		
44		
. 45		
46		
47		
48		
49		
50		
51		
52		
63		
66		
69		
71		
73		
73		

4.1	An aerial photo showing the similarity of homes in a neighbourhood.	83
4.2	Hierarchy graph of a sample house.	85
4.3	A tree graph used by Treemap algorithm and its corresponding rect-	
	angle division	86
4.4	Finding the corridor path.	86
4.5	Room placement steps for the hierarchy tree shown in Fig. 4.2	87
4.6	Optimizing the corridor with different actions.	88
4.7	A sample floor plan automatically generated by the algorithm.	90
4.8	Comparison of a sample generated floor plan with an architecture de-	
	signed floor plan	91
		01
5.1	Possible interferences in a two-tier femtocell/macrocell network [3].	96
5.2	Simulated PDF of the building shadowing $L_{\rm x}$ in the complete model.	99
5.3	A sample FBS propagation contour.	101
5.4	The simplified PDF of the building shadowing L_x	103
5.5	Probability mass function of the number of penetrated internal walls k_2 .	104
5.6	Analytical and simulated PDF of the received signal envelope given in	
	(5.26) for different received mean power of macrocell and femtocell.	111
5.7	The PDF of femtocell signal envelope for different values of K	118
5.8	Comparing the PDF of SIR versus SINR for a single interferer ($\omega = 0.2$)	
	as well as two interferers ($\omega_1 = \omega_2 = 0.1$), and $\Omega = 1. \ldots \ldots$	119
5.9	An outage probability map around three FBSs in a given scenario,	
	calculated by the proposed model	120
5.10	Analytical and simulated outage probability for single and multiple	
	FBSs.	121
5.11	Simulated BER versus SNR in the presence of single FBS using the	
	developed model, outer room model and LOS	122
6.1	The proposed TDI system in fading channels.	134
6.2	BER versus the threshold \mathcal{T}_1 used for sample blanking at different SNRs.	153
6.3	BER versus the threshold \mathcal{T}_1 for different values of \mathcal{T}_2 and SNR	154
6.4	BER performance of variants of the TDI system versus standard OFDM.	155
6.5	Noise enhancement of the TDI and OFDM systems with ZF and MMSE	
	equalizers for $SNR = 15$ dB	156
6.6	Effective SINR of TDI and OFDM for different values of ϵ	157
6.7	BER of the TDI and conventional OFDM using ZF and MMSE equalizers.	158
6.8	BER versus the threshold \mathcal{T}_1 for different values of \mathcal{T}_2 and SNR in	
	fading channels.	159
6.9	BER using ZF equalizer in the presence of IN	160
6.10	BER using MMSE equalizer in the presence of IN	161
6.11	BER of the TDI for different gating probability p	162

Acronyms

\mathbf{AF}	Assured Forwarding
AWGN	Additive White Gaussian Noise
BER	Bit Error Rate
\mathbf{BE}	Best Effort
BS	Base Station
\mathbf{CDF}	Cumulative Distribution Function
CSI	Channel State Information
DBA	Dynamic Bandwidth Assignment
DL	Download
\mathbf{EF}	Expedited Forwarding
EPON	Ethernet Passive Optical Network
FBS	Femtocell Base Station
FDTD	Finite-Difference Time-Domain
FFT	Fast Fourier Transform
FTTH	Fiber To The Home
GPON/GEPON	Gigabit Ethernet Passive Optical Network
HNB	Home NodeB
IEEE	Institute of Electrical and Electronics Engineers
IFFT	Inverse Fast Fourier Transform
IN	Impulsive Noise
IPTV	Intlernet Protocol TeleVision
LAN	Local Area Network
LTE	Long Term Evolution
MAC	Media/Medium Access Control
MCS	Modulation and Coding Scheme
MIMO	Multiple-Input, Multiple Output

MMSE	Minimum Mean Square Error
MPCP	MultiPoint Control Protocol
NSI	Noise State Information
OFDM	Orthogonal Frequency Division Multiplexing
OFDMA	Orthogonal Frequency Division Multiple Access
OLT	Optical Line Terminal
ONU	Optical Network Unit
ONU-BS	Optical Network Unit - Base Station
PAPR	Peak to Average Power Ratio
PDF	Probability Density Function
PHY	Physical Layer
PMP	Point-to-MultiPoint
PON	Passive Optical Network
\mathbf{QAM}	Quadrature Amplitude Modulation
\mathbf{QoS}	Quality of Service
QPSK	Quadrature Phase Shift Keying
RV	Random Variable
SINR	Signal to Interference and Noise Ratio
SIR	Signal to Interference Ratio
SNR	Signal to Noise Ratio
SONET	Synchronous Optical NETwork
TDI	Time Domain Interleaving
UE	User Equipment
UL	Upload
VoD	Video on Demand
VoIP	Voice over Internet Protocol (IP)
WDM-PON	Wavelength Division Multiplexed Passive Optical
	Network
WHT	Walsh-Hadamard Transform
WiMAX	Worldwide Interoperability for Microwave Access
\mathbf{ZF}	Zero Forcing

Chapter 1 Introduction

The modern society vastly depends on information and communication technologies. All kinds of information is digitized, and modern communication technologies remove the time and space restrictions on human activities. Having a personal communication device is a must, not a luxury anymore. Mobile communication is an essential part of our lives. It is widespread and common, yet transparent as air. The image of a person typing on a text-based, bulky computer is ancient. Today, people are using their computing machines at the palm of their hands. These newly emerged technologies provide a multimedia user experience that a while ago seemed impossible. With a few touches, users tap into colossal amount of information that is brought to their devices with the speed of light.

The key part in all these technologies is a broadband access network capable of ensuring Quality of Service (QoS) for different types of traffic. Foreseeing this, International Telecommunication Union-Radio communication/International Mobile Telecommunication (ITU-R/IMT) devises a target for 4th generation systems which is called IMT-advanced. An IMT-advanced network will be capable of over-the-air transfer in rates that exceeds 1 Gbps, thus supports various high-quality and high data rate applications and services. It will transfer different types of traffic, such as data, voice, and video on an all-IP infrastructure, and it is also designed to serve mobile users, as well as fixed users.

Two potential technologies are competing towards IMT-Advanced goals. Worldwide Interoperability for Microwave Access (WiMAX) was the first to enter the competition. It was originally designed to provide broadband access for fixed users. The mobility option is added later via the IEEE 802.16m standard. The second competitor is Long Term Evolution (LTE) which is the continuation of current 3G technology. To be able to fulfil the promises of IMT-Advanced, both WiMAX and LTE have to include a wide range of technologies. Some of these technologies are: smart antenna, multiple-input and multiple-output (MIMO), relay network and carrier aggregation. They should also consider working in different settings and use the spectrum flexibly. Being able to use multiple bandwidth settings and different multiplexing technologies, such as TDM or FDM, are just examples of the required flexibility. Having access to all these technologies, a 4G operator can easily configure the settings to achieve the required performance.

Another mandate for 4G technologies to achieve a high performance is their capability to seamlessly integrate with other access and distribution layer technologies. As an instance, most operators use optical networks to connect the wireless base stations to the core network. The ability to efficiently exploit the features of optical networks to improve the end-to-end QoS would result in a distinguished advantage for the technology. The work presented in Chapter 2 addresses the issue of improving end-to-end QoS for future wireless access networks, the backbone of which is typically composed of a Passive Optical Network (PON). PONs are proved to be a smart choice for backbone data networks, since they are capable of transferring massive amounts of data, satisfactory for the everyday growing demand of recent applications, and yet they have less maintenance cost in comparison to other wired networks.

The integration of optical and wireless networks offers an inexpensive broadband solution. In a hybrid network, such as the one shown in Figure 2.1, the optical part is used to connect several wireless base stations as well as broadband customers to the central office. The abundant bandwidth of optical network is divided among the wireless base stations to serve their wireless users. In addition, hybrid opticalwireless networks offer several value-added features, such as redundancy, mobile access and cooperative communication which make it more appealing. The hybrid opticalwireless networks seem to be the technology of future, but some challenging issues still need to be addressed. In Chapter 2, we focus on ensuring the QoS over a hybrid network, which is vital to offer a satisfactory service.

One of the next generation network objectives is to provide high quality service to the users regardless of their location. Today's wireless users generate 70% of data traffic and initiate 50% of voice calls within buildings [1]. This massive amount of indoor traffic puts a high pressure on next-generation technologies to develop revolutionary ideas to serve indoor users, which is very challenging due to the signal attenuation caused by traversing the walls and buildings' structure and their complex effect on the signal. Traditional solution of increasing the transmission power by a margin to counteract these effects is not an option anymore. This is mainly because doing that, increases the interference power in the current interference-limited networks, and also the available power for mobile devices is limited.

The remedy of choice by most next-generation networks is to rely on low-power indoor transmitters, called femtocell base station (FBS), to serve indoor users. These FBSs are connected to the distribution layer of the operator via a broadband Internet connection and share the medium with normal Internet traffic. The introduction of the femtocell paradigm to the network generates several challenges such as interference management, security, handover strategies, etc. Interference management is one of most important challenges, since it affects not only the indoor users that are served by FBSs, but also legacy macrocell users that are in proximity of those FBSs. Thus, improving the total performance of the network requires intelligent interference management schemes and realistic models for the femtocell interference.

Traditional wireless networks are either concerned with penetration of signal into buildings, such as legacy cellular networks, or covering the whole building area, like wireless LANs. Hence, a realistic model for describing the propagation of a signal from an indoor transmitter to outside, which is the main interest in case of femtocells, is left undeveloped. The typical research approach to study the femtocell interference is to choose a sample scenario with one or more buildings and predefined floor plan [2,3]. The results of this approach clearly depend on the selected floor plan and may considerably vary with it. This motivates us to develop a model, as realistic as possible, for this exact application.

The developed model is then used as the basis for a composite shadowing/fading model that jointly considers the effects of signal propagation through building structures and multipath fading. Moreover, several performance criteria of a two-tier femtocell/macrocell network such as signal to interference and noise ratio and outage probability are analyzed using the proposed model. The results reveal that in some scenarios, commonly used models for femtocell studies fail to grasp the complex effects of signal propagation in buildings and their predictions are far from reality.

One of the design goals of the next-generation networks is to provide high quality service anytime and anywhere. For the next work, we turn our attention to industrial environments where impulsive noise and multipath fading are main culprits in reduced QoS. Almost all of the proposed next-generation networks, employ a variant of Orthogonal Frequency Division Multiplexing (OFDM), which is proved to be susceptible to impulsive noise (IN). In presence of IN, the OFDM performance is limited by a high error floor. Therefore, in Chapter 6, we first model the effect of simultaneous fading and IN on OFDM systems, and later propose a technique to improve the performance of such systems with minimum added complexity. The proposed technique can be applied to any OFDM-based system that suffer from IN. For example, recent Power Line Communication (PLC) standard added an OFDMbased mode. Powerline communication is a prime example for channels with IN, and hence, the proposed technique can be used to mitigate the adverse effect of IN on such channels.

1.1 Thesis Outline

The works proposed in this thesis consists of five chapters, each is focused on a different part of the next generation network. The first part, which is described in Chapter 2, focuses on improving the quality of service in backbone part of the network. The next three chapters consider the performance of a two-tier femtocell/macrocell network. Chapter 3 presents a model for the femtocell interference based on a novel Building Architecture Model (BAM) described in Chapter 4. In Chapter 5, Some performance metrics, such as received Signal to Interference and Noise ratio (SINR) and outage probability is derived based on the proposed model in Chapter 4. The work presented in Chapter 6 considers the effect of Impulsive Noise (IN) on the performance of the network and present an innovative time-domain interleaving technique for OFDM systems that can combat the IN effects.

1.2 Contributions of the Thesis

The contributions of the thesis is summarized in the following list.

1.2.1 Improving quality of service (QoS) of hybrid optical-wireless networks

- First: A technique to extract information on incoming traffic from the wireless scheduler is formulated.
- Second: A Dynamic Bandwidth Assignment (DBA) algorithm for the upload channel of Ethernet PON (EPON) is developed.

1.2.2 Modeling the interference caused by femtocells in pre-planned wireless access networks

- First: An algorithm to generate random floor plans is developed.
- Second: A technique to stochastically model the randomness in building architectures is proposed.
- Third: An indoor-outdoor channel model to evaluate the interference caused by femtocell transmitters is developed.
- Fourth: Based on the proposed model, several performance metrics, including received SINR and outage probability is evaluated.

1.2.3 Improving the bit error rate performance of OFDM systems in frequency-selective fading channels in presence of the impulsive noise

First: The effects of IN in AWGN and frequency-selective channels are modeled.

Second: A system to improve the performance of OFDM-based systems in presence of IN is proposed.

Third: The performance of the proposed system in terms of received SINR and Bit Error Rate (BER) is evaluated.

References

- V. Chandrasekhar and J. Andrews, "Uplink capacity and interference avoidance for two-tier femtocell networks," *IEEE Trans. Wireless Commun.*, vol. 8, no. 7, pp. 3498–3509, Jul. 2009.
- [2] Technical Specification Group Radio Access Networks (TSG-RAN), "3G Home NodeB study item technical report (release 8)," 3GPP, Tech. Rep. TR 25.820, 2008.
- [3] H. Claussen, L. Ho, and L. Samuel, "An overview of the femtocell concept," Bell Labs Tech. J., vol. 13, pp. 221–245, 2008.

Part I

Improving Quality of Service in Hybrid Optical Wireless Network

Chapter 2 Improving the Quality of Service of Hybrid Optical Wireless Networks via a Novel Traffic-Aware Dynamic Bandwidth Assignment Algorithm

2.1 Introduction

In recent years, numerous bandwidth consuming applications have emerged. These applications like Internet Protocol TeleVision (IPTV), online gaming and Video-on-Demand (VoD), require considerable bandwidth and are attracting widespread attention among consumers. Moreover, service providers increasingly show interest and intention to offer integrated services, comprising of voice, video, data and even wireless services, to their customers based on an all-IP shared network infrastructure. As a result, a broadband access network capable of ensuring Quality of Service (QoS) for different service types is required.

Optical fiber networks are capable of transferring massive amounts of data, satisfactory for the everyday growing demand of recent applications. Modern passive optical networks, such as 10-Gigabit Ethernet Passive Optical Network (10G-EPON) or Wavelength Division Multiplexed Passive Optical Network (WDM-PON) can transfer several gigabits of data per second and yet they have less maintenance cost in comparison to old optical networks, like Synchronous Optical NETwork (SONET).

A version of this chapter has been published in [1]. The work has been partly presented in [2]. $[\bigcirc 2011 \text{ IEEE}]$

Nevertheless, it is still quite expensive to lay fiber to every user's premise and build a purely optical network. In addition to the cost problem, it is impractical to build a true optical network, such as Fiber To The Home (FTTH), in a congested and built-up urban area, where considerable number of broadband users are.

On the other hand, wireless networks are relatively inexpensive and have the unique feature of delivering data to mobile users. The drawbacks are that they offer much less bandwidth and they are error-prone. Another disadvantage of wireless networks is that the wireless spectrum is shared among many users, further limiting the bandwidth offered to each user.

The integration of optical and wireless networks offers an inexpensive broadband solution. In a hybrid network, such as the one shown in Fig. 2.1, the optical part is used to connect several wireless base stations as well as broadband customers to the central office. The abundant bandwidth of optical network is divided among the wireless base stations to serve their wireless users. In addition, hybrid optical-wireless networks offer several value-added features, such as redundancy, mobile access and cooperative communication which make it more appealing.

The hybrid optical-wireless networks seem to be the technology of future but some challenging issues still need to be addressed. Here, we focus on ensuring the QoS over a hybrid network, which is vital to offer a satisfactory service. The choice of how the scheduler assigns bandwidth to different flows has a tremendous impact on the QoS parameters. The scheduler typically relies on a Dynamic Bandwidth Assignment (DBA) algorithm to divide the available bandwidth between different flows.

In this work, we propose a method to enhance DBA algorithms in the optical part of hybrid optical-wireless networks. A prediction method, which uses the information in the wireless domain, helps decrease the delay bound of delay-sensitive flows, and hence results in improved QoS. The extensive simulation-based studies not only shed light on the factors, affecting the proposed algorithm, but also compare fixed and adaptive cycle length optical DBA algorithms side by side.

The content of this chapter addresses the integration of Worldwide Interoperability for Microwave Access (WiMAX) and Ethernet Passive Optical Network (EPON) as successful and relatively inexpensive technologies of their kind. However, most of the discussions can be applied to any hybrid optical-wireless network.

The chapter is organized as follows. Section 2.2 introduces the system model. Section 2.3 compares advantages and disadvantages of integration in hybrid opticalwireless networks. The proposed algorithm is presented in Section 2.5 and the maximum throughput of optical-wireless network is analytically studied in Section 2.7. The traffic prediction assisted algorithms, used in simulation experiments, are explained in Section 2.6, while Section 2.8 defines the simulation platform. Section 2.9 analyzes the simulation results and finally Section 2.10 concludes the chapter.

2.2 System Model

The topology of hybrid optical-wireless network is commonly considered to be tree [3–10].

The root of the tree is the Optical Line Terminal (OLT) which is located at the central office (CO) and is directly connected to the core network. Each Optical Network Unit (ONU), also called gateway [9], is connected to the OLT through one or several splitters, as shown in Fig. 2.1. WiMAX Base Stations (BS) are connected to an ONU via a standard interface, like Ethernet, or even implemented in the same box. We refer to the latter case as ONU-BS. In practice, the WiMAX BSs share the optical bandwidth with other users, such as residential or corporate local area networks.

The WiMAX BSs provide wireless connectivity to their subscriber stations (SS) through either a single hop legacy wireless network or a relay-based wireless network with mesh topology. In the latter case a routing algorithm, like the one presented in [9], is necessary to effectively deliver the packets.

Due to the intrinsic property of passive optical networks, downstream data, transmitted from OLT to ONU, is broadcasted to all ONUs. By contrast, upstream data, transmitted from the ONUs can only be received at the OLT. Thus, the downstream and upstream links are considered as Point-to-MultiPoint (PMP) and pointto-point (PPP), respectively.

In cases where the wireless base station and the OLT are collocated, the wireless



Figure 2.1: Integrated EPON and WiMAX architectures [3]. ©2007 IEEE

and optical parts can be integrated, resulting in an intelligent network. The key point in the integrated architecture is that the BS and the ONU can share their internal data, such as detailed information about bandwidth requests, bandwidth allocation and packet scheduling, which can be used to improve the overall performance. An intelligent scheduler at the integrated ONU-BS should leverage this extensive information to achieve improved performance and QoS guarantee.

2.2.1 Quality of service in hybrid wireless-optical networks

Quality of service is one of the major research areas in network-relates topics. It is also one of the most challenging ones, as a set of new challenges and requirements come along with each new application and for every emerging technology. To provide good quality of service, the application requirements should be satisfied with as less resources as possible. The Best Effort (BE) method of delivery inherited from the original Internet design is not suitable for all of the applications, and to ensure the quality of service, special designs are required.

The quality of service requirements of each application is typically described by the following parameters [11].

• Throughput: The effective number of data units that are transported in the

unit time.

- **Delay:** The time interval between the departure of last data bit from the source and its arrival at the destination. If original source and final destination are considered, it is called end-to-end delay. To focus on different network layers or sections of the network, various types of delays are typically measured.
- **Jitter:** The delay variation that is caused by traveling through nodes in the network.
- Loss: The percentage of the transmitted packets that are not received in predefined period of time.

Modern networks have a specific design, namely QoS framework, to satisfy the user/application requirements referred to as Service Level Agreement (SLA) or service commitment. In QoS framework a combination of scheduling, buffer management and policing algorithms are implemented to satisfy the service commitment.

Guaranteeing quality of service in a hybrid wireless-optical network is a formidable task. The two domains, *i.e.*, wireless and optical, have their own characteristics and face separate challenges. Some of these differences are listed as follows:

- The wireless link is much more susceptible to noise than optical link.
- The data rate of wireless network is much less than optical network.
- Unlike the optical network for which the state and attenuation of each ONUs is almost constant, wireless environment is dynamic and its characteristics are changing quite fast.
- WiMAX implements IntServ QoS framework, whereas EPON implements Diff-Serv.
- Traffic classes are defined differently in these networks.

In this study, we selected EPON as the optical network and WiMAX as its wireless counterpart due to their unique features. These two technologies have several similarities that work in our advantage; e.g., they both support QoS, and have a request/grant mechanism for assigning bandwidth to different users integrated within. Nevertheless, there are also several major differences that need to be addressed.

In EPON, the implementation details such as the number of service classes and their QoS parameters are different in each implementation. However, they all Table 2.1: Some example applications and their corresponding EPON QoS class and WiMAX service type

Application	EPON QoS class	WiMAX service type
Web browsing, email	BE	BE
File transfer	BE	nrtPS
Streaming audio/video	AF	rtPS
VoIP	EF	ertPS / UGS

recognize the types of service defined in IETF standard. The most well-known and typically implemented traffic classes, in order of their priorities, are Expedited Forwarding (EF), Assured Forwarding (AF) and Best Effort (BE).

WiMAX bandwidth management adapts a different approach. Five different classes of service are defined [12].

- Unsolicited Granted Service (UGS)
- extended real-time Polling Service (ertPS)
- real-time Polling Service (rtPS)
- non real-time Polling Service (nrtPS)
- Best Effort (BE)

In WiMAX, each flow has to initialize the session and specify its type of service. Depending on the requested type of service, it may need to request for bandwidth in each frame. UGS services specify their bandwidth requirement in the initialization phase. They are the highest priority class of service. On the other hand, polling services request for bandwidth in each frame. Scheduling algorithm processes these requests and grants the available bandwidth to them. The least important type of traffic is BE, which competes for the required bandwidth. WiMAX Scheduler processes all of these requests and grants the available bandwidth is then broadcasted to SSs. A number of applications along with their corresponding WiMAX type of service are listed in Table 2.1.

2.3 Literature Review

Researchers have proposed a number of methods to integrate wireless networks, such as WiMAX or LTE, with passive optical networks. The idea of using the information already available at the wireless BS to assist the optical part is first introduced in [3]. But no specific methods or performance evaluation has been done. To the best of our knowledge, none of the researchers devised a comprehensive method to use the information to improve the performance of the optical domain, and hence improve the overall performance.

Some addressed hybrid optical-wireless network by proposing a single MAC frame and QoS structure on both domains [13, 14] while others translated one technology's QoS parameters into another's [5–7, 10, 15, 16]. Due to the differences between optical and wireless systems, such as their scalability requirements, it is too difficult to propose a single QoS structure for both technologies without sacrificing scalability.

Sarkar *et al.* presented an architecture called Wireless-Optical Broadband Access Network (WOBAN) [17–24]; This work proposed several approaches on the placement of ONU-BS nodes. The work presented in [14] addressed the integration of WiMAX and Optical Burst Switching (OBS) and evaluated the performance for BE and UGS traffic flows. Another architecture, namely MARIN, that incorporates multiple OLTs in the network was proposed in [25].

Ou *et al.* in [5] devised a method to dynamically map the WiMAX classes to the Gigabit PON (GPON) classes. The design was implemented using commercially available devices to deliver video on demand service to test users. The work was later enhanced in [7] and an excess distribution DBA scheme was adopted. Yan *et al.* proposed a DBA algorithm that takes channel conditions into account [8]. Another channel aware algorithm was presented in [6]. The proposed scheduling algorithm considers the wireless condition of the cell and head of line delay of rtPS queue as well as the queue length. The authors of [10] proposed a two-step mapping that combines BE and nrtPS into a priority queue and rtPS and ertPS into another one and then maps them to EPON classes of traffic. Obele *et al.* in [13, 15] proposed an architecture based on a unified QoS infrastructure which is similar to IEEE 802.16a. The performance of their proposed architecture was analyzed under self-similar and long-reach-dependent traffic loads. Wang *et al.* in [16] and Luo in [26,27] also used a similar unified QoS scheme throughout the network. The work described in [4] used the next EPON polling time to control the admission of delay-sensitive traffic.

2.4 Improving QoS via Anticipated Traffic Information

Amongst the various challenges of ensuring QoS in EPON, the dynamic bandwidth assignment of the upload channel attracts more research attention, due to its inherent limitations. Unlike the download channel where the bandwidth assigner has full knowledge of the bandwidth needs, the upload bandwidth assigner, which is located at the OLT is not aware of real time bandwidth requirements of each ONU. Therefore, a mechanism is required to report real time bandwidth requirements of each individual ONU to the OLT.

To exchange the bandwidth requirement information, a request/grant mechanism, called Multi-Point Control Protocol (MPCP), is used. In MPCP, ONUs send information regarding their bandwidth requirements to the OLT. For example, in a widely used method, ONUs send their queue length to the OLT at the end of the grant slot. OLT processes the results and sends the bandwidth assignments back to ONUs by grant packets. The problem is that MPCP introduces latency which degrades the performance of the bandwidth assigner. *i.e.*, the bandwidth assigner in OLT can only know the ONU bandwidth requirements with some delay. The MPCP request is sent once in each EPON cycle. As a consequence, the packets that are received after sending the request remain in ONU buffer for an extra EPON cycle.

Using information about the incoming traffic can alleviate the degradation of the extra delay. Integrated WiMAX/EPON ONU-BS provides us with a tool to give this information to OLT. Prior to sending upstream data, the user requests the required wireless bandwidth from the WiMAX BS. The BS scheduler assigns the available bandwidth based on these requests and then sends the grant in UL map



Chapter 2: Improving the Quality of Service of Hybrid Optical Wireless Networks via a Novel Traffic-Aware Dynamic Bandwidth Assignment Algorithm

Figure 2.2: Timing diagram of sending a marked packet (a) Normal scheduling (b) Integrated scheduling.

field of WiMAX frame. Therefore, the BS has full knowledge of the incoming data for at least the next coming frame. If this information becomes accessible to the OLT scheduler, it can be used to predict the incoming traffic which would result in a more effective bandwidth scheduling. However, in order to achieve optimal performance, an intelligent design that considers the EPON cycle and WiMAX frame length as well as the traffic load and classes is required.

Predicting the incoming traffic makes it possible to request the required bandwidth for transmitting a packet that has not yet arrived. At the next EPON cycle, due to pre-assignment of the required bandwidth, the packets that have just arrived at the ONU, are transmitted to the OLT right away without experiencing any additional delay for exchanging MPCP request/grant messages.

Since the WiMAX BS assigns bandwidth to the Mobile Subscribers (MS) at the beginning of each frame, it is aware of the quantity of data to be received during the following frame. Fig. 2.2 compares the process of sending a marked packet in normal and integrated scheduling. As it can be seen in the Fig. 2.2, predicting the arrival of a packet reduces its queuing delay. This helps ensure QoS for delay sensitive applications.

17

In order to use knowledge of incoming traffic to optimize the bandwidth assignment in hybrid optical-wireless networks, the length of the WiMAX frame and EPON cycles should be carefully adjusted. The essential condition for predicting incoming traffic is that the WiMAX frame length should be longer than the EPON cycle length. For better illustration, consider Fig. 2.3(a) in which the WiMAX frame is considerably shorter than EPON cycle. Since the BS knows only the traffic of the current frame, it is not possible to accurately predict the amount of traffic that accumulates until the next EPON cycle. Therefore, only a fraction of incoming traffic can be predicted. Hence, the performance improvement that could have been gained from full knowledge diminishes. By contrast, Fig. 2.3(b) illustrates the case in which WiMAX frame is longer than EPON cycle. Hence, the ONU-BS knows the quantity of data to be sent to the OLT at the next EPON cycle.

The key part of the proposed algorithm is estimating the amount of traffic that an ONU-BS will have to relay to the OLT in the next grant. Without lack of generality, we assume that all received data at the ONU-BS should be relayed to the OLT. *i.e.*, there is no local traffic at ONU-BS. Given that the amount of incoming traffic is known, ONU-BS requests OLT for the amount of bandwidth, denoted by B_r below.

$$B_r = Q + B_p(T_{nx}), \tag{2.1}$$

where Q denotes the current length of the queue and $B_p(T_{nx})$ is the predicted incoming traffic that will be received up to beginning of the next grant, which is represented by T_{nx} .

It is essential to know the time of next transmission to estimate the bandwidth requirement. If EPON uses fixed length cycles, the next time of transmission is simply calculated as

$$T_{nx} = T_q + C, \tag{2.2}$$

where T_g is the grant window beginning time and C is the cycle length. It is worth mentioning that this method does not account for the variations that occur because of different ONU grant sizes in each EPON cycle. A more accurate approach is to



Chapter 2: Improving the Quality of Service of Hybrid Optical Wireless Networks via a Novel Traffic-Aware Dynamic Bandwidth Assignment Algorithm

Figure 2.3: EPON cycle in comparison with WiMAX frames.

(b)

WiMAX frame

calculate T_{nx} based on the internal data of OLT scheduler. This approach is applicable to both fixed-length cycle and adaptive-length cycle algorithms.

2.5 Implementing Prediction Mechanism in DBA Algorithms

The proposed algorithm consists of two parts. The core of the algorithm is the inter-ONU scheduling which is performed at the OLT scheduler. This part of the algorithm does not normally consider different classes of traffic and assigns the bandwidth to each ONU based on the available bandwidth, and ONU requests. The prediction based technique enhances this part and allows it to use traffic prediction as well as ONU requests to fine tune the assigned bandwidth. An example of augmenting an inter-ONU scheduling algorithm with prediction is described in Section 2.5.2.

The second part of the algorithm, namely intra-ONU scheduling, is performed



Figure 2.4: Relationship between F(t) and WiMAX frame.

at each ONU to assign the granted bandwidth to each traffic class. An example of such an algorithm and our proposed modification based on traffic prediction is explained in Section 2.6.2.

The algorithm is independent of the WiMAX BS scheduler. Therefore, it does not impose any requirement on the WiMAX scheduler. The only modification needed at WiMAX BS is that its upload bandwidth assignments, *i.e.*, UL map, should be made accessible to the ONU scheduler. Although choosing the WiMAX scheduler may affect the overall performance, the study of WiMAX scheduling algorithms and their impact is beyond the scope of this work. A comprehensive survey of WiMAX scheduling algorithms can be found in [28].

2.5.1 Predicting incoming traffic

Information about incoming traffic in the next WiMAX frame is abstracted in the function F(t), which is defined as the aggregated traffic at time t. F(t) predicts the ONU queue length at time t. It is an increasing function of time that yields the amount of traffic that has been received up to time t. The relation between WiMAX frame and F(t) is illustrated in Fig. 2.4.

Every time a UL map is generated at the BS, F(t) is defined for the next WiMAX frame. Note that it models the worst case scenario, *i.e.*, maximum aggregated traffic, and depending on the type of the scheduler used in BS, the real incoming traffic can be slightly less.

The summarized information, *i.e.*, F(t), is sent to the OLT along with the

bandwidth requests to be used in the scheduling process. In order to limit message exchanges and keep the OLT scheduler as simple as possible, the information regarding the incoming traffic class is not transmitted to the OLT. The simulations show that for the selected DBA algorithm, it is not required to transmit the class-based information. However, since this approach can be applied to any DBA algorithm, sending the class-based information can be easily implemented, if required. Without considering the classes of traffic, F(t) can be modeled by a linear piecewise function of time.

$$F(t) = \begin{cases} a_0 t + b_0, & t_0 < t \le t_1 \\ a_1 t + b_1, & t_1 < t \le t_2 \\ a_2 t + b_2, & t_2 < t \le t_3 \\ \vdots \\ a_n t + b_n, & t_n < t \le t_{n+1} \end{cases}$$
(2.3)

where a_i and b_i , i = 1, ..., n are coefficients. To simplify the calculations and prohibit time wrapping, time is typically measured from a common origin such as beginning of the cycle. Note that F(t) is only defined in the prediction range $(t_0, t_{n+1}]$ which is discussed more in the next paragraphs. In the case of considering multiple classes of traffic, each class can be described by an equation similar to (2.3).

The slope of F(t) depends on the modulation and coding scheme selected by MSs in the corresponding frame. In scenarios where all of the MSs use the same modulation and coding technique and TDD duplexing is used, F(t) could be simplified to

$$F(t) = \begin{cases} b_0, & t \le t_1 \\ a_1 t + b_1, & t_1 < t \le t_2 \\ b_2, & t_2 < t \le t_3 \end{cases}$$
(2.4)

This case is not unreal and it happens if the network operates at the maximum capacity, where all users transmit at the modulation and coding rate that yields maximum throughput. In addition, the condition has to be met in the scenarios
Chapter 2: Improving the Quality of Service of Hybrid Optical Wireless Networks via a Novel Traffic-Aware Dynamic Bandwidth Assignment Algorithm



Figure 2.5: A representation of F(t): (a) General case, (b) Simplified

where maximum throughput analysis is of interest, which is the common analysis approach for DBA algorithms, since more meaningful insight can be achieved by studying the DBA algorithms under those conditions. An example of F(t) is shown in Fig. 2.5.

The scheduler uses F(t) to estimate the incoming traffic and grant the appropriate amount of bandwidth. Since it is assumed that TDD is used as the duplexing method, there are active and idle periods in the upload channel. Data packets are arriving during the active periods, where SSs are allowed to send. In WiMAX, the active period for upstream is the part of the UL subframe that is granted to the SSs. During idle periods, that BS transmits download data, users are not allowed to send, and hence, F(t) remains constant. Considering this, F(t) is

$$F(t) = \begin{cases} b_0, & t_0 < t \le t_s^1 \\ a_1 t + b_1, & t_s^1 < t \le t_e^1 \\ a_1 t_e^1 + b_1, & t_e^1 < t \le t_s^2 \\ \vdots \\ a_n t_e^n + b_n, & t_e^n < t \le t_s^{n+1} \end{cases}$$
(2.5)

where t_s^i and t_e^i are the beginning and end of ith activity period, respectively. From WiMAX TDD point of view, these are equivalent to beginning and end of the UL slot. Normally, the WiMAX scheduler puts together the grant slot of all users that transmit with the same modulation and coding. It also packs the grant at the beginning of the slot [29]. Therefore, it is not expected to have more than one activity period in the prediction range, which simplifies the transmission of abstract data.

Considering the WiMAX rate, time division duplexing and request scheme, F(t) is

$$F(t) = \begin{cases} Q, & t_r < t \le t_s^1 \\ R_W t + Q - R_W \times min(t_s^1, t_r), & t_s^1 < t \le t_e^1 , \\ R_W t_e^1 + Q - R_W \times min(t_s^1, t_r), & t_e^1 < t < t_s^2 \end{cases}$$
(2.6)

where R_W is the WiMAX transmission rate, Q is the request or equivalently, ONU's queue length at the time of transmitting the request, and t_r is the time of request transmission. F(t) is illustrated in blue in Fig. 2.6. At the time of sending the request, F(t) is initialized to Q. Then it increases with the WiMAX transmission rate when the UL period begins. The increase continues until the end of UL period and then remains constant.

The prediction information is updated regularly in a way that $t_r < t_s^2$, *i.e.*, the updated prediction range overlaps with the previous one. This makes the continuous prediction possible.



Figure 2.6: Illustration of F(t) and request corrections method.

2.5.2 Employing predictions in DBA algorithms

In order to implement the predictions in the bandwidth assignment algorithm, it is necessary to correct the requests to consider not only the packets that were in the queue at the time of transmitting the requests, but also the packet that arrived later. This way, it is possible to estimate the total number of the packets in the queue in near future, including the time that the grant will be received. Also, since the WiMAX frame is supposed to be longer than the EPON cycle (refer to Section 2.4 for the explanation), it is possible that receiving data continues between the grant periods. The ideal solution is to consider all of these newly arrived packets and extend the grant to transmit them, if possible. The grant can be extended until the ONU's transmission buffer becomes empty and all of the newly arrived packets are transmitted.

In our proposed algorithm, F(t) predicts the newly arrived data. In order to consider this prediction in the dynamic bandwidth assignment algorithm, it is required to update the requests, fed into the DBA, with the predictions. Since the correction depends on the bandwidth assignment, *i.e.*, DBA results, it is done iteratively. This way the DBA bandwidth assignments and request corrections are jointly calculated.

The DBA outcome without applying the request corrections is considered as the initial bandwidth assignment. To calculate the request correction, the granted bandwidth is modeled with the same approach as the incoming data. This model, which is denoted by G(t), on the other hand, describes transmitted bits. It is modeled as a line segment which starts from t_g with a slope of R_E . t_g represents the start time of the granted window and R_E represents the EPON transmission rate. These models are illustrated in Fig. 2.6.

$$G(t) = \begin{cases} 0, & t \le t_g \\ R_E \times (t - t_g), & t > t_g \end{cases}$$
(2.7)

The reception prediction function F(t) and the transmission function G(t) coincide with each other at time t_c . This is the time that ONU queue is expected to become empty if enough bandwidth is available. Therefore

$$F(t_c) = G(t_c) \tag{2.8}$$

By substituting F(.) and G(.) from 2.6 and 2.7, respectively and solving the equation for t_c , we have,

$$t_{c} = \begin{cases} t_{g} + \frac{Q}{R_{E}}, & t_{g} \leq t_{s} - \frac{Q}{R_{E}} \\ \frac{R_{W}t_{s} - Q - R_{E}t_{g}}{R_{W} - R_{E}}, & t_{s} - \frac{Q}{R_{E}} < t_{g} < t_{e} - \frac{1}{R_{E}}(Q + R_{W}t_{e} - R_{W}t_{s}), \\ t_{g} + \frac{Q + R_{W}t_{e} - R_{W}t_{s}}{R_{E}}, & t_{g} \geq t_{e} - \frac{1}{R_{E}}(Q + R_{W}t_{e} - R_{W}t_{1}) \end{cases}$$
(2.9)

where t_s and t_e denote the beginning and the end of the next grant slot, respectively. The calculation should be done for all ONUs, before starting a new iteration. Then, the request correction of each ONU is

$$R_{new}^k = R_E \times (t_c^k - t_g^k), \qquad (2.10)$$

where R_{new}^k is the new request for ONU k. t_c^k and t_g^k denote t_c and t_g for ONU k, respectively.

The correction process is repeated for each ONU until the granted bandwidths stall or a predetermined maximum number of iterations is reached. Then, the granted bandwidth is sent to the ONUs along with the grant start times and predicted incoming traffic.

2.6 Traffic-Prediction-Assisted Algorithms

2.6.1 Inter-ONU scheduling

The proposed traffic prediction mechanism can be applied to any DBA algorithm. Since it provides the DBA algorithms with real-time information of traffic, it generally improves the delay performance. It is worth mentioning that the most important modifications have to be done in the inter-ONU scheduling algorithms, which are discussed here. There are two approaches among DBA algorithms regarding EPON cycle length, which is a significant factor in the performance of the proposed method. Some DBA algorithms, such as the well-known Interleaved Polling with Adaptive Cycle Timing (IPACT) [30], use dynamic cycle length while others, such as Excess Distribution [31], adapt fixed cycle length. In order to investigate the proposed method's performance in both categories, the proposed method is applied to a DBA algorithm from each category.

The first DBA algorithm is Excess Distribution which is first proposed in [31] and later improved in [32]. This DBA scheme first divides the available bandwidth equally among the ONUs to define the threshold window. All of the ONU's requests that are less than the threshold are granted. Then, the unused remaining bandwidth is distributed equally between the unsatisfied ONUs. The variant used in this study adapts a two-stage scheduling algorithm, as described in Section 2.6.2, to allocate bandwidth to each of the traffic classes. This algorithm is referred to as Excess Distribution or ED, while the traffic-prediction-assisted version of it, is referred to as Integrated ED. The psuedo-code of the integrated ED is shown in Algorithm 1. It demonstrates how the prediction can be applied to a DBA algorithm. The notations used in the pseudo-code are explained in Table 2.2. Note that the time frame in each ONU and OLT is different due to propagation delay. For the sake of simplicity, this effect is not considered in the pseudo-code, but it is considered in simulations.

Unlike the first algorithm that maintains the fixed EPON cycle length, the second algorithm, namely CB-IPACT, adjusts EPON cycle length according to the introduced traffic load. Another difference is that CB-IPACT grants the requests on-the-fly, *i.e.*, it processes each request independent of others. CB-IPACT grants

Algorithm 1 proposed Inter-ONU DBA pseudo-code

Require: Q(i), i = 1, ..., Nfor i := 1 to N do 1: 2: $R(i,1) \leftarrow Q(i)$ 3: end for for k := 1 to Ite_{max} do 4: $R_{ex}^t \leftarrow 0$ 5: $B_{ex}^{\breve{t}^{w}} \leftarrow 0$ 6: 7: for i := 1 to N do if $R(i,k) > W_{max}$ then 8: $R_{ex}(i) \leftarrow R(i,k) - W_{max}$ 9: $\vec{R_{ex}^t} \leftarrow R_{ex}^t + R_{ex}(i)$ 10: 11: else $B_{ex}^t \leftarrow B_{ex}^t + W_{max} - (R(i,k))$ 12:end if 13:end for 14:for i := 1 to N do 15:if $B_{ex}^t > R_{ex}^t$ then 16: $G(i) \leftarrow R(i,k) + \frac{B_{ex}^t - R_{ex}^t}{N}$ 17:else 18:19:if $R(i,k) < W_{max}$ then 20: $G(i) \leftarrow R(i,k)$ else 21: $G(i) \leftarrow W_{max} + \frac{R_{ex}(i)}{R_{ex}^t} \times B_{ex}^t$ 22: 23: end if end if 24:end for 25:26:Update T_{sch} Update $P(i, G, T_{sch}, F_i(t))$ 27: $R(i, k+1) \leftarrow R(i, k) + P(i, G, Tsch)$ 28:29: end for Update W_{max}

Notation	Description
N	Number of ONUs
R(i,k)	Request of ONU i in iteration k
Q(i)	Queue Length at the beginning of the grant
Itemax	Maximum iteration for solving the bandwidth allocation
R_{ex}^t	Total extra requested bandwidth
B_{ex}^t	Total extra bandwidth available
G(i)	Granted bandwidth to ONU i
$P(i, G, T_{sch})$	Prediction of incoming traffic for ONU i according to
	granted time slots. Calculated according to (2.9)

Table 2.2: Notations used in Algorithm 1.

the request if it is less than a predefined maximum grant. Otherwise, it grants the maximum grant.

$$Grant = min(Request, MaxGrant)$$
 (2.11)

Applying traffic prediction to this algorithm by the method, explained in Section 2.5, is fairly simple. The traffic-prediction-assisted CB-IPACT is referred to as Integrated-CBIPACT. Then, the two-stage approach, explained in Section 2.6.2, is employed to distribute the granted bandwidth to different traffic classes.

2.6.2 Intra-ONU scheduling

After the available bandwidth is divided between the ONUs, the intra scheduler at ONU has to divide the granted bandwidth between different traffic classes. Here, we adopt strict priority in the ONU scheduler. Therefore, the ONU scheduler serves EF, the highest priority class first. Then, it serves AF and after that BE queue will be served.

When strict priority is applied in the ONU scheduler, high priority packets that are recently received can consume the granted bandwidth that was originally requested by the lower priority packets. It causes the delay of lower priority packets to increase in light load ONUs. The phenomenon is called light load penalty. A remedy for this problem that helps satisfy the QoS requirement for each traffic class is suggested in [33]. The suggestion is to move the contents of queues in order of their

Notation	Description
N_c	Number of classes
R(i)	Length of the queue i at the time of request
Q(i)	Queue Length at the beginning of the grant
G	Granted bandwidth
P	Predicted incoming traffic
$G_c(i)$	Granted bandwidth for class i
B_{rem}	Remaining unused bandwidth

Table 2.3: Notations used in Algorithm 2.

priority to a separate transmit buffer at the time of sending the request. This buffer will be transmitted at the next grant regardless of the higher priority packets that may arrive after that. After the transmit buffer is emptied, other queues are served in order of their priority. In other words, the transmit buffer acts as the highest priority queue. This algorithm is called two-stage DBA.

We propose a variant of two-stage DBA that does not require another buffer. At the time of sending the request, the ONU records the length of each queue. The recorded request is then used to determine how many packets can be sent from each queue during the next grant.

The ONU scheduler works in two rounds. In the first round, it serves the queues in order of their priority up to the saved request for each queue. Since the queues are First-In-First-Out (FIFO), it means that only the packets that were in the queue at the time of request (t_r) , are served. After that, the scheduler starts transmitting the next priority queue. The second round starts after all queues are swept once and all packets that were in the queues at the request time are transmitted. In this cycle the ONU scheduler performs normal strict priority scheduling to serve the newly arrived packets. The pseudo-code of the algorithm is shown in Algorithm 2.

The traffic-prediction enhanced DBA assigns bandwidth to incoming packets. But as previously described in Section 2.6, the class of the incoming traffic is not differentiated in order to keep the algorithm simple. To be able to use the bandwidth that is assigned to incoming packets, the accounted predicted traffic, represented by G_p , is also received along with the grant size. The scheduler adds G_p to the highest priority traffic request. After the incoming traffic at the highest priority is taken

Chapter 2: Improving the Quality of Service of Hybrid Optical Wireless Networks via a Novel Traffic-Aware Dynamic Bandwidth Assignment Algorithm

Algorithm 2 The intra-ONU scheduling algorithm pseudocode

```
Require: R(i), i = 1, \ldots, N_c
               Q(i), i = 1, ..., N_c
Require:
Require: G, P
  First round:
1: R(1) \leftarrow R(1) + G_p
    for i := 1 to N_c do
2:
       G_c(i) \leftarrow min(R(i), G - \sum_{k=1}^{i-1} G_c(k))
3:
4:
   end for
  Second round:
    for i := 1 to N_c do
5:
       B_{rem} \leftarrow \min(0, G - \sum_{k=1}^{N_c} G_c(k))

G_c(i) \leftarrow G_c(i) + \min(Q(i) - R(i), B_{rem})
6:
7:
    end for
8:
```

into account, the remaining of G_p is used to transmit the newly received packets of the next priority queue. This continues to lower priority queues. After the first round, the second round is performed without any modification. The pseudo-code of the algorithm is shown in Algorithm 2. The notations used are described in 2.3.

2.7 Maximum Throughput Analysis

The maximum throughput is achieved when the whole granted bandwidth is used. Therefore, in order to calculate the maximum throughput, it is assumed that all ONUs are backlogged, which has the desired result that the whole granted bandwidth would be used.

To calculate the maximum throughput, the wasted bandwidth has to be estimated. In each EPON cycle, which is denoted by C, the following factors waste the bandwidth. The required gap between ONU's granted slots, known as inter-ONU guard time G_1 , is a contributing factor to bandwidth waste. In each cycle, there are N gaps between the grants, where N is the number of active ONUs in the network. Another contributing factor is the gap between two consecutive cycles, which we represent as G_2 . The bandwidth that is used for sending the report messages also reduces user bandwidth. Again, there are N report messages in each EPON cycle. Adding up all of the waste factors, it is possible to work out the maximum throughput. It is assumed that the DBA algorithm that is used and the round trip time of ONUs, let the grant slots to be assigned without any gap in between other than the described ones. The maximum throughput is

$$R_{max} = 1 - \frac{N \times G_1 - G_2 - N \times \frac{L_{rep}}{R_E}}{C}, \qquad (2.12)$$

where R_{max} is the maximum throughput and L_{rep} is the length of the report message in bits.

Some DBA algorithms such as the one presented in [31], require all ONU requests to process the bandwidth assignment. In these algorithms, the idle time between frames, namely walk time, wastes a considerable amount of bandwidth. The walk time is basically composed of three different components; waiting time for the reception of the last ONU request, the processing time of scheduling algorithm and the round trip time of first ONU. The last factor comes from the fact that the link remains idle for the duration of transmitting the first ONU's grant and receiving data from it. Supposing that the processing time is negligible, it is equal to the first scheduled ONU's round trip time. In summary, walk time is defined as the time between reception of last report message and reception of the first data bit of the first ONU. In this period, the link is idle and the corresponding bandwidth is wasted. Taking the walk time into account the maximum throughput is worked out as

$$R_{max} = 1 - \frac{1}{C} \times \left(N \times G_1 - G_2 - N \times \frac{L_{rep}}{R_E} - T_{rtt} - T_{proc} \right)$$

where T_{rtt} is the round trip time of the ONU that its report is the last report and T_{proc} is the processing time of DBA algorithm.

2.8 Simulation Platform

The performance of the proposed algorithm is extensively studied through simulations under different scenarios. Simulation experiments are conducted with the OPnet simulation package for wireless internetworking V.15 [34]. Custom models for OLT,



Figure 2.7: Simulation Scenario.

and ONU-BS are developed to ensure utmost similarity to the real equipment.

WiMAX physical layer supports several technologies. This includes Orthogonal Frequency-Division Multiple Access (OFDMA), Orthogonal Frequency-Division Multiplexing (OFDM) and Single-Carrier Time Division Multiple Access (SC-TDMA); each supports Time Division Duplexing (TDD) or Frequency Division Duplexing (FDD) in the UL/DL channels. In this study, without lack of generality, we consider SC-TDMA with TDD.

The simulation scenario is depicted in Fig. 2.7. Unless mentioned otherwise, the scenario given in Fig. 2.7 is used as the simulation topology. This simulation model consists of one OLT which is connected to 32 ONU-BSs. The performance of the DBA algorithms is analyzed by simulating the entire system under different working conditions. The performance of the enhanced algorithms, developed in Section 2.6, as well as the original algorithms is studied.

The Ethernet and WiMAX workstations are responsible for generating different classes of traffic. Since we focus on the performance of the uplink direction, all of the traffic load is destined toward a server connected to the OLT. The default simulation

Simulation parameters	Value
EPON cycle length	2 ms
WiMAX PHY	TDD SC
WiMAX frame size	5 ms
InterONU guard time	$1 \ \mu s$
Propagation delay	$10 \ \mu s$

Table 2.4: Simulation parameters.

parameters are shown in Table 2.4.

To simulate the WiMAX part of the network, a modified version of standard OPnet WiMAX models is used. The modifications enable these models to share some of their internal information with the developed EPON models, which is vital to implement the proposed DBA algorithms for hybrid optical-wireless networks.

The EPON part of the network is custom designed to implement the point to multipoint aspects of the EPON as well as its Ethernet frame architecture. The EPON models consist of the link model, OLT MAC layer, and ONU MAC layer. The last two are illustrated in Fig. 2.8 and Fig. 2.9, respectively. These models are uploaded to the OPnet users' forum and accessible from [34].

Optical networks typically perform as the backbone infrastructure. In real implementations, they connect corporate networks as well as wireless base stations to the operator's core. This realistic scenario has been studied in Section 2.9.4.

In the realistic scenario, half of the ONU-BSs have been replaced by the legacy ONUs, that serve the corporate users, to study their effect on the performance of the proposed algorithm. The legacy ONUs cannot predict the traffic and have to request bandwidth with the traditional method.

In addition, in a real network the real-time load of the users are not equal. To model this effect, in the realistic scenario, 80% of the total traffic is generated by top 25% high load ONUs and ONU-BSs. The remaining load is then generated by low load ONUs and ONU-BSs. The traffic load of the high load and low load ONUs and ONU-BSs are shown in Table 2.5.



Chapter 2: Improving the Quality of Service of Hybrid Optical Wireless Networks via a Novel Traffic-Aware Dynamic Bandwidth Assignment Algorithm

Figure 2.8: OLT MAC layer process model in OPNet.

2.8.1 Traffic generation and types

Three traffic types are considered in the simulation experiments. The highest priority traffic is the Expedited Forwarding (EF) class which models the voice or video traffic. The quality of service requirements of this class are defined by the maximum tolerable delay, maximum jitter, maximum loss and throughput. The second priority traffic is the Assured Forwarding (AF). Multimedia streaming traffic falls into this category. This class of traffic is delay-constrained and typically its delivery is assured as long

	Low Load (Mbps)	High Load (Mbps)
ONU	12.5	150
ONU-BS	4.5	50

Table 2.5: Traffic loads in realistic scenario.



Chapter 2: Improving the Quality of Service of Hybrid Optical Wireless Networks via a Novel Traffic-Aware Dynamic Bandwidth Assignment Algorithm

35

Figure 2.9: ONU MAC layer process model in OPNet.

as its throughput remains within predefined limits. The third, and most basic traffic class is the Best Effort (BE) traffic that models file transfers, email and web traffic. This class has no specific requirement in terms of maximum delay or jitter. However, the throughput plays an important role for this class, and hence, some researchers define a minimum throughput requirement to prohibit BE traffic from being cut-off from the network in highly loaded conditions.

WiMAX subscriber stations (Fig. 2.7) are responsible for generating traffic with the required distribution. The EF traffic load is generated by a Poisson process while AF and BE traffic classes are generated by self-similar traffic generators to better model real traffic in the network. Moreover, to model the traffic more realistically, EF traffic is composed of packets with fixed length equal to 100 bytes, while packet length in AF and BE classes is uniformly distributed between 100 and 1500 bytes. The total traffic is composed of BE, AF and EF; each of them shares 0.4, 0.4 and 0.2 of the total traffic, respectively [35]. Moreover, it should be noted that unless stated otherwise, the network load is sufficient enough to saturate the network. The measured throughput is normalized to nominal capacity of the link which is 1 Gbps in EPON.

A simple approach to generate self-similar traffic, is to multiplex several traffic flows with a distribution that has long-range dependence (LRD) property, like Pareto distribution [36]. The density function of Pareto distribution is,

$$f(x) = \frac{\alpha}{k} \left(\frac{k}{x}\right)^{\alpha+1}; \quad x > k, \alpha > 0$$
(2.13)

Since a traffic with heavy-tailed distribution is desired, the shape parameter is bounded $(1 < \alpha < 2)$. The degree of self-similarity which is normally expressed as Hurst parameter (H) is related to shape parameter by,

$$H = \frac{3 - \alpha}{2}, \quad 1 < \alpha < 2$$
 (2.14)

The model used in simulation experiments, is a superposition of Fractal Point Processes (Sup-FRP). The Hurst parameter is set to 0.8 which is suitable for web traffic as well as file transfer [37]. The fractal onset time scale is set to 0.1 second.

The simulation scenario consists of three users per each ONU-BS. Each user is responsible of generating one class of traffic. This gives us the flexibility of modifying each traffic class and its corresponding parameters without affecting other classes of traffic.

2.9 Simulation Results

The results of simulating the algorithms in different loads, and several cycle and frame length are given in this section. In addition, the performance of the algorithms in realistic conditions are reported in Section 2.9.4.



Figure 2.10: Normalized throughput versus normalized load.

2.9.1 Load analysis

In the following paragraphs, the behavior of the proposed algorithms under different loads is investigated. Under light and moderate loads, the network throughput, shown in Fig. 2.10, is the same for all algorithms. This is an obvious result of introducing less load than network capacity which all of the DBA algorithms are capable of delivering. The throughput curves separate when the introduced load approaches the link capacity.

Fig. 2.11 compares the delay of the EF flow for different algorithms as the load increases. As expected, an increase in load results in more delay. Under lower loads, the gradual increase in the delay of CB-IPACT is mainly because of the increase in the cycle length. The cycle length of CB-IPACT extends with load which in turn results in approximately 1 ms increase in the average delay. The probability density function (PDF) of delay of EF flows (Fig. 2.13) clearly shows that increasing the load results in longer delays. In full load conditions, cycle length saturates to its maximum and the EF delay is increased by more than the maximum cycle length. The same



Figure 2.11: Average delay of EF flow in different loads.

discussion applies for AF flows with the minor difference that AF flows experience a faster increment due to their lower priority.

The overheads associated with the integration cause a small increase in the average delay in the low and moderate loads in integrated versions of the algorithms. Specifically, the integrated-ED suffers from this effect. However, the reward of bearing a slightly longer delay in light loads is a significant delay decrease in high load, as shown in Fig. 2.11 and Fig. 2.12.

Both Excess Distribution (ED) and integrated-ED algorithms fix the cycle length equal to the maximum cycle length of CB-IPACT and integrated-CBIPACT. As a result, their cycle length is longer than the average cycle length of CB-IPACT and integrated-CBIPACT in light and moderate loads. Therefore, their delay is generally larger than that of CB-IPACT and integrated-CBIPACT. ED algorithm can keep constant delay when load is up to 75% of link capacity for EF traffic and up to 50% of link capacity for AF traffic. This difference is due to the fact that higher priority traffic can take advantage of the excess bandwidth which is distributed between



Figure 2.12: Average delay of AF flow in different loads.

ONUs, to send their packets. Again, since the AF class has lower priority, it is more susceptible to increasing load and the adverse effects can be seen sooner.

Without traffic prediction, the delay of both flows increases significantly at high load. However, if the proposed traffic prediction algorithm is applied, it counteracts the load effect and maintains the delay at approximately the same level. The effect of prediction can be best demonstrated with the PDF of the delay that is shown in Fig. 2.14.

Fig. 2.11 shows an interesting phenomenon. The average delay of the EF flow decreases as load increases from 0.8 to 1. The reason is that in moderate loads, there is an amount of excess bandwidth that is distributed among the ONUs. This excess bandwidth varies significantly for each cycle and as a result, the granted bandwidth also significantly varies. Therefore, the position of the granted slot in the cycle changes, which results in a variation in the measured delay from one grant time slot to the next. This might increase the cycle length that is measured in some ONUs and as a result, increase their average delay. Not all of the ONUs are affected by the

Chapter 2: Improving the Quality of Service of Hybrid Optical Wireless Networks via a Novel Traffic-Aware Dynamic Bandwidth Assignment Algorithm



Figure 2.13: PDF of EF delay using CB-IPACT and integrated-CBIPACT.

phenomenon. Since all algorithms serve ONUs in a round robin fashion, the variation in grants does not affect first and last ONUs. However, the full effect of variation can be seen at the ONUs that are granted in the middle of the cycle. When the load increases to full load, less excess bandwidth remains and the granted slots become equal. Therefore, the granted slot does not change its position and the phenomenon vanishes, which is shown in Fig. 2.11 as a slight decrease in average EF delay.

In summary, from QoS point of view, the throughput is almost the same for all four algorithms. The delay bound of the algorithms is the same in light loads. But in heavy loads, the integrated algorithms, due to their unique prediction method, allow considerably lower delay bound, at the expense of an insignificant delay increase in light loads. Moreover, a comparison of fixed and adaptive cycle length algorithms reveals that although both categories have the same delay bound, the adaptive cycle length algorithms perform better in terms of delay, in light load conditions. This holds true for the integrated algorithms.

Chapter 2: Improving the Quality of Service of Hybrid Optical Wireless Networks via a Novel Traffic-Aware Dynamic Bandwidth Assignment Algorithm



Figure 2.14: PDF of EF delay using ED and integrated ED.

2.9.2 EPON cycle length analysis

Since the ability to predict incoming traffic depends on the ratio between the lengths of the EPON cycle and WiMAX frame, as discussed in Section 2.5, changing the cycle length has a significant impact on the performance of the proposed algorithm. In saturated conditions, where all of the ONUs are backlogged, almost all DBAs assign the available bandwidth equally to the ONUs. Therefore, the difference in performance is mainly due to the traffic prediction. Hence, by analyzing the full load behavior of the algorithms, it is possible to investigate the prediction performance in nearly isolated environment.

Fig. 2.15 compares the network throughput in different cycle lengths. As can be seen from the figure, the normalized throughput varies between 0.92 and 0.99. The difference is due to the fact that CB-IPACT assigns the bandwidth on the fly. *i.e.*, as the scheduler receives each bandwidth request, it processes the request and sends the grant. On the other hand, ED processes all requests at the same time. Hence, walk time causes a reduction in ED's maximum throughput in comparison



Figure 2.15: Normalized throughput versus cycle length.

with CB-IPACT. Static bandwidth assignment does not perform the assigning based on requests and its maximum throughput is the same as CB-IPACT.

Fig. 2.15 also shows that the maximum throughput increases with longer cycle lengths. This is mainly because there is less frequent request/grant exchanges which directly translates in reduced walk time. Therefore, the bandwidth consumed for the request/grant pairs and also the bandwidth wasted in walk times can be conserved. Besides, comparing the integrated versions of the algorithms in with the original ones, reveals the throughput penalty caused by the extra message exchange to share the traffic information with the OLT in integrated versions.

The simulated throughput as well as the analytical throughput are shown in Fig. 2.16. The minor difference between simulation and analytic throughput is caused by the fragmentation effect and the added message exchanged, which we did not consider in analysis. Since fragmentation of packets between cycles is not allowed in EPON, the assigned bandwidth is wasted if the packet is larger than the remaining granted bandwidth. This slightly decreases the throughput in simulation experiments. For



Figure 2.16: Normalized analytic throughput versus cycle length.

shorter EPON cycles, this can happen more frequently, and thus the effect is larger. The message exchange overhead also increases in shorter cycles.

Fig. 2.17 shows the average delay of the EF stream with different cycle length. Generally, by increasing the cycle length, the average delay also increases, because there are less frequent grant slots available to the ONU. A packet may remain in queue longer in longer cycles. This justifies the linear increase in the delay shown in Fig. 2.17. By knowing the incoming traffic in advance, the proposed algorithms can save the time that is normally spent on exchanging request/grant. As expected, this results in almost a cycle length reduction in delay when full knowledge of incoming traffic exists. For example, assuming the frame length is 5 ms, the ONU has full knowledge of incoming traffic when the cycle length is equal to 1 ms or 2 ms (refer to Section 2.5 for details). It is backed by 1 ms and 2 ms delay reduction when cycle length is set to 1 ms, and 2 ms, respectively.

By contrast, given that the WiMAX frame length is 5 ms, in 5 ms and 10 ms EPON cycle lengths, ONU has only partial knowledge on incoming traffic, and hence



Figure 2.17: Average delay of EF stream versus EPON cycle length.

a part of incoming traffic is handled in conventional manner, *i.e.*, request/grant mechanism. Therefore, it results in only a partial reduction in delay. This phenomenon can be best understood through careful investigation of the PDF of the EF delay, shown in Fig. 2.18. When EPON cycle length is set to 2 ms, adding prediction to the algorithm clearly shifts the PDF leftward and decreases the delay. However, when EPON cycle length is 5 ms, it is not always possible to send the required information to the prediction module in time and information may be received when it is expired. This is demonstrated in Fig. 2.19. The required information for the prediction of incoming traffic which is extracted from the DL/UL map is transmitted to the OLT scheduler along with the request. In the case of BS1, the information is updated after transmission of request. Therefore, the incoming packet has to wait for the next request to be sent if there is no extra granted bandwidth available for sending it. On the other hand, BS2 updates the information before sending request and hence the incoming packet is predicted in the scheduler and the required bandwidth is granted. This results in the PDF shown in Fig. 2.18.

Chapter 2: Improving the Quality of Service of Hybrid Optical Wireless Networks via a Novel Traffic-Aware Dynamic Bandwidth Assignment Algorithm



Figure 2.18: Comparison of the PDF of EF delay for different EPON cycle lengths.

The same effect can be seen in Fig. 2.20 for the AF streams. However, because the prediction module does not differentiate between traffic types, the bandwidth assigned to the predicted traffic is assigned to the highest priority flow, which is the EF flow. This does not affect the performance when ONU has full knowledge of the coming traffic. But in case of partial knowledge, *i.e.*, when EPON cycle length is 5 ms or 10 ms, the unpredicted EF packets can take advantage of the bandwidth, originally assigned to AF packets. This deteriorates the performance of lower priority packets when only partial knowledge is available.

In summary, throughput is not affected in the integrated algorithms when cycle length changes. On the other hand, the proposed algorithms, can decrease the delay of the high priority classes significantly.

2.9.3 WiMAX frame length analysis

Changing the length of the WiMAX frame affects the performance metrics of the proposed algorithms. Similar to the previous case, the only performance metric which





Figure 2.19: Timing diagram for partial knowledge of incoming traffic.

is not affected is the system throughput which is shown in Fig. 2.21. The EPON cycle length is fixed to 2 ms in all simulation experiments. Like the previous discussion, the ED's throughput is slightly less than CB-IPACT due to walk time.

The delay of EF and AF streams for different frame lengths is shown in Fig. 2.22 and Fig. 2.23, respectively. Although, changing the frame length does not affect the amount of traffic load, it changes the frequency and shape of the incoming traffic; *i.e.*, when the frame length is extended, the frequency of incoming traffic decreases, and at the same time, the amount of traffic delivered in each frame to optical network increases.

The delay of the EF stream remains fixed for WiMAX frame lengths up to 5 ms, and then, it slightly increases, while the delay of AF streams is only fixed for WiMAX frame lengths up to 2 ms, and then, it increases dramatically afterward. Because WiMAX TDD is used as the physical layer of the wireless part of the network, the incoming traffic is bursty in nature. This does not affect the delay, if WiMAX frame length and EPON cycle length are set in a way that EPON sees continuous traffic, *i.e.*, the amount of traffic introduced to EPON is almost the same in each EPON cycle. But as WiMAX frame length grows to more than that of the EPON cycle, the incoming traffic becomes bursty and pulse-shaped. The amount of the traffic in each pulse also increases. In long frames, the amount of traffic delivered to the ONU is more than the maximum grant size. Thus, it is divided among several EPON cycles, resulting in increased queuing delay. Since the prediction algorithm does not provide the information about the amount of traffic in each class, the biggest share of the predicted bandwidth is allocated to the EF stream, and hence, it exhibits



Figure 2.20: Average AF delay versus EPON cycle length.

the increasing trend later than AF stream does. The phenomenon can be clearly seen in Fig. 2.22 and Fig. 2.23.

Extending the frame length leads to extending the range of the prediction, and in turn, decreases the mean EF delay. However, the decreasing trend stops at frame length equal to 5 ms, and increases thereafter. As the WiMAX frame is extended, the amount of traffic aggregated in each frame grows. The aggregated traffic is delivered almost instantaneously to the ONU for transmission. In the short frames, the amount of traffic is little enough to be sent in a single EPON cycle. Again, the lack of distinction between the traffic classes in prediction acts in favor of high priority traffic, since EF traffic can also use the granted bandwidth that is originally assigned to AF flows. When the aggregated EF traffic is increased to more than the granted bandwidth, it spreads over two or more EPON cycles, which results in increased delay.

In summary, increasing the WiMAX frame length in comparison to EPON cycle length helps the prediction algorithm predict traffic for a longer range. On the other



Figure 2.21: Maximum throughput versus frame length.

hand aggregating traffic, caused by longer WiMAX frames, can result in the loss of statistical multiplexing gain. Long WiMAX frame lengths also increases the queuing delay which contracts the delay reduction gained by traffic prediction. Therefore, it is vital to intelligently choose the EPON cycle length in relation to the WiMAX frame length to take advantage of the prediction.

2.9.4 A realistic scenario

The delay of EF and AF traffic classes are shown in Fig. 2.24. It shows the delay of high and low load ONU and ONU-BSs when various DBA algorithm are employed. As seen in the figure, applying the proposed prediction-based DBA algorithm does not affect the delay of legacy ONUs. However, it lowers the delay of ONU-BSs' delay-sensitive packets dramatically, regardless of their traffic load. This is due the fact that in cases where the whole network is working with full capacity, the total requested bandwidth is greater than the total available bandwidth. Thus, the ONUs and ONU-BSs are granted with the requested bandwidth or less which results in an extra delay



Figure 2.22: Average EF delay versus frame length.

for requesting bandwidth. Since the proposed scheme pre-grant the packets from ONU-BSs, the algorithm benefits both of the high and low load ONU-BSs.

Fig. 2.25 shows the throughput of ONU/ONU-BSs. In this figure, high load ONU and ONU-BS are represented by ONU_H and ONUBS_H, whereas ONU_L and ONUBS_L represent low load ONUs and ONU-BSs, respectively. It is concluded from the figure that the delay decrement in delay-sensitive classes of traffic comes with a slight decrease of BE throughput in high load ONUs. On the other hand, the high load ONU-BSs are granted more bandwidth to enable them to better serve the delay-sensitive classes. It is worth mentioning that the throughput of low load ONUs and ONU-BSs are not affected. Also, applying the proposed algorithm does not harm the EF and AF throughput of high load ONUs.



Figure 2.23: Average AF delay versus frame length.

2.10 Chapter Summary

Hybrid optical-wireless network is a promising technology that is relatively inexpensive and can address the huge bandwidth need of modern applications. In this work, a mechanism was proposed to enhance the performance of the optical scheduler algorithm by augmenting it with incoming traffic prediction, which is extracted from the internal information of the wireless scheduler. The prediction method was described and investigated in detail. Based on the prediction method, two DBA algorithms were proposed. Extensive simulation experiments proved the performance of the algorithms and showed that due to the employed prediction method, the delay of both algorithms in high loads can be decreased by a factor of two, without affecting the throughput. The effects of changing the main parameters were also investigated. In addition to that, the performance of the proposed algorithm was proved in a real world scenario which consists of ONUs and ONU-BSs.

The results showed that predicting the traffic removes the significant delay increase which normally occurs in high load conditions. In fact, the delay of high

51



Figure 2.24: The delay of EF and AF traffic classes in the realistic scenario.

priority traffic remains the same for high and light load conditions. This enables the service providers to establish a lower delay bound and a better quality of service for consumers, even in high load conditions.

Since the proposed algorithm is superior to conventional algorithms, only in the high load conditions, and also because the main part of the algorithms runs in the OLT, it is very simple to further modify the algorithm to activate only in high loads, where it is useful the most, and to remain idle otherwise. This results in a simpler DBA in light load conditions and a more efficient one in high load conditions. As a future work, it is useful to establish a proper load threshold to activate the proposed DBA algorithm.



(sqdM) tudhgud (Mbps)

0

9

100 -

Throughput (Mbps)

150 J

50

0

50



References

- M. Mirahmadi and A. Shami, "Traffic-prediction-assisted dynamic bandwidth assignment for hybrid optical wireless networks," *Computer Networks*, vol. 56, no. 1, pp. 244–259, 2012.
- [2] —, "Traffic-prediction-assisted dynamic bandwidth assignment for hybrid wireless optical networks," in *Int. Wireless Commun. and Mobile Computing Conf. (IWCMC)*, Istanbul, Turkey, Jul. 2011, pp. 2103–2108.
- [3] G. Shen, R. Tucker, and C.-J. Chae, "Fixed mobile convergence architectures for broadband access: Integration of EPON and WiMAX," *IEEE Commun. Mag.*, vol. 45, no. 8, pp. 44–50, Aug. 2007.
- [4] Y. Yan, H. Yu, H. Wang, and L. Dittmann, "Integration of EPON and WiMAX networks: uplink scheduler design," in *Proc. of SPIE*, vol. 7137, no. 1, 2008, p. 713704.
- [5] S. Ou, K. Yang, M. Farrera, C. Okonkwo, and K. Guild, "A control bridge to automate the convergence of passive optical networks and IEEE 802.16 (WiMAX) wireless networks," in 5th Int. Conf. Broadband Commun., Networks and Syst. (BROADNETS), Sep. 2008, pp. 514–521.
- [6] F. Alsolami and P. Ho, "A novel two-level hierarchical scheduling scheme for integrated WiMAX and EPON," *IEEE Commun.*, Surveys and Tutorials, pp. 46–60, 2009.
- [7] K. Yang, S. Ou, K. Guild, and H.-H. Chen, "Convergence of Ethernet PON and IEEE 802.16 broadband access networks and its QoS-aware dynamic bandwidth allocation scheme," *IEEE J. Sel. Areas Commun.*, vol. 27, no. 2, pp. 101–116, Feb. 2009.

- [8] Y. Yan, H. Yu, and L. Dittmann, "Wireless channel condition aware scheduling algorithm for hybrid optical/wireless networks," in 3rd Int. Conf. Access Networks (Access Nets). Las Vegas, NV: Springer, Oct. 2009, pp. 397–409.
- [9] A. Reaz, V. Ramamurthi, S. Sarkar, D. Ghosal, S. Dixit, and B. Mukherjee, "CaDAR: An efficient routing algorithm for a wirelessoptical broadband access network (WOBAN)," *J. Optical Commun. and Networking*, vol. 1, no. 5, pp. 392–403, Oct. 2009.
- [10] I.-S. Hwang, J.-Y. Lee, C.-W. Huang, and Z.-D. Shyu, "Advanced dynamic bandwidth allocation and scheduling scheme for the integrated architecture of EPON and WiMAX," in 10th Int. Conf. Mobile Data Management (MDM), vol. 0, Los Alamitos, CA, 2009, pp. 655–660.
- [11] M. El-Gendy, A. Bose, and K. Shin, "Evolution of the Internet QoS and support for soft real-time applications," *Proc. IEEE*, vol. 91, no. 7, pp. 1086–1104, Jul. 2003.
- [12] IEEE Standard for Local and metropolitan area networks Part 16: Air Interface for Broadband Wireless Access Systems, IEEE Std. 802.16m, 29 2009.
- [13] B. O. Obele, M. Iftikhar, S. Manipornsut, and M. Kang, "Analysis of the behavior of self-similar traffic in a QoS-aware architecture for integrating WiMAX and GEPON," J. Optical Commun. and Networking, vol. 1, no. 4, pp. 259–273, Sep. 2009.
- [14] K. Katrinis, A. Tzanakaki, S. Dweikat, S. Vassilaras, R. Nejabati, D. Simeonidou, and G. Zervas, "Backhauling wireless broadband traffic over an optical aggregation network: WiMAX over OBS," in 6th Int. Conf. Broadband Commun., Networks, and Syst. (BROADNETS), Sep. 2009, pp. 1–4.
- [15] B. O. Obele and M. Kang, "Fixed mobile convergence: A self-aware QoS architecture for converging WiMAX and GEPON access networks," 2nd Int. Conf. Next Generation Mobile Applications, Services and Technologies (NGMAST), vol. 0, pp. 411–418, Sep. 2008.

- [16] S. Wang, H. Li, and Y. Ji, "Distribution QoS scheme for a novel of hybrid optical wireless network," in *Proc. of SPIE*, vol. 6784, Oct. 2007.
- [17] S. Sarkar, B. Mukherjee, and S. Dixit, "Optimum placement of multiple optical network units (ONUs) in optical-wireless hybrid access networks," in *Optical Fiber Communication Conf. (OFC)*, Anaheim, CA, Mar. 2006.
- [18] S. Sarkar, H.-H. Yen, S. Dixit, and B. Mukherjee, "DARA: Delay-aware routing algorithm in a hybrid wireless-optical broadband access network (WOBAN)," in *IEEE Int. Conf. Commun. (ICC)*, Glasgow, Scotland, Jun. 2007, pp. 2480–2484.
- [19] —, "Radar: Risk-and-delay aware routing algorithm in a hybrid wirelessoptical broadband access network (WOBAN)," in *Optical Fiber Communication Conf. (OFC)*, Anaheim, CA, Mar. 2007, pp. 1–3.
- [20] —, "A mixed integer programming model for optimum placement of base stations and optical network units in a hybrid wireless-optical broadband access network (WOBAN)," in *IEEE Wireless Commun. and Networking Conf.* (WCNC), Hong Kong, Mar. 2007, pp. 3907–3911.
- [21] S. Sarkar, S. Dixit, and B. Mukherjee, "Hybrid wireless-optical broadband-access network (WOBAN): A review of relevant challenges," J. Lightwave Technology, vol. 25, no. 11, pp. 3329–3340, Nov. 2007.
- [22] S. Sarkar, H.-H. Yen, S. Dixit, and B. Mukherjee, "A novel delay-aware routing algorithm (DARA) for a hybrid wireless-optical broadband access network (WOBAN)," *IEEE Netw.*, vol. 22, no. 3, pp. 20–28, May/Jun. 2008.
- [23] —, "Hybrid wireless-optical broadband access network (WOBAN): Network planning and setup," *IEEE J. Sel. Areas Commun.*, vol. 26, no. 6, pp. 12–21, Aug. 2008.
- [24] S. Sarkar, Y. Hong-Hsu, S. Dixit, and B. Mukherjee, "Hybrid wireless-optical broadband access network (WOBAN): Network planning using lagrangean relaxation," *IEEE/ACM Trans. Netw.*, vol. 17, no. 4, pp. 1094–1105, Aug. 2009.

- [25] W.-T. Shaw, S.-W. Wong, N. Cheng, and L. Kazovsky, "MARIN hybrid opticalwireless access network," in *Optical Fiber Communication Conf. (OFC)*, Anaheim, CA, Mar. 2007, pp. 1–3.
- [26] Y. Luo, T. Wang, S. Weinstein, M. Cvijetic, and S. Nakamura, "Integrating optical and wireless services in the access network," in *Nat. Fiber Optic Engineers Conf. (NFOEC)*, Anaheim, CA, 2006.
- [27] Y. Luo, S. Yin, T. Wang, Y. Suemura, S. Nakamura, N. Ansari, and M. Cvijetic, "QoS-aware scheduling over hybrid optical wireless networks," in *Nat. Fiber Optic Engineers Conf. (NFOEC)*, 2007.
- [28] C. So-In, R. Jain, and A.-K. Tamimi, "Scheduling in IEEE 802.16e mobile WiMAX networks: key issues and a survey," *IEEE J. Sel. Areas Commun.*, vol. 27, no. 2, pp. 156 –171, Feb. 2009.
- [29] F. Ohrtman, WiMAX handbook: building 802.16 wireless networks, ser. McGraw-Hill communications. McGraw-Hill, 2005.
- [30] G. Kramer, B. Mukherjee, and G. Pesavento, "IPACT: A dynamic protocol for an Ethernet PON (EPON)," *IEEE Commun. Mag.*, vol. 40, no. 2, pp. 74–80, Feb. 2002.
- [31] C. Assi, Y. Ye, S. Dixit, and M. Ali, "Dynamic bandwidth allocation for qualityof-service over Ethernet PONs," *IEEE J. Sel. Areas Commun.*, vol. 21, no. 9, pp. 1467 – 1477, Nov. 2003.
- [32] A. Dhaini, C. Assi, and A. Shami, "Dynamic bandwidth allocation schemes in hybrid TDM/WDM passive optical networks," in 3rd IEEE Consumer Commun. and Networking Conf. (CCNC), vol. 1, Las Vegas, NV, Jan. 2006, pp. 30 – 34.
- [33] G. Kramer, B. Mukherjee, S. Dixit, Y. Ye, and R. Hirth, "Supporting differentiated classes of service in Ethernet passive optical networks," J. of Optical Networking, vol. 1, no. 8, pp. 280–298, 2002.
- [34] OPnet simulation software. [Online]. Available: www.opnet.com

- [35] A. Shami, X. Bai, N. Ghani, C. Assi, and H. Mouftah, "QoS control schemes for two-stage ethernet passive optical access networks," *IEEE J. Sel. Areas Commun.*, vol. 23, no. 8, pp. 1467 – 1478, Aug. 2005.
- [36] L. Ying, K. Feng-ju, Z. Lian-jiong, and W. Zhi-Guang, "A self-similar traffic generation method and application in the simulation of mobile ad-hoc network," in *ISECS Int. Colloq. Computing, Communication, Control, and Management* (CCCM), vol. 4, Sanya, China, 8-9 2009, pp. 229 –233.
- [37] J. Potemans, B. V. den Broeck, Y. Guan, J. Theunis, E. V. Lil, and A. V. de Capelle, "Implementation of an advanced traffic model in opnet modeler," in *OPNETWORK 2003*, Washington D.C., 2003.
Part II

Managing Femtocell Interference in Preplanned Cellular Network

Chapter 3 Predicting Femtocell Interference for Next-Generation Networks via a Novel Building Architecture Model

3.1 Preamble

One of the main goals of the fourth generation (4G) wireless networks is to offer high data rates to all users anywhere within the coverage area. However, delivering reliable services to users inside buildings remains a major challenge due to the significant attenuation of the buildings' walls. An effective solution that has been recently adopted by several standards is to create a hierarchical cell structure where a macrocell is overlaid over a number of femtocells, which are devoted to provide coverage for indoor users. In such configuration, the end user will be connected to the femto base station (FBS) via a wireless link, but the FBS will be connected to service providers' core network through a broadband connection such as the digital subscriber line (xDSL) or fiber to the x (FTTx). The FBS is known as "femto BS" in WiMAX or "Home NodeB (HNB)" in 3GPP terminology.

Depending on the spectrum availability, femtocells can be configured to function in dedicated channels, or share the spectrum with other existing networks [2]. In the latter case, deploying femtocells within the coverage area of a macrocell may deteriorate the performance of the macrocell users due to the mutual interference between the two systems. Although the FBS maximum transmit power can be adjusted

A version of this chapter has been published in [1]. [©2012 IEEE]

Chapter 3: Predicting Femtocell Interference for Next-Generation Networks via a Novel Building Architecture Model 60

dynamically according to the macro user interference level, such process is not trivial and includes several trade-offs to consider. To achieve the desirable result, a receiver inside the building should get a high signal to interference and noise ratio (SINR) anywhere inside the building while the leakage of the FBS signal to the outside world should be kept to minimum. Achieving these two conflicting goals simultaneously is very challenging without considering the signal attenuation due to the propagation through walls, doors, windows and other building materials. Moreover, because of the complexity of the mixed indoor-outdoor environment, analyzing the interference in such environments using ray tracing [3] or finite-difference time-domain (FDTD) methods [4] would be highly complex, time-consuming, and it will be valid only for the case in study. Furthermore, it requires detailed information about the environment, such as the building's floor plan, which is typically not accessible. Therefore, a simple model that captures the fundamental properties of such scenarios is indispensable.

In other words, buildings act as a shield that reduces the mutual interference between the macro and femto users. However, wall attenuation changes considerably in a broad range from 5 dB to 20 dB or more, based on the wall properties such as the type of material used and thickness. Furthermore, a signal transmitted through windows or doors is attenuated less than 3 dB.

In the literature, most of the works focused on the outdoor-to-indoor propagation models, aimed at extending the coverage of an outdoor transmitter to indoor receivers, and in fact a few papers considered the indoor-to-outdoor case. A comprehensive study of the outdoor-indoor interface that includes the effects of different materials is reported in [5]. The final report of the European Co-Operation in the field of Scientific and Technical research, Action 231 (COST-231) [6] also includes a model for penetration of the signal inside buildings. The model which takes several factors such as frequency and distance into account, is calibrated empirically. Oestges *et al.* [7] proposed a similar model with the difference that the effect of the angle of incident was included. Several measurements were used to calibrate the model at 2.5 GHz. Another model which considers the distance and carrier frequency is explained in ITU-R M.1225 [8]. The models in [9] calculate the indoor path loss as a linear function of indoor distance. Extensive research was conducted to deterministically model the propagation in mixed outdoor-indoor scenarios using ray-tracing or other methods [4, 10–12]. Such methods require a large amount of details including the topography and digitized 3D map of the city and their results are site-specific. In [13], the authors modeled the transmission of a signal from the rooms that have external walls or the rooms adjacent to them. The measurements have been taken in several frequencies from 0.9-3.5 GHz. The proposed frequency-dependent model also considers the number of walls between the transmitter and outside. Although this work is unique, it does not model the transmitters farther from the external walls which limits its applicability.

This chapter presents a statistical model for indoor-to-outdoor path loss where the transmitter is a FBS inside a suburban house. In particular, the randomness due to building floor plans and FBS's position is modeled. The proposed model is then used to analyze the femtocell signal interference on macro users. The chapter also investigates the issue of FBS's placement to minimize the interference on macrocell users. The rest of the chapter is as follows. Section 3.2 describes the signal propagation into buildings. The proposed propagation model is discussed in Section 3.3. Note that the proposed model is based on a procedural building generation (PBG) algorithm, developed in Chapter 4. A very concise description of the algorithm is presented in this chapter for the sake of completeness. The algorithm is presented in details in the next chapter. Section 3.4 proposes the stochastic femtocell signal model. In Section 3.5, as an example of the propoed model's applications, it is used to analyze the effects of FBS's position on its interference on a macrocell user outside the building, and finally, Section 3.6 concludes the chapter and gives some ideas for further studies.

3.2 Signal Propagation into Buildings

Radio frequency signal propagation inside buildings is a combination of several complex physical effects. In practice, empirical models are used to calculate the signal strength. A successful mixed indoor-outdoor propagation model should consider the main characteristics that distinguish the channel from typical urban wireless channels.

First, it should estimate the signal attenuation due to passing through walls, doors, and windows. Second, regardless of walls' attenuation, the indoor channel should be modeled differently from the outdoor channel. Third, a method of transition between the models should be implemented.

In general, the average received signal power $P_{\rm r}$ at a distance $d_{\rm x}$ from a transmitter located inside a particular building can be written as [14]

$$P_{\rm r} = P_{\rm t} - L_{\rm T} \quad (\rm dB) \tag{3.1}$$

where $P_{\rm t}$ is the transmitted signal power and $L_{\rm T}$ is the total attenuation. Modeling the mixed indoor-outdoor channel is beyond the scope of this work. Hence, we rely on widely-accepted channel models, such as COST-231 MultiWall Model (MWM) [6], which is a semi-empirical model that has been augmented with some deterministic predictions, namely, the attenuation of walls. Most 3GPP recommendations used this model [2].

In this model, the signal attenuation can be computed as,

$$L_{\rm T} = 20 \log \left(\frac{4\pi d_{\rm x}}{\lambda_{\rm c}}\right) + k_{\rm f}^{\left(\frac{k_{\rm f}+2}{k_{\rm f}+1}-0.46\right)} \alpha_{\rm f} + \sum_{i=1}^{W} k_i \alpha_i \quad ({\rm dB})$$
(3.2)

where d_x is the distance between the transmitter and receiver, λ_c denotes the carrier wavelength, $k_{\rm f}$ is the number of penetrated floors, $\alpha_{\rm f}$ is the loss between adjacent floors, W is the number of considered wall types, α_i and k_i are the attenuation and number of penetrated walls of type i, respectively. The attenuation factors α_i and α_f depend on the dimension of the walls and floors, as well as their constituting materials. The attenuation of typical walls at several frequencies is given in [7]. Femtocells are typically used to cover residential buildings where the FBS is installed at the street level. In such scenarios, the effect of the floors in (3.2) can be omitted, which leads to a two dimensional (2D) propagation model, depicted in Fig. 3.1. Therefore, (3.2)



Figure 3.1: Cost MultiWall Model (MWM) concept and a sample.

is reduced to

$$L_{\rm T} = \underbrace{20 \log \left(\frac{4\pi d_{\rm x}}{\lambda_{\rm c}}\right)}_{L_{\rm FS}} + \underbrace{\sum_{i=1}^{W} k_i \alpha_i}_{L_{\rm B}} \quad (dB).$$
(3.3)

The free space path loss $L_{\rm FS}$ depends on the distance between the transmitter and receiver, and the building loss $L_{\rm B}$ depends on the walls, doors, windows and other obstacles in the building. The signal attenuation incurred due to passing through walls are added to the basic transmission loss, i.e., the free space path loss [15]. Obviously, the attenuation of walls depends on their dimension as well as the constituting materials. Table 3.1 presents some approximate values of the attenuation of walls with typical dimensions at different frequencies. It is worth mentioning that the COST-231 MWM considers only the walls that are traversed on a straight line connecting the transmitter and the receiver. Therefore, the attenuation factors α_i are not directly used and subsequently tuned empirically to account for reflections.

It should be mentioned that there are mainly two approaches to model the signal attenuation in indoor environments. The first approach is based on employing the logarithmic distance in the COST-231 MWM, and the other approach based on representing the attenuation as a multiplication of distance $(L_{\rm T} = a \times d_{\rm x})$, where a is an empirical coefficient. Although some measurement results show that the COST-231 approach describes the walls attenuation more accurately than other models [15], some guidelines chose to model the indoor attenuation as a linear function of distance.

Chapter 3: Predicting Femtocell Interference for Next-Generation Networks via a Novel Building Architecture Model

Wall Material	Attenuation (dB) in 900 MHz	Attenuation (dB) in 1800 MHz	Attenuation (dB) in 2.5 GHz	
Wooden walls	8	10	12.3	
Concrete walls	14-24	16-26		
Stucco walls			13.1	

Table 3.1: Attenuation of walls in different frequencies [7]

The reasoning for this selection is that the COST-231, despite its superiority, requires typically inaccessible and detailed information about the floor plan of the building. Therefore, the model proposed in this work enhances the COST-231 MWM with a stochastic model that can be used to describe the signal attenuation.

The path loss in urban and suburban environments outside the high rise core of the cities can be calculated by [16],

$$L_{\rm o}(d) = 40(1 - 4 \times 10^{-3}\Delta h_b)(\log(d) - 3) - 18\log(\Delta h_b) + 21\log(f) + 80 \text{ dB}, (3.4)$$

where Δh_b is the BS antenna height measured in meters from the rooftop and f is the frequency in MHz. This model is useful in urban environments, however, according to studies reported in [15], the COST-231 MWM can be used to directly compute $L_{\rm T}$ in the proximity of buildings where there are no obstructions between the outer walls of the building and the receiver, which is normally the case in streets and around the buildings in urban areas. This approach is more accurate for short distances outside houses. For longer distances, the model described in (3.4) has to be used. In the latter case, the approach proposed in [12, 13] can be used at the indoor-outdoor interface. In the model, an imaginary boundary is placed around the building on the exterior walls. When the signal passes through the boundary, outdoor path loss model is applied. The indoor and outdoor path loss components are added together to form the final result [13].

Since this work mainly intends to model the interference in the proximity of the building, COST-231 MWM described in (3.3) is used to estimate the entire loss.

64

3.3 Modeling Methodology

To design a general model that can describe the FBS outdoor signal propagation, a statistical model for typical residential and small businesses buildings is required. Moreover, the model should consider the impact of the building floor plan, the location of the transmitter inside the building, and the location of the measurement point outside the building. The attenuation between the transmitter and receiver is a function of d_x , k_i , α_i and W. In this work we consider the floor plan and the location of the transmitter to be random, and hence k_i and d_x will be random as well. Consequently, unlike most of the work reported in the literature, the developed model will not be site-specific [11].

A possible approach to model the effects of buildings' floor plans is to set up a mobile measurement testbed. The transmitter will be placed in a certain point inside the building, several measurements are then recorded at different locations around the building. Then, the transmitter location is changed and the process is repeated. Finally, this tedious process should be repeated for several buildings. Obviously, such approach is unfeasible because of the tremendous effort required. Such experiments are usually performed for one transmitter location at one building [11] and the results are site-specific.

As an alternative to the exhaustive measurements approach, we design an automatic floor plan generator algorithm to generate a number of floor plans enough to build a statistical model. The algorithm, namely Procedural Building Generation (PBG) [17], is able to generate realistic floor plans based on some parameters that can be estimated in each neighborhood. Having the floor plan, a well-accepted deterministic analysis tool is used to calculate the signal strength in the predetermined measurement points. This method allows us to develop the femtocell signal model based on a large number of samples, which can never be obtained by any measurement campaign. Consequently, the resulting model would be applicable to more general scenarios.

Chapter 4 explains PBG algorithm in details. A concise summary is brought here to illustrate its mechanism:



Figure 3.2: Measurement setup.

- 1. The number of rooms is generated randomly according to the data that can be obtained from the neighborhood or census.
- 2. The area of each room is randomly generated based on its functionality.
- 3. The area and outer shape of the house is determined.
- 4. The rooms are placed in the house.
- 5. A connectivity graph that demonstrates the doors is produced.
- 6. A corridor to connect the rooms together is added, if necessary.
- Windows and doors are added randomly according to the connectivity graph. A sample of the resulting floor plans is depicted in the Fig. 4.7.

Next, a transmitter is dropped at a random point inside the house. Using existing models of indoor signal propagation, and including the attenuation of walls, the radio frequency field strength is calculated and the signal power at the measurement points are estimated. The process is run several times in a Monte-Carlo manner to construct the sample set.

In general, the exact location of FBS inside the house is unknown to the operator. Thus, the model should be independent of the exact location of the transmitter. Towards this goal, the measurements are gathered on a circle that enclosed the house and co-centered with it as depicted in Fig. 3.2. From the network planning point of view, it is possible to remove the house from the calculations and abstract the FBS inside the house as a single point transmitter at the center of the house, with random gain in each direction in far field. *i.e.*, a transmitter with random pattern. A sample pattern for a particular floor plan is extracted from one simulation run is depicted in Fig. 3.3. The abrupt changes in the signal power, which is a result of passing through walls is clearly shown in the figure. As it can be noted from the above discussion, the proposed model gives a statistical representation of the FBS signal with all the complexities due to buildings' floor plan without explicitly modeling the building itself. This enables the optimization of various design trade-offs and evaluate the performance of the next generation wireless networks more accurately.

3.4 Constructing Femtocell Signal Propagation Model Based on Building Architectural Model (BAM)

In order to calculate the FBS's signal strength in and around the house, the COST-231 model, described in Section 3.2 is adopted as the main model. Since, the objective is to find the interference caused by FBS's signal around the house, the model is also applied for short distances around the house. Therefore, the path loss can be calculated as,

$$L_{\rm T}(x) = L_{\rm m}(d_{\rm c}) + L_{\rm x}(x),$$
 (3.5)

where $L_{\rm m}(d_{\rm c})$ is the path loss estimated by the main path loss model of the medium at distance $d_{\rm c}$ from the center of the house, and $L_{\rm x}(x)$ is the excess path loss experienced at point x. For short distances around the house, the excess path loss models the effect of error in the transmitter position as well as the signal attenuation due to walls.

$$L_{\rm X}(x) = 20 \log\left(\frac{4\pi d_{\rm X}}{\lambda}\right) + \sum_{i=1}^{N} k_i \alpha_i - 20 \log\left(\frac{4\pi d_{\rm C}}{\lambda}\right), \qquad (3.6)$$

where d_x is the distance between point x and the receiver. Since the network operator is not aware of the internal architecture of the house and the transmitter position, the modeling idea is to assume that transmitter is at the center of the house and use the main path loss model to calculate the path loss. The error due to position of the transmitter and the walls' attenuation is then added as the excess path loss. After some straightforward simplifications we obtain,

$$L_{\mathbf{x}}(x) = 20 \log\left(\frac{d_{\mathbf{x}}}{d_{\mathbf{c}}}\right) + \sum_{i=1}^{N} k_i \alpha_i.$$
(3.7)

The proposed model considers two types of walls, light walls which are not bearing load such as plasterboards, particle boards or thin (<10 cm) concrete walls [18], and heavy walls which mainly correspond to the outer walls of the building. The model can be extended to include other types of walls, if needed. The suggested attenuation in COST-231 for the loss coefficients are $\alpha_1 = 3.7$ dB for light walls and $\alpha_2 = 6.9$ dB for heavy walls. These attenuation coefficients have been extracted empirically and consider the effects of reflections as well. Windows and doors are assumed to have attenuation factors of $\alpha_3 = 1$ and $\alpha_4 = 3$ dB, respectively [18].

To build the model, 5×10^3 different floor plans were generated using the PBG algorithm, and the signal strength is sampled at a uniformly distributed random point on a circle that encloses the house as depicted in Fig. 3.2. The basic parameters of the software such as the number of rooms and bedrooms in each house are extracted from the recent Census data [19]. The complimentary parameters that are necessary to generate floor plans are adjusted to model a suburban dwelling.

The probability distribution function of the excess loss is shown in Fig. 3.4. The peaks shown in the graph are clearly due to the attenuation of walls. The more walls, the signal passes, the more loss it would endure. The histogram composed of several parts, each of them can be modeled by a Gaussian distribution. Therefore, we will model the data with Gaussian Mixture distribution.



(b) Antenna pattern representation of the house.

Figure 3.3: Angular pattern representation of a FBS's signal leaking outside the building.

3.4.1 Estimating the attenuation with Gaussian-mixture distribution

From the probability distribution function of the excess loss, shown in Fig. 3.4, it can be noted from the figure that the histogram can be divided into several attenuation groups each of which is associated with a path from the transmitter to the measurement point. Moreover, each door and window in the floor plan behaves as an aperture [20], which created the peaks observed in the histogram. Therefore, each group can be associated by passing through a specific number of walls and/or windows. The signal strength measurements for several buildings taken as part of WINNER II [21] and OFCom [15] projects clearly corroborate with the obtained results. A key observation is that each signal group can be modeled as a Gaussian distribution, resulting in a Gaussian-Mixture model (GMM) whose probability distribution function (PDF) is the weighted sum of the probability distribution functions of M Gaussian distributions. Thus,

$$p_l(l_{\mathbf{x}}|\lambda) = \sum_{i=1}^M \mathbf{w}_i p_{\mathcal{N}}(l_{\mathbf{x}}|\mu_i, \sigma_i)$$
(3.8)

$$= \sum_{i=1}^{M} \frac{\mathbf{w}_i}{\sqrt{2\pi\sigma_i^2}} \exp\left(-\frac{(l_{\mathbf{x}}-\mu_i)^2}{2\sigma^2}\right), \qquad (3.9)$$

where $p_{\mathcal{N}}(l_x|\mu_i, \sigma_i)$ represents the PDF of a Gaussian random variable with mean μ_i and standard deviation σ_i . It is basically several Gaussian random variables added together with weight w_i . It is obvious that $\sum_{i=1}^{M} w_i = 1$. Expectation Maximization (EM) algorithm has been used to find the proper coefficients. The parameters of the resulting Gaussian Mixture Model (GMM) are given in Table 3.2. The probability density function and cumulative density function of the distribution are shown in Fig. 3.4.

3.4.2 Recived power distribution

The local received power ω from the transmitter can be modeled by

Chapter 3: Predicting Femtocell Interference for Next-Generation Networks via a Novel Building 71Architecture Model



Figure 3.4: Probability distribution function of building attenuation.

i	1	2	3 4		5	6	
μ_i	0.25	13.31	22.36	29.94	37.77	40.53	
σ_i	4.19	12.20	2.60	5.95	0.68	45.16	
ω_i	0.11	0.45	0.20	0.15	0.03	0.06	

Table 3.2: GMM paramters

$$\omega = P_{\rm av} 10^{-\frac{L_{\rm X}}{10}},\tag{3.10}$$

where $P_{\rm av} = 10^{\frac{P_{\rm t} - L_{\rm m}}{10}}$ is the average received power without considering the excess loss. Note that the received power ω is not in decidel. The PDF of ω is calculated as

$$p_{\omega}(\omega) = p_l \left(-10 \log\left(\frac{\omega}{P_{\rm av}}\right)\right) \times \frac{\mathrm{d}}{\mathrm{d}\omega} \left(-10 \log\left(\frac{\omega}{P_{\rm av}}\right)\right) \\ = p_l \left(-10 \log\left(\frac{\omega}{P_{\rm av}}\right)\right) \times \frac{-10}{\ln(10)P_{\rm av}\omega} \\ = \sum_{i=1}^{M} \frac{\mathrm{w}_i \xi}{\sqrt{2\pi\sigma_i^2}\omega} \exp\left(-\frac{\left(-10 \log\left(\frac{\omega}{P_{\rm av}}\right) - \mu_i\right)^2}{2\sigma^2}\right), \quad (3.11)$$

where $\xi = \frac{-10}{\ln(10)P_{\text{av}}}$. Thus, the received power can be modeled by a lognormal-mixture distribution.

3.5 An Application: Optimal FBS Placement

The signal power of FBS leaked to outside of the building depends on the floor plan of the building and position of the FBS. A transmitter seated near a window would have the most powerful signal outside, whereas one placed in an enclosed room has a limited effect on macro users outside. In this section, we study the effect of FBS's position and find the best possible position that minimizes the expected interference. The interference at a particular point can be minimized by;

$$\overline{I}_{\mathrm{m}} = \min_{x_t} \int_c I(x_t, x_r) f_{X_r}(x_r) dc, \qquad (3.12)$$

where x_t and x_r represent transmitter and receiver coordinates, respectively. $I(x_t, x_r)$ is the interference at x_r caused by a transmitter at x_t , and $f_{X_r}(x_r)$ is the probability density function of the receiver position. The interference is to be minimized on c which is defined as a circle around house with radius R centred at the center of the house. Assuming equal probability of receiving signal at each angle, we may write,

$$\overline{I}_{\rm m} = \min_{x_t} \quad \frac{1}{2\pi} \int_{0}^{2\pi} I(x_t, [R, \theta]) d\theta.$$
(3.13)

For every realization in the building database, an extensive search on every possible transmitted point results in the optimal position. The diagram in Fig. 3.5 illustrates the expected interference at some potential FBS's position. FBSs are represented by red crosses. The radius of each circle is proportional to the expected interference. Large circles show high level of interference, whereas small circles indicates that little interference is expected, which could be potential candidates for placing a FBS. High level of interference is expected around the windows, which makes this regions as the worst for placing FBS. On the other hand, enclosed areas inside the house with no windows, like corridors or corner of rooms have a better chance of limiting interference to outside. To study the effect of FBS's location, the calculations are repeated for 500 realizations and the interference is measured on a circle with radius equal to 15 m.





Figure 3.5: Expected signal attenuation map of a sample house.



Figure 3.6: The histogram of \overline{I}_m .

The histogram of the expected path loss (\overline{I}_{m}) of a transmitter placed at the optimum point is shown in Fig. 3.6.

The average value of the path loss for optimal position over all realizations is 80.4 dB. Considering 62 dB path loss at the distance of 15 m, the mean excess path loss due to building is worked out to be 18.4 dB. The simulations also show that the expected interference can increase up to 22.9 dB by moving the FBS in the house. These values can be used as a rule of thumb to consider the attenuation effects of the building.

3.6 Chapter Summary

This chapter presented a novel model that considers the complex effects of buildings' floor plan on the signal power. The proposed model, namely Building Architectural Model (BAM), predicts the power of the femtocell signal leaking outdoors. It considers the main effects of the buildings, including signal attenuation due to passing through walls/windows/doors and a mixed indoor-outdoor channel. The model is based on widely accepted propagation models. However, validation of the results by measurements would be highly desirable.

The model can be used to derive several performance metrics of the network. Moreover, it can be used by the network operator to predict the effect of the installation of new femtocells on its network. It is also used as a basis for a composite shadowing/multipath fading model derived in Chapter 5. As an application of the model, it was used in Section 3.5 to minimize the femtocell power leakage outside the building by optimizing the FBS's placement and its interference. The achieved results confirmed that a proper placement of a FBS can decrease the mean interference around the house by as much as 23 dB.

References

- M. Mirahmadi, A. Shami, and A. Al-Dweik, "A building architecture model for predicting femtocell interference in next-generation networks," in *IEEE Int. Conf. Communications (ICC)*, Ottawa, Canada, Jun. 2012, pp. 5059–5063.
- [2] Technical Specification Group Radio Access Networks (TSG-RAN), "3G Home NodeB study item technical report (release 8)," 3GPP, Tech. Rep. TR 25.820, 2008.
- [3] A. Al-Dweik, B. Sharif, and C. Tsimenidis, "Accurate BER analysis of OFDM systems over static frequency-selective multipath fading channels," *IEEE Trans. Broadcast.*, vol. PP, no. 99, p. 1, Dec. 2011.
- [4] A. Valcarce *et al.*, "Applying FDTD to the coverage prediction of WiMAX femtocells," *EURASIP J. Wireless Commun.*, vol. 2009, pp. 1–13, 2009.
- [5] M. Dohler and A. Aghvami, "An outdoor-indoor interface model for radio wave propagation for 2.4, 5.2 and 60 GHz," Master's thesis, King's College, London, 1999.
- [6] Digital Mobile Radio Towards Future Generation Systems: COST Action 231, European Co-Operation in the field of Scientific and Technical research (COST) Final Report COST 231.
- [7] C. Oestges and A. Paulraj, "Propagation into buildings for broad-band wireless access," *IEEE Trans. Veh. Technol.*, vol. 53, no. 2, pp. 521–526, Mar. 2004.
- [8] Guidelines for evaluation of radio transmission technologies for IMT-2000, International Telecommunication Union (ITU) Recommendation ITU-R M.1225, Feb. 1997.

- [9] D. Bacon and N. Thomas, "Simulation and modelling of propagation paths involving the indoor/outdoor interface," in *Proc. Int. Symp. on Advanced Radio Technologies (ISART)*, Boulder, USA, Feb. 2007.
- [10] C. Takahashi *et al.*, "Propagation-prediction and site-planning software for wireless communication systems," *IEEE Antennas Propag. Mag.*, vol. 49, no. 2, pp. 52–60, Apr. 2007.
- [11] M. Lillywhite and G. Liang, "Prediction of indoor and outdoor signal strengths from indoor transmitter," in *IEEE Antennas and Propagation Society Int. Symp.*, vol. 1, 2005, pp. 661–664.
- [12] R. Wahl and G. Wolfle, "Combined urban and indoor network planning using the dominant path propagation model," in *Proc. 1st European Conf. Antennas* and Propagation, Nov. 2006, pp. 1–6.
- [13] A. Valcarce and J. Zhang, "Empirical indoor-to-outdoor propagation model for residential areas at 0.9-3.5 GHz," *IEEE Antennas Wireless Propag. Lett.*, vol. 9, pp. 682–685, Jul. 2010.
- [14] T. S. Rappaport, Wireless communications: principles and practice, 2nd ed. Prentice Hall PTR New Jersey, 2002.
- [15] "Predicting coverage and interference involving the indoor-outdoor interface," OFCOM, Tech. Rep. SES-2005-08, 2005.
- [16] Selection procedures for the choice of radio transmission technologies of the UMTS, 3GPP Std. UMTS 30.03 v3.2.0, 1998.
- [17] M. Mirahamdi and A. Shami, "A novel algorithm for real-time procedural generation of building floor plans," innovation Center for Information Engineering (iCIE), Tech. Rep., Dec. 2011. [Online]. Available: http: //arxiv.org/abs/1211.5842
- [18] H. Claussen, L. Ho, and L. Samuel, "An overview of the femtocell concept," Bell Labs Tech. J., vol. 13, pp. 221–245, 2008.

- [19] 2001 census of canada. Statistics Canada. [Online]. Available: http: //www.statcan.gc.ca/
- [20] Z. Zhang, R. K. Sorensen, Z. Yun, M. F. Iskander, and J. Harvey, "A ray-tracing approach for indoor/outdoor propagation through window structures," *IEEE Trans. Antennas Propag.*, vol. 50, no. 5, pp. 742–748, May 2002.
- [21] J. Meinilä, P. Kyösti, T. Jämsä, and L. Hentilä, "WINNER II channel models," *Radio Technologies and Concepts for IMT-Advanced*, pp. 39–92, 2009.

Chapter 4 Novel Algorithm for Real-time Procedural Generation of Building Floor Plans

4.1 Preamble

Dynamic generation of virtual environments gained popularity in the last decade. For example, in modern games, generating dynamic virtual environments for each round of the game allows even the savvy gamers to enjoy the game endlessly. A lot of games' scenes take place in a town which has to be generated either manually or with a rulebased procedure. In some massive multiplayer games, producing the whole world is required. Besides the scale of the virtual environment, its details are also essential. Creating environments where the player can go inside the buildings adds a layer of sophistication which multiplies its complexity. As a result, creating and managing such environments constitute a large portion of game design process.

A model describing the architecture of a city or inside the buildings can also be used in different disciplines for various applications. As an instance, such an algorithm can be used to statistically model the signal propagation in built-up areas or used as a hypothetical signal propagation benchmark which can be customized to adapt to different scenarios. The model developed in this Chapter has been used in the works presented in Chapter 3 and Chapter 5 to statistically model the propagation of a radio signal transmitted from an indoor transmitter to outside, which extremely depends on the floor plan of the building that encloses the transmitter.

In this chapter, we introduce a novel procedural floor plan generation for suburban houses. The remaining of the chapter is organized as follows. Section 4.2 reviews some of the previous work related to automatic generation of buildings. Section 4.3 describes the proposed algorithm. Section 4.4 discusses the results, and finally Section 4.5 concludes the chapter.

4.2 Literature review

Several algorithms have been proposed in literature to create architectures. The functionality of these algorithms can be categorized in three different classes: creating the city maps, creating the 3-dimensional appearance of a city or a building, and creating the interior design of a building.

Müller *et al.* propose a rule-based algorithm capable of generating streets and urban-looking buildings which results in realistic environments with high visual quality [2]. Several simple 3D shapes, such as cubes, cones, etc, are placed or intersected together to form the shell of the buildings. Then the proposed shape grammar, which is an extensive set of rules governing the shapes, is processed to refine the buildings. First, the grammar looks for areas that shapes intersect. The algorithm discards these areas when putting windows or other ornaments. The next set of grammar rules tries to snap windows and other accessories to the visual lines in the structure which makes the resulting building more realistic. The algorithm produces high quality modern looking city landscape. However, the disadvantage is that it cannot be run in realtime. Running the algorithm possibly takes up to several days depending on the size of the requested scenario. In addition, it only creates the shell of the buildings.

The work presented in [3] focuses on real-time interior generation of large buildings, such as office buildings. The authors propose a method to manage the huge memory and processing power requirement for real-time floor plan generation of large buildings as well as maintain the environment persistent. The building spaces are generated in a top to down fashion, from larger spaces to internal smaller spaces. Therefore, it is possible to generate them in a lazy fashion and create the spaces that are necessary. If the player goes into a space, that space is then divided into smaller rooms and other accessories are added to it. The algorithm generates the floor plan in the following steps: building setup, floor division, hallway division, room cluster division, and built region generation. In each step, more details are added to the floor plan, while the invisible architectures are removed form the cache. This approach is highly scalable and can be applied to scenarios with a number of large buildings without significant increase in memory or processing power requirements.

Another algorithm that aims to create the interior floor plan for buildings, especially for houses, is proposed in [4]. The algorithm divides the available space using Squarified Treemap algorithm [5]. It then connects the rooms together and placed doors between them based on a connection graph which is randomly produced at a previous stage based on some rules. An example rule can be expressed as bedrooms should not be connected together. In the next step, if the rooms that are supposed to be connected are not adjacent, a procedure is applied to place a corridor to connect the rooms. However, the procedure wastes some space in the house which is not desirable and makes the floor plans unreal.

Since the creation of the building appearance is highly required, the issue has been addressed and several academic and commercial tools exist to generate urban environment with great realism. However, to the best of our knowledge, there are just a few algorithms that can create building interior and none of them gives a detailed and realistic floor plan and most often they just fit several rectangles which demonstrate rooms into a predefined area.

4.3 Proposed Algorithm for Constructing Floor Plans

This section describes our algorithm for creating floor plans. The proposed algorithm is based on the work presented in [4]. However, the corridor placement step is revised greatly, and also, an optimization stage is added to the design to make the resulting floor plans more realistic. Moreover, several rules have been introduced to prohibit the creation of bizarre looking rooms. We refer to the proposed enhanced version as procedural building generation (PBG) in this thesis. Just like the work in [4], PBG has been tuned to create random floor plans for suburban houses. However, it can be used to create the floor plan for any type of building with some modifications. The proposed algorithm creates floor plans in the following steps:

- 1. Determining the area and outer shape of the house.
- 2. Placing the rooms inside the house.
- 3. Creating the connectivity graph. This graph specifies if there is a way from one room to the other.
- 4. Creating a corridor to connect the rooms together, if necessary.
- 5. Placing windows and doors.

Each of the above steps is described in detail in the corresponding section. The algorithm is primarily based on the following concepts.

- The outer shape of the house, or its facade, is rectangular.
- Each house is composed of several connected rooms.
- No two spaces of the house can overlap.
- The building shape is preferred to be square or square-like. *i.e.*, the square shape is preferred over long rectangles.
- No room can protrude the supposed boundary of the house.
- All significant portions of space should be used [3].
- The area of each room is a random variable whose statistical properties are based on its functionality.
- The floor plan should be connected. *i.e.*, there is at least one way to go from one room to the other. In other words, the connectivity graph is a connected graph.
- Generating narrow long rooms which would be perceived unnatural should be avoided as much as possible.
- There are several windows, and at least a door connecting the house to the outside.

These notions are implemented in the proposed algorithm. The following sections explain the algorithm in detail.

4.3.1 Determining the outer shape of the house

The outer shape of most of the buildings is rectangular and does not have curved facade. The buildings with unusual shapes are in most cases public buildings with architectural importance. The proposed algorithm focuses on typical buildings and does not include these types of buildings.

The shape of building facade is modeled by its aspect ratio, which is defined as,

$$AR = max(b/h, h/b) \tag{4.1}$$

where AR, b, and h are the aspect ratio, base and height of the rectangle representing the building facade, respectively. An area with a narrow shape, like a rectangle with large aspect ratio, is simply regarded as unsuitable for a house in real life, and thus, discarded in the proposed algorithm.

The area and aspect ratio of the buildings are random variables with predetermined distribution. For a given neighbourhood, the statistical properties of these random variables can be simply sampled from aerial photos. Fig. 4.1 shows an aerial image of a residential region in London, Ontario. This type of suburban landscape is quite widespread in North America and it is probably the main design scheme of the residential areas.

The output of the first step of the algorithm is a rectangle representing outer walls of the building.

4.3.2 Rooms placement

Each house is composed of rooms and probably corridors to connect them together. The number of rooms in a house varies from house to house, but some simple statistics can be found in Census. The statistics such as the number of rooms and bedrooms are enough for our modeling purpose. Moreover, the parameters can be adjusted manually to simulate any scenario.

In the proposed algorithm, the random variables are generated based on the joint probability distribution of the number of bedrooms and the number of rooms which is extracted from 2001 Census of Canada [6] in Table 4.1. It lets the generator



Figure 4.1: An aerial photo showing the similarity of homes in a neighbourhood. [Courtesy of Google Inc.]

to follow the real distribution of a neighbourhood.

The area of each room is a random variable whose distribution depends on its functionality. As an instance, the living room is typically the largest room in a house, while the storage rooms are the smallest ones. According to [4], rooms of a house can be divided into three categories based on their functionality; service area includes kitchen and laundry room, private area composed of bedrooms and bathrooms, and social area like living room and dining room. After generating the area of each room,

Table 4.1: Joint probability of the number of bedrooms (rows) and the number of rooms (columns).

	1	2	3	4	5	6	7	8	9	10
0	8.0e-3	1.4e-2	6.3e-4	0	0	0	0	0	0	0
1	0	0	8.0e-2	4.3e-2	1.2e-2	3.4e-3	1.2e-3	5.8e-4	2.1e-4	1.1e-4
2	0	0	0	9.7e-2	9.1e-2	3.9e-2	1.6e-2	8.6e-3	3.0e-3	1.5e-3
3	0	0	0	0	7.6e-2	1.1e-1	8.0e-2	5.5e-2	2.6e-2	2.6e-2
4	0	0	0	0	0	1.4e-2	3.5e-2	5.2e-2	4.0e-2	6.9e-2

the area of each part of the house, *i.e.*, social, service, and private part, is calculated.

The rooms are then put into a hierarchical tree graph based on their functionality and the functionality of other rooms. The hierarchy starts from outside which is normally directly connected to the living room as the center of the daily life. The other rooms are included as branches below it.

One important issue which is commonly left out in the algorithms is how to assign area to each room and how to determine if a house is equipped with a room. The proposed algorithm uses a collective distribution for all types of the rooms. This approach uses the Census data to find some distribution parameters.

To associate the functionalities to the generated rooms, the algorithm uses a list called priority list. It simply picks the N most important rooms, where N is the number of rooms in the house. Then, the area of each room, is specified based on the selected functionality which determines its distribution. To customize the output and match it to a different type of building, the priority list and area distributions can be adjusted.

In the next step, a rule-based algorithm puts the rooms in the hierarchy tree based on their functionality. As an instance, the kitchen is connected to the living room either directly or via the dining room, and hence, it is placed under the living room directly or via the dining room. In some cases, the position of a room in the hierarchy tree also depends on the functionality of other rooms. For example, in a typical house there should be a bathroom connected to the common area. The extra bathrooms are typically inside master bedroom or other large bedrooms. Therefore, there is a bathroom connected directly to the common area and there may be several other bathrooms connected to bedrooms. Some basic rules are as follows,

- 1. Place outside node as the root.
- 2. Place living room below outside.
- 3. If there is a kitchen, place it below living room.
- 4. If there is any bedroom, place the largest one under living room and name it master bedroom.
- 5. If there is just one bathroom, place it below living room.
- 6. Place the remaining bathroom below bedrooms, starting from the largest bed-



Figure 4.2: Hierarchy graph of a sample house.

room.

7. Place laundry and pantry below kitchen, if any.

The hierarchical tree of rooms of a sample house is demonstrated in Fig. 4.2.

The next step is placing the rooms in the house. An algorithm called Squarified Treemap [5], places the rooms. The objective of the Squarified Treemap algorithm is to divide a region of space into several smaller regions with predefined area without any unused space. It also tries to minimize the aspect ratio of each block to be more square. Squarified treemap [5] algorithm is an extension of standard Treemap algorithm [7] which gives priority to the square or square-like shapes. The original Treemap algorithm organizes the spaces in a tree graph like the one shown in Fig. 4.3. As can be seen in the figure, the method can generate elongated rectangular sub-divisions. These subdivisions are not favourable in floor plans. Therefore, Squarified treemap algorithm is adapted for the proposed algorithm. For a detailed explanation about Squarified Treemap algorithm refer to [5].

The automatic floor plan generator algorithm uses Squarified Treemap in a step by step approach. It first places the rooms in the first level of the hierarchy. At this step, the algorithm puts rooms with total surface area of all the rooms below it. Then, it moves to each room and places the smaller rooms below it inside. The steps of putting rooms inside a house is illustrated in Fig. 4.5.



Figure 4.3: A tree graph used by Treemap algorithm and its corresponding rectangle division [7].



Figure 4.4: Finding the corridor path.

4.3.3 Corridor placement

In a typical house all of the rooms are connected either directly or via some corridors to the living room, as the center of activities. The proposed algorithm includes this fact into the floor plan designs. It identifies the rooms that are required to be connected directly to the living room and if they are not adjacent to the living room, it places a corridor, connecting the rooms to the living room.

Depending on the number and type of the rooms and their placement, it may be required to place a corridor to connect the rooms together. For example, in Fig. 4.4a Room #3 cannot be connected to the living room. Thus, a corridor is required to connect the leftmost room (Room #3) to the living room.

The Corridor Placement algorithm functions as follow:

1. The rooms that requires connections are identified. We refer to these rooms as the corridor rooms.



Figure 4.5: Room placement steps for the hierarchy tree shown in Fig. 4.2.

- A graph is constructed with the walls of the corridor rooms as well as the living room. The outer walls of the building is not considered in this graph. (Figure 4.4b)
- At the next stage the graph is pruned and the edges that connects to a vertex with degree one is removed from the graph. This graph is called corridor graph. (Figure 4.4c)
- 4. The shortest path in the graph connecting all the rooms in the graph is chosen using standard shortest path algorithm [8]. A room is considered connected if it shares an edge or a vertex with the graph. The use of the shortest path algorithm is justified by the fact that corridor area is a wasted area in the house and has to be minimized. Thus, we choose the shortest path which is translated to the smallest area.
- 5. If a room is connected by a vertex and does not have any shared edge, it is required to modify the graph to be able to place a door for the room. This is done through either shifting the edge of the graph inside the room or lengthen the edge connected to the shared vertex. The process is illustrated in Fig. 4.6. All of the possible choices to modify the corridor graph constitutes the action



(a) Shifting the corridor downward to align with the living room wall



(b) Lengthen the corridor to make room for a door to the living room



(c) A combination of shifting upward and lengthen the corridor

Figure 4.6: Optimizing the corridor with different actions.

set.

$$A = \{a_i = Shift | Lenghten\},\$$

$$\forall e_i \in Edges, i = 1 \dots Number of edges (4.2)$$

Selecting proper actions for each edge results in a small corridor. Therefore, an optimization is required to prohibit the generation of bulky corridors and minimize corridor area.

6. The corridor should not change the shape of surrounding rooms in a way that their space become unusable. Therefore, the corridor graphs that leaves peculiar shape rooms behind are filtered out. Then, An algorithm compares the area of the corridor resulting from each modified corridor graph and chooses the corridor with the smallest area. The resulting graph is called the optimized corridor graph.

7. At the last step, the polygon representing the corridor is constructed based on the optimized corridor graph and extruded from the overlapping rooms. The output is the final floor plan. It should be noted that the area of the corridor itself is regarded as the extension of the living room. Therefore, in all subsequent steps, the corridor walls are considered as the living room walls.

4.3.4 Placing windows and doors

The next step places the connections between the rooms, *i.e.*, doors, and windows in the floor plan. The studies [4] shows that not all of adjacent rooms can be connected together. As an instance, bedrooms can not be connected to the kitchen.

The hierarchy tree is normally the basis of connection graph which may then augmented by several other edges. For example, it is possible to connect kitchen and dining room. The decision to add these optional edges is taken randomly based on the functionality of both rooms.

The doors are then placed randomly at the shared walls between the rooms. The door size is fixed and is adjusted manually by the algorithm designer. However, their position in walls are randomly chosen.

The same approach is taken for placing windows. The only difference is that the connection graph is constructed regarding the placement of the rooms as well as some restrictive rules. As a general rule, the rooms that share a wall with outside are equipped with a window unless it is prohibited. For example, a sample rule is that a window cannot be installed in bathrooms.

4.4 Results

In this section, several generated floor plans as well as their parameters are presented. The parameters used to generate the floor plans are specified in Table 4.2.

It should be noted that the number of bedrooms and the number of rooms are random variables generated based on the distribution in Table 4.1. Bathrm-9

Bedrom-7

MBedrm-8

Parameter	Distribution			
bedroom area	Uniform(8,18)			
other room's area	Uniform $(3,11)$			
building aspect ratio	Uniform $(1,2)$			
	·			
	1 I			

Kitchn-1

Living-4

Landry-3

Pantry-2

Dining-5

Table 4.2: List of the parameters used in generating floor plans.

Figure 4.7: A sample floor plan automatically generated by the algorithm.

A sample floor plan generated with the parameters in Table 4.2 is shown in Fig. 4.7. The house has two bedrooms and a bathroom. The room boundaries before adding corridor is depicted with thin lines whereas the dashed bold lines show the boundaries after placing the corridor. The floor plan looks natural and it could be used in any application. For example, it can be used by a game engine to produce authentic floor plans. The secondary bedroom was only connected to the master bedroom and bathroom. Since connecting two bedrooms together or through bathroom is prohibited in the rules, a corridor is placed to connect the secondary bedroom to the living room. The position of the corridor is optimized to occupy the least possible space.

Several other possibilities of the placing the corridor is shown in Fig. 4.6. Fig. 4.6a shows the case where the graph is shifted downward to align with living room walls, however, this case is not valid as the corridor can not accommodate a door to the smaller bedroom. Fig. 4.6b shows the case where a graph edge is lengthen to make room for a door from corridor to the living room. A combination of shifting upward and lengthening the corridor is shown in Fig. 4.6c. And in the final floor



(a) A generated floor plan with augmented details.



(b) A real floor plan [4].



(c) 3D realization of the generated floor plan [9].

Figure 4.8: Comparison of a sample generated floor plan with an architecture designed floor plan

plan shown in Fig. 4.7, the corridor generated by a combination of shifting upward to align with the room boundaries and lengthening to make room for a connection to the living room is selected. It is worth mentioning that according to the discussions in Section 4.3.3, the corridor with least consumed space is selected.

To give a sense of the generated floor plans, we have added some furniture to a sample floor plan generated by our algorithm. The floor plan as well as its 3D realization is illustrated in Fig. 4.8a and 4.8c. A real architect-designed floor plan is also shown in Fig. 4.8b for the sake of comparison. The generated floor plan shows utmost similarity to the real one. It is worth mentioning that the improvements in the algorithm compared to the algorithm in [4] prohibits the generation of long unused corridors which would make unused space.

4.5 Chapter Summary

This work proposed an algorithm for real-time generation of floor plans. The dynamic generation of floor plan algorithm developed in this Chapter can also be used in all of the applications concerning buildings. For example, based on this algorithm, we proposed a model to predict the strength of a radio signal of an indoor transmitter, like a WiFi access point, outside the building. The model is reported in Chapter 3, and [1].

The procedural building generation (PBG) algorithm focused on suburban houses. However, it can be tailored to accommodate other types of buildings. The algorithm generates every aspect of the house randomly, resulted in dissimilar floor plans. The random generation includes shape of the house, its area, number of rooms and their functionality and the position of windows and doors. Comparing to similar algorithms, in PBG, special attention is devoted to placing a corridor inside the house. The proposed corridor placement algorithm optimizes the place and shape of the corridor to produce natural-looking designs and reduces the area of the corridor which is regarded as a wasted space in house.

The results show floor plans with utmost similarity to real floor plans. If used to produce game scenarios, it can be imagined that the gamer that is playing inside the 3D realization of such generated floor plan, would find it very natural.

References

- M. Mirahmadi, A. Shami, and A. Al-Dweik, "A building architecture model for predicting femtocell interference in next-generation networks," in *IEEE Int. Conf. Commun. (ICC)*, Ottawa, Canada, Jun. 2012, pp. 5059–5063.
- [2] P. Müller, P. Wonka, S. Haegler, A. Ulmer, and L. Van Gool, "Procedural modeling of buildings," ACM Trans. Graph., vol. 25, pp. 614–623, Jul. 2006.
- [3] E. Hahn, P. Bose, and A. Whitehead, "Persistent realtime building interior generation," in *Proc. ACM SIGGRAPH Symp. Videogames*, ser. Sandbox '06, New York, NY, USA, 2006, pp. 179–186.
- [4] F. Marson and S. R. Musse, "Automatic real-time generation of floor plans based on squarified treemaps algorithm," *Int. J. Computer Games Tech.*, vol. 2010, pp. 1–10, 2010.
- [5] M. Bruls, K. Huizing, and J. van Wijk, "Squarified treemaps," in Proc. IEEE/TCVG Symp. Visualization Joint Eurographics (VisSym), Vienna, Austria, May 1999, pp. 33–42.
- [6] 2001 census of canada. [Online]. Available: http://www12.statcan.ca/english/ census01/products/standard/
- B. Johnson and B. Shneiderman, "Tree-Maps: a space-filling approach to the visualization of hierarchical information structures," in *Proc. IEEE Conf. Visualization*, ser. VIS '91, Los Alamitos, CA, USA, 1991, pp. 284–291. [Online]. Available: http://portal.acm.org/citation.cfm?id=949607.949654
- [8] E. Dijkstra, "A note on two problems in connexion with graphs," Numerische Mathematik, vol. 1, no. 1, pp. 269–271, 1959.
[9] [Online]. Available: www.floorplanner.com

Chapter 5 Indoor-Outdoor Channel Characterization and its Application to Interference Analyses of Femtocells

5.1 Preamble

The employment of femtocell BSs (FBS) in the cellular network present a new paradigm in communication science. In Chapter 3, we have proposed a statistical model for analyzing femtocell interference which can estimate the strength of a signal propagating from inside a house. In this chapter, we put a step forward and propose a composite model for multipath fading and building shadowing, which we defined as the random loss of power due to traversing through building structure and modeled it in Chapter 3.

Employing FBSs in the legacy cellular network causes several types of interference. These interferences, illustrated in Fig. 5.1, are [2,3]:

- A Macrocell Downlink Interference to the Femtocell UE Receiver.
- B Macrocell Uplink Interference to the Femtocell BS Receiver.
- C Femtocell Downlink Interference to the Macrocell UE Receiver.
- D Femtocell Uplink Interference to the Macrocell NodeB Receiver.
- E Femtocell Downlink Interference to Nearby Femtocell UE Receivers.
- F Femtocell Uplink Interference to Nearby Femtocell Receivers.

A version of this chapter has been submitted for publication [1].



Chapter 5: Indoor-Outdoor Channel Characterization and its Application to Interference Analyses of Femtocells 96

Figure 5.1: Possible interferences in a two-tier femtocell/macrocell network [3].

- G Macrocell Downlink Interference to the adjacent channel Femtocell UE Receiver.
- H Macrocell Uplink Interference to the adjacent channel Femtocell Receiver.
- I Femtocell Downlink Interference to the adjacent channel Macrocell UE Receiver.
- J Femtocell Uplink Interference to the adjacent channel Macrocell NodeB Receiver.

Among all of the above interferences, scenarios A, D, B, E, and F are more important and have to be carefully studied [2]. The model proposed in this work mainly intends to estimate the femtocell uplink interference to a macrocell user equipment (MUE) nearby, which is listed as scenario D. It can also be used to analyze interference type F.

The macrocell-femtocell interference analysis have been considered widely in the literature. However, most of the work reported ignores the random nature of the floor plans. For example, Kim *et al.* [4] considered a two-tier macro-femto architecture where the buildings' internal and external walls attenuate the FBS signal by a fixed value of 10 dB. Lee *et al.* [5] evaluated the macro-femto system with fractional frequency reuse by considering only the free space path loss. Chandrasekhar and Andrews [6] considered the capacity analysis and developed an interference avoidance strategy. In their work, the indoor attenuation is approximated by a lognormal distri-

Chapter 5: Indoor-Outdoor Channel Characterization and its Application to Interference Analyses of Femtocells 97

bution, which to some extent considers the randomness of the FBS location. However, the lognormal distribution does not accurately model the indoor-to-outdoor propagation as it is demonstrated in this work. Park *et al.* [7] proposed beam subset selection as a strategy for interference reduction in two-tier femtocell networks where the cross-tier interference from the FBSs to the macrocell users is approximated by a Poisson shot noise. The caveat is that a fixed attenuation of 5 dB is considered in this model by assuming that the FBS signal penetrates exactly one wall. In an effort to model indoor-to-outdoor propagation, an empirical model is developed by Valcarce and Zhang [8] for the frequencies in the range of 0.9 to 3.5 GHz. Although the model is developed based on a massive number of measurements, the model assumes that the FBS is located in an outer room or a room adjacent to an outer room, which limits the number of penetrated walls. Moreover, utilizing the model requires the knowledge of transmitter position and partial knowledge of the surrounding floor plan.

Unlike the models described above, this chapter presents a statistical attenuation model for the signal propagating from an indoor FBS to the outdoor environment. The statistical attenuation model is based on a random floor plan generator algorithm that can generate large number of realistic floor plans. To build the statistical model, the attenuation for each generated floor plan is computed using standard path loss models. This process is repeated for a large number of realizations, which ultimately produces the desired statistical model. The developed model is used to compute the distribution of the received signal envelope at the receiver front-end, as well as the outage probability of the macrocell user, for which a closed-form solution is derived. Moreover, the generated model is used to evaluate the uncoded bit error rate (BER) in different operating scenarios.

The rest of this chapter is organized as follows. Section 5.2 presents the development of an accurate and simplified channel models that describe the signal propagation in indoor-to-outdoor environments based on a stochastic floor plan generation algorithm. Section 5.3 applies the proposed model to derive the received signal of an Orthogonal Frequency Division Multiplexing (OFDM) transmitter. In Section 5.4, the performance of a hybrid system with femtocells and a macrocell is evaluated analytically based on the approximated version of the proposed composite model. Numerical results are given in Section 5.5, and finally, the concluding remarks are given in Section 5.6.

5.2 Proposed Composite Shadowing/Multipath Fading in Mixed Indoor-Outdoor Environments

The same propagation model used in Chapter 3 is used as the basis. In this section, we start with a summary of the propagation model, and then, continue to develop the proposed composite model.

Similar to the approach taken in Chapter 3, to generate a large number of floor plans sufficient to build a statistical floor plan model we use the automated Procedural Building Generation (PBG) algorithm was developed in Chapter 4, which can generate realistic floor plans based on parameters that are common for particular architectural styles used in certain neighborhoods.

Once a floor plan is generated, the next step is to drop a transmitter at a random point inside the house and then the propagation model given in (3.3) is invoked to calculate the signal strength at predetermined measurement points, which are located on a circle that encloses the house and is co-centered with it as depicted in Fig. 3.2. The process is repeated several thousand times in a Monte-Carlo fashion to estimate the statistical properties of the attenuation.

In a typical application such as cell planning, the designer assumes the receiver's position. Therefore, the path loss component of the attenuation in a static environment is deterministic in nature. Hence, we focus on the random part of the attenuation incurred due to passing through buildings, namely $L_{\rm x}$.

$$L_{\mathbf{x}} = \sum_{i=1}^{W} k_i \alpha_i + \zeta \quad (dB)$$
(5.1)

where ζ is a random variable that captures the effect of not knowing the exact position



Chapter 5: Indoor-Outdoor Channel Characterization and its Application to Interference Analyses of Femtocells 99

Figure 5.2: Simulated PDF of the building shadowing $L_{\rm x}$ in the complete model described in (5.3).

of FBS inside the building, which in turn depends on the architecture of the house. The effect of ζ will be captured and statistically described in the model. Using this re-definition, the total path loss is

$$L_{\rm T} = 20 \log \left(\frac{4\pi d_{\rm c}}{\lambda}\right) + L_{\rm x} \quad ({\rm dB}) \tag{5.2}$$

where d_c is the distance from the center of the building. From cell planning point of view, the first term is deterministic and known, whereas the second term L_x is random and it will be denoted as the building shadowing.

The histogram of the building attenuation L_x is presented in Fig. 5.2. In this chapter, since a less complex model is more desirable, without loss of generality, fewer number of Gaussian components are used to fit the GMM distribution. The PDF of

Chapter 5: Indoor-Outdoor Channel Characterization and its Application to Interference Analyses of Femtocells 100

GMM distribution is

$$p_{L_{\mathbf{X}}}(l_{\mathbf{X}}) = \sum_{k=1}^{M} \mathbf{w}_k \, p_{\mathcal{N}}(l_{\mathbf{X}}|m_k, \sigma_k) \, (\mathrm{dB})$$
(5.3)

where l_x is measured in dB and $p_{\mathcal{N}}(.)$ represents the PDF of Gaussian distribution with parameters (m_k, σ_k) . The weighting parameters w_k are selected such that $\sum_{k=1}^{M} w_k = 1$. Expectation Maximization (EM) algorithm is used to find the proper parameters to maximize the goodness of the fit [9]. Moreover, after some basic manipulation, the PDF in (5.3) can be transformed to linear which results in lognormal mixture distribution

$$p_{\omega}(\omega) = \sum_{k=1}^{M} w_k \frac{\xi}{\sqrt{2\pi\omega\sigma_k}} \exp\left(-\frac{(10\log(\omega) - m_k)^2}{2\sigma_k^2}\right)$$
(5.4)

where $\xi = \frac{10}{\ln 10}$ and ω is the received average power normalized by the free space path loss $L_{\rm FS}$.

From network planning perspectives, using the developed model in (5.2) enables the network designers to ignore the house structure and abstract the FBS inside the house by a transmitter with a random gain in each direction in far field. The notion illustrated in Fig. 5.3 shows the received signal power for a sample floor plan.

It is worth noting that the random floor plan is not the sole source for the random behavior of the signal strength at the receiver side, multipath fading caused by the indoor reflections of the FBS signal may also change the instantaneous signal strength substantially. Consequently, the received signal will experience a composite fading/shadowing effects. Therefore, the PDF of the received signal envelope in the composite model is [10],

$$p_r(r) = \int_0^\infty p_{r|\omega}(r|\omega) p_\omega(\omega) \ d\omega.$$
(5.5)

Note that $p_{r|\omega}(.)$ describes the received signal envelope given that the average received power is ω .



Figure 5.3: A sample FBS propagation contour.

Several measurements [11, 12] indicate that indoor fading distribution can be modeled by a Rician PDF. Therefore, it is reasonable to assume such distribution for a femtocell signal group. Therefore, the conditional PDF of the signal envelope can be written as

$$p_{r|\omega}(r|\omega) = \frac{2(K+1)r}{\omega} \exp\left(-K - \frac{(K+1)r^2}{\omega}\right) I_0\left(2\sqrt{\frac{K(K+1)}{\omega}}r\right) \triangleq p_{\mathcal{R}}(r, K, \omega)$$
(5.6)

where K is the Rician K-factor and I_0 is the 0th order modified Bessel function of the first kind. Considering the complete model of shadowing and Rician fading, and substituting (5.4) and (5.6) in (5.5) yields a mixture of composite Rician/lognormal Chapter 5: Indoor-Outdoor Channel Characterization and its Application to Interference Analyses of Femtocells 102

distributions,

$$p_r(r) = \sum_{k=1}^M w_k \int_0^\infty p_{\mathcal{R}}(r, K, \omega) \frac{\xi}{\omega\sqrt{2\pi\sigma_k}} \exp\left(-\frac{(10\log(\omega) - m_k)^2}{2\sigma_k^2}\right) d\omega.$$
(5.7)

In general, closed-form solutions for composite fading/shadowing models are not available in the literature [13], particularly when the conditional PDF of the received signal amplitude involves complicated distributions such as the Rician mixture PDF described in (5.7). Consequently, numerical solutions or Monte Carlo simulation can be invoked to evaluate the system performance. Alternatively, the model can be slightly relaxed to allow for a relatively simpler model that can be solved analytically, as it is described in the next subsection.

5.2.1 A simplified stochastic attenuation model

The developed model in the previous section can be used to accurately describe the stochastic behavior of the femtocell signal propagating from the indoor transmitter to the outdoor environment. However, since the main objective of the femtocell deployment is to cover the interior part of the building itself, it will be reasonable to assume that the FBS transmitter location is fixed in the center of the building. Furthermore, the contribution of the doors and windows to the overall attenuation is much smaller than the contribution of the internal and external walls, hence the attenuation of the doors and windows can be ignored with no major impact on the model accuracy. Consequently, L_x is reduced to

$$L_{\rm x} = \sum_{i=1}^{W} k_i \alpha_i \quad (dB) \tag{5.8}$$

where $k_1 \in \{0, 1\}$ and $k_2 \ge 0$ are random variables that represent the number of outer and interior walls crossed, respectively. While k_1 can be modeled as a Bernoulli random variable with probability $p(k_1 = 0)$ [9], the simulation results have demonstrated that k_2 can be approximated by a Poisson distribution with a mean value



Figure 5.4: The simplified PDF of the building shadowing $L_{\rm x}$ given in (5.8).

 $\lambda = 1.2,$

$$p_{k_2}(k_2) = \frac{\lambda^{k_2}}{k_2!} e^{-\lambda} \,. \tag{5.9}$$

Fig. 5.5 and 5.4 present the PDF of the number of internal walls k_2 and the simplified building shadowing L_x , respectively. As it can be noted from Fig. 5.4, the shadowing is essentially the summation of M shifted Dirac functions scaled by the factor w_k , hence it can be written as

$$p_{L_{\mathbf{x}}}(l_{\mathbf{x}}) = \sum_{k=1}^{M} \mathbf{w}_k \delta(l_{\mathbf{x}} - l_k) \quad (dB)$$
 (5.10)

where $\delta(l_{\rm x} - l_k) = 1$ for $l_{\rm x} = l_k$ and 0 otherwise. By a simple transformation, the





Figure 5.5: Probability mass function of the number of penetrated internal walls k_2 .

PDF in (5.10) in linear can be expressed as

$$p_{\omega}(\omega) = \sum_{k=1}^{M} \mathbf{w}_k \delta(\omega - \omega_k)$$
(5.11)

where $\omega_k \triangleq 10^{\frac{l_k}{10}}$.

Finally, the PDF of the received signal amplitude can be achieved by substituting (5.11) into (5.5), which gives the following composite fading/shadowing model,

$$p_r(r) = \sum_{k=1}^{M} \mathbf{w}_k \ p_{\mathcal{R}}(r, K, \omega_k).$$
(5.12)

This model can be applied to other fading models where the Rician fading can be replaced by other models such as Nakagami, or Rayleigh.

5.3 Received Signal Model

This work focuses on studying the adverse effect of installing FBS in a preplanned legacy cellular network. Specifically, the effects of FBSs' downlink signal on the downlink reception of a macro user outside the house is studied. All of the FBSs and the macro base stations (MBS) use OFDM in their downlink transmission, due to its unique performance in fading environments.

5.3.1 OFDM signal model

The OFDM symbol duration $T_{\rm s}$ is composed of two parts, $T_{\rm s} = T_{\rm u} + T_{\rm CP}$. The useful data transmission part is denoted by $T_{\rm u}$, whereas $T_{\rm CP}$ represents the cyclic prefix added to the symbol to preserve orthogonality and remove inter-symbol interference (ISI) in delay-dispersed channels. The bandpass representation of an OFDM signal with $N_{\rm c}$ subcarriers during the *l*th signaling period is given by [14],

$$s(t) = \sum_{l=-\infty}^{\infty} \sum_{k=0}^{N_{\rm c}-1} d_k(l) \cos[2\pi (f_{\rm c} + f_k)t_l + \varphi_k(l)]$$
(5.13)

where $d_k(l) = |d'_k(l)|$, $d'_k(l)$ is the complex data symbol that modulates the kth subcarrier at frequency f_k , $\varphi_k(l) = \arg(d'_k(l))$, f_c represents the carrier frequency and $t_l = t - lT_s$.

Assuming that the signal passes through a multipath channel with L discrete components each of which has a gain $h_i(t)$ and delay $\tau_i(t)$, then the received signal can be expressed as

$$r(t) = \sum_{l=-\infty}^{\infty} \sum_{i=1}^{L} \sum_{k=0}^{N_{\rm c}-1} h_i(l) d_k(l) \cos[2\pi f_{{\rm c},k}(t_l - \tau_i(l)) + \varphi_k(l)] + n(t)$$
(5.14)

where n(t) represents a zero-mean white Gaussian noise process with power spectral density of $\frac{N_0}{2}$ W/Hz. By considering slow fading channel, it is reasonable to assume that channel gain and delay remain fixed for the period of one symbol. Moreover, assuming that $T_{\rm CP} > \max{\{\tau_i, i = 1, ..., L\}}$, i.e., no intersymbol interference, it is possible to simplify the received signal to

$$r(t) = \sum_{i=1}^{L} \sum_{k=0}^{N_{\rm c}-1} h_i(l) d_k(l) \cos[2\pi f_{{\rm c},k} t_l + \phi_k(i,l)] + n(t), \quad t \in [lT_{\rm s}, (l+1)T_{\rm s}].$$
(5.15)

Note that $\phi_k(i,l) \triangleq \varphi_k(l) - 2\pi f_{c,k}\tau_i$. It is reasonable to assume that $\phi_k(i,l)$ are independent and identically distributed random variables with uniform distribution in $(0, 2\pi]$. This is due the fact that although τ_i might not be very different from each other, $1/f_{c,k}$ is very small and hence, any small change in the delays can change the phase by 2π .

The channel gains are typically assumed to be Gaussian. Thus, it is possible to view (5.15) as the summation of Gaussian vectors with uniform phases. A key observation is that if h_i are i.i.d. and circularly invariant complex Gaussian variates with zero mean and variance $\mu_{\rm hh} = \frac{1}{2}E\{h_i^2\}$, the summation amplitude is Nakagamim distributed with $\Omega = 2\mu_{\rm hh}$ and fading parameter m = L, which is assumed to be integer in this model [10]. Thus, we can write (5.15) as

$$r(t) = \sum_{k=0}^{N_{\rm c}-1} d_k(l) r_k(l) \cos[2\pi f_{{\rm c},k} t_l + \psi_k(l)] + n(t), \quad t \in [lT_{\rm s}, (l+1)T_{\rm s}]$$
(5.16)

where $r_k(l) \triangleq \sqrt{\sum_{i=1}^{L} |h_i \exp(\phi_k(i, l))|^2}$ represents the signal amplitude and $\psi_k(l)$ represents the resulting phase, which is uniformly distributed over $(0, 2\pi]$.

5.3.2 Macrocell/Femtocell signal model

It can be directly concluded from (5.16) that the received OFDM signal can be decomposed into N_c signals distributed over N_c subcarriers. Therefore, for the remaining part of the chapter, we focus on the signal on each subcarrier, and therefore the subcarrier index k is dropped from all subsequent equations. Moreover, the macrocell macrocell user is considered as the desired user while the femtocell users are considered as interferers. Assuming a Nakagami-m fading channel, the macro user signal is given by

$$s_{\rm m}(t) = \sqrt{P_{\rm m}} \varrho(l) d_{\rm m}(l) \cos(2\pi f_{\rm c} t_l + \psi_{\rm m}(l))$$
(5.17)

where $P_{\rm m}$ is the average received signal power, $d_{\rm m}(l)$ denotes the transmitted data symbol, $\psi_{\rm m}$ is the received signal phase, and ρ represents the fading amplitude, which follows Nakagami-m distribution with parameters (m, Ω) ,

$$p_{\varrho}(\varrho) = \frac{2m^{m}\varrho^{2m-1}}{\Omega^{m}\Gamma(m)} \exp\left(-\frac{m\varrho^{2}}{\Omega}\right), \quad \varrho \ge 0.$$
(5.18)

Here, we also assume a slow-fading channel.

The interference is composed of N FBSs that are transmitting from inside N different houses with unknown architectural details. This unknown architecture imposes large scale fading, i.e., building shadowing, on the femtocell signal that is received outdoor. Therefore, the received interference can be modeled as

$$s_{\rm f}(t) = \sum_{n=1}^{N} \sqrt{P_n} \rho_n(l) d_n(l) \cos\left[2\pi f_{\rm c}(t_l - \tau_n) + \psi_n(l)\right]$$
(5.19)

where P_n is the received average power of the n^{th} FBS, and $\rho_n(l)$ is the composite shadowing/fading amplitude gain. Based on the simplified model presented in (5.12), the PDF of ρ_n is given by

$$p_{\rho_n}(\rho_n) = \sum_{k=1}^M \mathbf{w}_k \ p_{\mathcal{R}}(\rho_n, K_n, \omega_{n,k})$$
(5.20)

where $\omega_{n,k}$ is the average power of k^{th} component of n^{th} interferer and w_k is the weight of k^{th} component. Note that subscript n is added to represent the n^{th} interferer. It is worth mentioning that we have included the free-space path loss L_{FS} in P_n , whereas the shadowing/fading amplitude gain includes the losses due to the propagation through obstacles, i.e., the building shadowing, and multipath fading as described in Section 5.2.

The reference user receives the desired signal as well as N interfering signals transmitted from N FBSs. Therefore, the received signal is a summation of femtocells and macrocell signals

$$r(t) = \sqrt{P_{\rm m}} \varrho(l) d_{\rm m}(l) \cos(2\pi f_{\rm c} t_l + \psi_{\rm m}(l)) + \sum_{n=1}^{N} \sqrt{P_n} \rho_n(l) d_n(l) \cos\left[2\pi f_{\rm c}(t_l - \tau_n) + \psi_n(l)\right] + n(t). \quad (5.21)$$

The received signals' phases $\psi_{\rm m}(l)$ and $\psi_n(l)$ are assumed to be mutually independent and uniformly distributed over $[0, 2\pi)$. In order to consider the worst-case scenario, it is assumed that the macro user is in the proximity of the houses where the FBS are deployed.

5.4 Performance Analysis of a Macro User in the Presence of Femtocells

5.4.1 Femtocell signal approximation using Nakagami-m distribution

Based on the envelope distribution of the femto signal given in (5.20), the performance analysis should involve the Rician PDF. However, manipulating the Rician PDF can be highly complicated due to the Bessel function part of the PDF. Therefore, we adopt the approximation of Rician distribution by Nakagami-m PDF [15]. Furthermore, several experimental results show that the Nakagami distribution can be adopted for characterizing the indoor channel fading at particular frequencies [11]. Therefore, given a random variable $X \sim \mathcal{R}(r, K, \omega)$, then X can be approximated by

$$X \simeq \mathcal{K}(r, \mu, \omega) \tag{5.22}$$

Chapter 5: Indoor-Outdoor Channel Characterization and its Application to Interference Analyses of Femtocells 109

where \mathcal{K} is the Nakagami-m distribution with parameters ω and $\mu = \frac{(1+K)^2}{1+2K}$. Applying the approximation to (5.20) gives

$$p_{\rho_n}(\rho_n) \simeq \sum_k w_k p_{\mathcal{K}}(\rho_n | \mu_n, \omega_{n,k})$$
$$= \sum_k w_k \frac{2\mu_n^{\mu_n}}{\Gamma(\mu_n)\omega_{n,k}^{\mu_n}} \rho_n^{2\mu_n - 1} \exp\left(-\frac{\mu_n}{\omega_{n,k}}\rho_n^2\right)$$
(5.23)

where $p_{\mathcal{K}}$ is the Nakagami PDF and μ_n is related to Rician K_n as described under (5.22). Therefore, the approximation can be explained as a Nakagami-Mixture.

5.4.2 PDF of the received signal envelope

This subsection describes the derivation of the received signal envelope PDF for the single interferer case, the same approach can be applied to extend the results to multiple interferers. The PDF of the received signal envelope is the vector sum of the femtocell and macrocell signals. Given that the phases of the two signals are independent and uniformly distributed, the PDF of the sum of two vectors is [16]

$$f_{B_2}(b_2) = \int_0^\infty \int_{M_1}^{M_u} \frac{2b_2}{\pi\sqrt{4A_1^2A_2^2 - (b_2^2 - A_1^2 - A_2^2)^2}} p_{A_1}(A_1) \ p_{A_2}(A_2) \ dA_1 \ dA_2 \quad (5.24)$$

where A_i and p_{A_i} , i = 1, 2 are the *i*th signal envelope and its corresponding PDF, respectively. The lower and upper limits of the envelope are represented by $M_1 =$ $||A_1| - |A_2||$ and $M_u = |A_1| + |A_2|$, respectively. Plugging (5.18) and (5.23) into (5.24) gives

$$p_{r}(r) = \int_{0}^{\infty} \int_{|\varrho-\rho|}^{\varrho+\rho} \frac{2r}{\pi\sqrt{4\rho^{2}\rho^{2} - (r^{2} - \varrho^{2} - \rho^{2})^{2}}} \frac{2m^{m}}{\Gamma(m)\Omega^{m}} \varrho^{2m-1} \exp\left(-\frac{m}{\Omega}\varrho^{2}\right) \\ \times \sum_{k} w_{k} \frac{2\mu^{\mu}}{\Gamma(\mu)\omega_{k}^{\mu}} \rho^{2\mu-1} \exp\left(-\frac{\mu}{\omega_{k}}\rho^{2}\right) d\varrho d\rho.$$
(5.25)

By extracting the summation from the integral, it can be rewritten as

$$p_r(r) = \sum_k w_k \int_0^\infty \int_{|\varrho-\rho|}^{\varrho+\rho} \frac{2r}{\pi\sqrt{4\rho^2 \varrho^2 - (r^2 - \varrho^2 - \rho^2)^2}} \times \frac{2m^m}{\Gamma(m)\Omega^m} \varrho^{2m-1} \exp\left(-\frac{m}{\Omega} \varrho^2\right) \frac{2\mu^\mu}{\Gamma(\mu)\omega_k^\mu} \rho^{2\mu-1} \exp(-\frac{\mu}{\omega_k} \rho^2) \, d\varrho \, d\rho. \quad (5.26)$$

The double integral part of (5.26) represents the summation of two phase vectors with Nakagami distributed amplitude and uniform phase $[R = X \exp(j\phi_x) + Y \exp(j\phi_y)]$. The two Nakagami components are $X \sim \mathcal{K}(m, \Omega)$ and $Y \sim \mathcal{K}(\mu, \omega_k)$.

A closed-form solution for the addition of L Nakagami vectors with uniform phases and integer m is given in [17]. The PDF of the summation envelope is worked out to be

$$f_{|H|}(r) = \mathcal{G}_L \left\langle \frac{r^{2j+1} e^{-\frac{r^2}{4U_L}}}{U_L^{j+1} 2^{2j+1}} \right\rangle$$
(5.27)

where the operator $\mathcal{G}_L \langle . \rangle$ is defined as

$$\mathcal{G}_L \langle \chi \rangle \triangleq \sum_{i_1=0}^{m_1-1} \dots \sum_{i_L=0}^{m_L-1} \sum_{j=0}^{S_L} \frac{\prod_{l=1}^L \left(\frac{(1-m_l)_{i_l}}{(i_l!)^2}\right) (-S_L)_j \ S_L! Y_L}{(j!)^2 U_L^{S_L}} \chi$$
(5.28)

where L is the number of Nakagami signals, m_i and Ω_i , i = 1, ..., L are the Nakagami parameters, and the pochhammer symbol is denoted by $(z)_n$. The coefficients in (5.28) are related to the parameters of the individual Nakagami distributions as follows,

$$S_L = \sum_{l=1}^{L} i_l, \qquad Y_L = \prod_{l=1}^{L} \left(\frac{\Omega_l}{4m_l}\right)^{i_l}, \qquad U_L = \sum_{l=1}^{L} \frac{\Omega_l}{4m_l}.$$

Substituting (5.27) into (5.26) gives the envelope PDF in closed-form

$$p_r(r) = \sum_k w_k \mathcal{G}_{2,k} \left\langle \frac{r^{2j+1} e^{-\frac{r^2}{4U_{2,k}}}}{U_{2,k}^{j+1} 2^{2j+1}} \right\rangle.$$
(5.29)



Figure 5.6: Analytical and simulated PDF of the received signal envelope given in (5.26) for different received mean power of macrocell and femtocell.

Note that the Nakagami-related coefficients are different for each component of the mixture and should be calculated accordingly, which is emphasized by adding the subscript k to the operator \mathcal{G} and to the coefficients. Fig. 5.6 presents the PDF of the received signal envelope for different values of femtocell and macrocell average received power. As it can be seen from the figure, the simulation and analytical results are almost identical.

5.4.3 Signal to interference ratio (SIR)

In order to analyze the performance of the interference-limited wireless network, it is vital to have an analytical model for the signal to interference ratio (SIR), which is denoted as η and defined as

$$\eta = \frac{\gamma_{\rm m}}{\sum_n \gamma_n} \tag{5.30}$$

where $\gamma_{\rm m}$ is the instantaneous macrocell signal power and γ_n is the $n^{\rm th}$ interference power. Note that the interferences are added incoherently. Assuming that SNR $\gg 1$, the system performance will be dominated by the interference while the effect of the AWGN noise can be ignored. The instantaneous power of a Nakagami faded signal, such as the macrocell signal considered in this work, is known to be gamma distributed [10],

$$p_{\gamma_{\rm m}}(\gamma_{\rm m}|m,\Omega) = \frac{m^m}{\Gamma(m)\Omega^m} \gamma_{\rm m}^{m-1} \exp\left(-\frac{m}{\Omega}\gamma_{\rm m}\right), \quad \gamma_{\rm m} \ge 0.$$
(5.31)

Similarly, given that the PDF of the femtocell envelope is given by (5.23), the PDF of instantaneous power can be written as

$$p\gamma_n(\gamma_n|\mu_n, [\omega_{n,k}, k=1, ..., M]) = \sum_k w_k \frac{\mu_n^{\mu_n}}{\Gamma(\mu_n)\omega_{n,k}^{\mu_n}} \gamma_n^{\mu_n - 1} \exp\left(-\frac{\mu_n}{\omega_{n,k}}\gamma_n\right).$$
(5.32)

5.4.3.1 Single interferer

Given that the interference is composed of a single FBS signal, the SIR definition is reduced to $\eta = \frac{\gamma_{\rm m}}{\gamma_1}$. Substituting (5.31) and (5.32) into the ratio distribution equation $(p_X(z) = \int_{-\infty}^{\infty} |y| \ p_{X,Y}(zy, y) \ dy)$ gives

$$p_{\eta}(\eta) = \frac{m^m}{\Gamma(m)\Omega^m} \eta^{m-1} \sum_k w_k \frac{\mu^{\mu}}{\Gamma(\mu)\omega_k^{\mu}} \int_0^\infty y^{m+\mu-1} \exp\left(-\left(\frac{m}{\Omega}\eta + \frac{\mu}{\omega_k}\right)y\right) \, dy.$$

Using the standard integral [18, 3.381.4] $(\int_0^\infty x^{\nu-1} \exp(-\mu x) dx = \mu^{-\nu} \Gamma(\nu)$, for $\Re\{\mu\} > 0$, $\Re\{\nu\} > 0$), where $\Re\{x\}$ denotes the real part of x, the PDF of η is calculated as

$$p_{\eta}(\eta) = \frac{\Gamma(m+\mu)}{\Gamma(m)\Gamma(\mu)} \vartheta^{m} \eta^{m-1} \times \sum_{k} w_{k} \, \theta_{k}^{\mu} \, (\vartheta\eta + \theta_{k})^{-m-\mu}$$
(5.33)

where $\theta_k = \frac{\mu}{\omega_k}$ and $\vartheta = \frac{m}{\Omega}$.

5.4.3.2 Multiple interferers

In the presence of multiple interfering signals, it is typically assumed that the interfering signals are added non-coherently, i.e., the total interference power is computed as the sum of the individual interfering signals power. Thus, the distribution of total power is required to achieve a closed-form formula for the SIR. The characteristic function of the total interference power is defined as

$$\Phi_{\rm I}(t) = \prod_n \int_0^\infty p_{\gamma_n}(s) \exp(i\,t\,s) \,ds \tag{5.34}$$

where $i = \sqrt{-1}$. Substituting $p_{\gamma_n}(s)$ by the PDF described by (5.32) and solving the integral yields

$$\Phi_{\mathbf{I}}(t) = \prod_{n} \sum_{k} w_k \left(1 - \frac{it}{\theta_{n,k}} \right)^{-\mu_n}$$
(5.35)

where $\theta_{n,k} = \mu_n / \omega_{n,k}$. After a few simple algebraic manipulations $\Phi_{I}(t)$ can be written as

$$\Phi_{\mathrm{I}}(t) = \underbrace{\sum_{k_1=1}^{M} \dots \sum_{k_N=1}^{M}}_{N} \prod_{n=1}^{N} w_k \left(1 - \frac{it}{\theta_{n,k_n}}\right)^{-\mu_n}$$
(5.36)

where M is the number of coefficients used to model a femtocell signal and N is the number of interfering femtocell signals. Therefore, the PDF of the total power is the inverse Fourier transform

$$p_{\mathbf{I}}(s) = \frac{1}{2\pi} \sum_{k_1=1}^{M} \dots \sum_{k_N=1}^{M} \prod_{n=1}^{N} w_k \int_0^\infty \prod_{n=1}^{N} \left(1 - \frac{it}{\theta_{n,k_n}}\right)^{-\mu_n} \times \exp(-its) \ dt.$$
(5.37)

The integral in (5.37) represents the sum of N Gamma distributed variates. According to [19], given the variates $X_n \sim \mathcal{Q}(\alpha_n, \beta_n)$, where \mathcal{Q} is the Gamma distribution with

Chapter 5: Indoor-Outdoor Channel Characterization and its Application to Interference Analyses of Femtocells 114

a given parameters, the PDF of $Y = \sum_{n=1}^{N} X_n$ can be expressed as

$$p_Y(y) = U \sum_{k=0}^{\infty} \frac{\delta_k y^{M+k-1} \exp\left(-\frac{y}{\beta_{\min}}\right)}{\beta_{\min}^{M+k} \Gamma(V+k)}, \quad y \ge 0$$
(5.38)

where $\beta_{\min} = \min\{\beta_n\}$, $U = \prod_{n=1}^N \left(\frac{\beta_{\min}}{\beta_n}\right)^{\alpha_n}$ and $V = \sum_{n=1}^N \alpha_n$, and the coefficient $\delta_k = 1$ for k = 0, and otherwise it can be computed recursively

$$\delta_k = \frac{1}{k} \sum_{i=1}^k \left[\sum_{j=1}^N \alpha_j \left(1 - \frac{\beta_{\min}}{\beta_j} \right)^i \right] \delta_{k-i}, \quad k = 1, 2, \dots$$
(5.39)

Applying this formula, a closed-form expression for the PDF of the total interference power can be obtained

$$p_{\mathrm{I}}(s) = \sum_{k_1=1}^{M} \dots \sum_{k_N=1}^{M} U_{\vec{K}} \sum_{l=0}^{\infty} \frac{\delta_l s^{V+l-1} \exp\left(-\theta_{\vec{K},\max} \; y\right)}{\theta_{\vec{K},\max}^{-V-l} \; \Gamma(V+l)}$$
(5.40)

where $\theta_{\vec{K},\max} = \max_{n} \{\theta_{n,k_n}\}, U_{\vec{K}} = \prod_{n=1}^{N} \omega_{k_n} \left(\frac{\theta_{n,k_n}}{\theta_{\vec{K},\max}}\right)^{\mu_n}, V = \sum_{n=1}^{N} \mu_n$ and

$$\delta_{\vec{K},l} = \frac{1}{l} \sum_{i=1}^{l} \left[\sum_{j=1}^{N} \mu_j \left(1 - \frac{\theta_{j,k_j}}{\theta_{\vec{K},\max}} \right)^i \right] \delta_{l-i}, \quad l = 1, 2, \dots$$
(5.41)

Using the distribution of the total interference power, the same approach used in Section 5.4.3.1 can be applied to derive the SIR distribution. Thus, the PDF in (5.40) is inserted in the ratio distribution equation,

$$p_{\eta}(\eta) = \frac{\vartheta^{m}}{\Gamma(m)} \eta^{m-1} \sum_{k_{1}=1}^{M} \dots \sum_{k_{N}=1}^{M} U_{\vec{K}} \sum_{l=0}^{\infty} \int_{0}^{\infty} \frac{\delta_{\vec{K},l}}{\theta_{\vec{K},\max}^{-M-l} \Gamma(V+l)} y^{\mu_{m}+M+l-1} \times \exp\left(-\left(\vartheta\eta + \theta_{\vec{K},\max}\right)y\right) dy \quad (5.42)$$

Chapter 5: Indoor-Outdoor Channel Characterization and its Application to Interference Analyses of Femtocells 115

Using the standard integral [18, 3.381.4] yields

$$p_{\eta}(\eta) = \frac{\vartheta^m}{\Gamma(m)} \eta^{m-1} \sum_{k_1=1}^M \dots \sum_{k_N=1}^M U_{\vec{K}} \sum_{l=0}^\infty \frac{\delta_{\vec{K},l}}{\theta_{\vec{K},\max}^m} (V+l)^{(m)} \left(\frac{\vartheta}{\theta_{\vec{K},\max}}\eta + 1\right)^{-(m+V+l)}$$
(5.43)

where $x^{(n)}$ is the rising factorial. The PDF can be represented in terms of \mathcal{H} operator defined as

$$\mathcal{H}\langle\chi\rangle \triangleq \frac{\vartheta^m}{\Gamma(m)} \sum_{k_1=1}^M \dots \sum_{k_N=1}^M U_{\vec{K}} \sum_{l=0}^\infty \frac{\delta_{\vec{K},l}}{\theta_{\vec{K},\max}^m} (V+l)^{(m)} \chi \tag{5.44}$$

where χ is a mathematical expression that may depend on \overrightarrow{K} or l. Note that the \mathcal{H} operator is defined using the same approach as \mathcal{G} operator and bears the same basic properties [17]. Using \mathcal{H} operator (5.43) can be rewritten as

$$p_{\eta}(\eta) = \eta^{m-1} \mathcal{H}\left\langle \left(\frac{\vartheta}{\theta_{\vec{K},\max}} \eta + 1\right)^{-(m+V+l)} \right\rangle.$$
(5.45)

Tran and Sesay [20] have investigated the truncation error of the last summation of the \mathcal{H} operator. They showed that the truncation error for sum of Gamma variates is

$$E(\ell) = 1 - U \sum_{l=0}^{\ell} \delta_l$$
 (5.46)

where E(.) is the truncation error and ℓ is the upper limit of the last summation. Therefore, for predetermined error threshold ε , $\exists \ell E(\ell) < \varepsilon$.

In this work, the inevitable truncation error of numerical analysis can be kept less than a predetermined threshold by controlling the truncation error of each summation so that $\frac{\vartheta^m}{\Gamma(m)} \sum_{k_1=1}^M \dots \sum_{k_N=1}^M E(\ell)|_{\overrightarrow{K}} < \varepsilon$. It is worth noting that (5.44) reduces back to (5.33) for the case of single interferer by simple algebraic manipulations.

5.4.4 Outage probability

The outage probability is a useful measure to evaluate the performance of interferencelimited wireless networks. In such scenarios, the outage probability is defined as the probability that the SIR becomes less than a particular value $\eta_{\rm th}$, namely, the power protection ratio. Thus the outage probability can be defined as,

$$P_{\rm out} = \Pr(\eta < \eta_{\rm th}). \tag{5.47}$$

For the case of single interferer, the outage probability can be found directly by integrating (5.33). For multiple interferers, we apply the same approach and calculate the CDF of SIR,

$$P_{\eta}(\eta) = \int_{0}^{\eta} x^{m-1} \mathcal{H}\left\langle \left(\frac{\vartheta}{\theta_{\vec{K},\max}} x + 1\right)^{-(m+V+l)} \right\rangle \, dx.$$
 (5.48)

The \mathcal{H} operator is defined by a number of linear operations and has the properties $\alpha \mathcal{H} \langle \chi \rangle = \mathcal{H} \langle \alpha \chi \rangle$ and $\int_a^b \mathcal{H} \langle \chi \rangle = \mathcal{H} \left\langle \int_a^b \chi \right\rangle$. Therefore,

$$P_{\eta}(\eta) = \mathcal{H}\left\langle \int_{0}^{\eta} x^{m-1} \left(\frac{\vartheta}{\theta_{\vec{K},\max}} x + 1 \right)^{-(m+V+l)} dx \right\rangle.$$
(5.49)

The integral in (5.49) can be solved with the help of the following standard integral [18, 2.111.2]

$$\int \frac{x^n}{z^m} dx = \frac{x^n}{z^{m-1}(n+1-m)b} - \frac{na}{(n+1-m)b} \int \frac{x^{n-1}}{z^m} dx$$
(5.50)

where z = a + bx. Given that n is integer, it is possible to get a direct closed-form solution with some algebraic manipulations

$$\int \frac{x^n}{z^m} \,\mathrm{d}\, x = \sum_{i=0}^n c_i f_i(x) \tag{5.51}$$

where

$$c_{i} = \prod_{j=i+1}^{n} \left(-\frac{ja}{b(j+1-m)} \right)$$
(5.52)

$$f_i(x) = \frac{z^{1-m}}{b(i+1-m)} x^i.$$
(5.53)

Substituting (5.51), (5.52) and (5.53) in (5.49), it is possible to calculate the outage probability. Given that m is integer,

$$P_{\eta}(\eta) = \mathcal{H}\left\langle \sum_{i=0}^{m-1} c_i \times f_i(\eta) - c_0 f_0(0) \right\rangle$$
(5.54)

where

$$c_i = \prod_{j=i+1}^{m-1} \left(-\frac{j\theta_{\vec{K},\max}}{\vartheta(j+1-m-V-l)} \right)$$
(5.55)

$$f_i(\eta) = \frac{\theta_{\vec{K},\max} \left(\frac{\vartheta}{\theta_{\vec{K},\max}} \eta + 1\right)^{1-m-V-l}}{\vartheta(i+1-m-V-l)} \eta^i.$$
(5.56)

5.5 Numerical Results

This section presents analytical and simulation results based on the developed interference model where the macrocells signal is the desired signal and the femtocell signals are the interferers. The OFDM system considered has $N_c = 128$ subcarriers and 16 cyclic prefix samples. The simulation results are obtained based on the assumption that the macrocell signal is Nakagami faded, while the fading/shadowing of the femtocell signal follows the composite shadowing/multipath fading model described in (5.20). Because the macro user is in the proximity of the FBS, the channels for both the macro and femto signals are assumed to have $\tau_i \ll T_{CP}$, i = 1,..., L. Unless stated otherwise, the simulation parameters are $\eta_{\text{th}} = 10$ dB, m = 8, and the femtocell signal is generated according to the developed model in Section II.C, and the Rician factor K = 7 dB. The analytical results are based on the approximated



Chapter 5: Indoor-Outdoor Channel Characterization and its Application to Interference Analyses of Femtocells 118

Figure 5.7: The PDF of femtocell signal envelope described in (5.23) for different values of K.

model described in Section IV.A.

The analytical PDF of a FBS signal envelope constructed according to (5.29) for K = 4, 7, 12 and 15 dB is depicted in Fig. 5.7. It can be noted from the figure that the femto signal PDF deviates from any well documented distribution such as the Nakagami, particularly for the case of low SIR values. Consequently, adopting the Nakagami distribution for the envelope of the interfering signal may cause significant error.

Fig. 5.8 compares the SIR and SINR distributions in the presence of one and two FBSs. The transmission power of the FBSs is adjusted such that the total received interference would be equal in both cases. The figure shows that the two FBSs increase the possibility of low SIR and SINR. It also reveals that the effect of noise is much less when two sources of interference exist. In both cases, the difference between SIR and SINR is insignificant for higher signal strength levels. Hence, it is



Figure 5.8: Comparing the PDF of SIR versus SINR for a single interferer ($\omega = 0.2$) as well as two interferers ($\omega_1 = \omega_2 = 0.1$), and $\Omega = 1$.

expected that the effect of noise can be ignored in those regions.

The outage probability contour map for a scenario with three FBSs is illustrated in Fig. 5.9. The FBSs are represented by a cross and the MBS is at (0, 300). For simplicity, the map ignores the effect of other urban structure around the houses. The figure demonstrates the high probability of outage in proximity of the houses. It is mainly due to the fact that line of sight reception of FBS signal through windows are considered in the proposed model. It also shows the significance of the FBS effect in further distances from the MBS, especially when adaptive femtocell transmission power is not employed. Therefore, this figure clearly indicates the need to scrutinize the cost and benefit of each individual FBS implementation.

The outage probabilities for single and multiple femtocells are compared in Fig. 5.10. It can be noted that increasing the number of femtocells increases the outage probability even when the total received interference power is maintained



Figure 5.9: An outage probability map around three FBSs in a given scenario, calculated by the proposed model. The MBS is at (0, 300).

fixed. Such results contradicts the intuitive results that increasing the number of femtocell interferers does not affect the outage probability given that the total received interference power is kept fixed. Furthermore, it can be noted that the simulation results matches the analysis very well.

The BER performance of a macro user versus SNR in the presence of a single FBS is given in Figure 5.11. The BER is obtained using the developed PBGA-based model presented in Section II.C, and it is compared to the conventional model which assumes that the FBS is installed at a random location inside an outer room (OR) that faces the macro user. The FBS signal envelope PDF of the conventional model is considered as a single Nakagami-m distribution. The results for all cases are generated assuming that the FBS transmitter power is fixed at $P_{tx} = 15$ or 20 dBm, and the MBS transmit power is 43 dBm. The worst case scenario where the FBS is in the line-of-sight (LOS) of the macro user is considered as well. The macro user is located



Figure 5.10: Analytical and simulated outage probability for single and multiple FBSs.

at a distance of 300 m from the MBS and 35 m from the FBS. As it can be noted from the figure, the outer room assumption leads to pessimistic results that are close to the LOS case. Although the results are not included, it can be directly concluded from the figure that the BER results would be highly optimistic if we assume that the FBS is installed in an internal room, which is due to the additional ~10 dB attenuation.

5.6 Chapter Summary

This work presented a novel composite shadowing/fading model that captures the complex effects of transmitting from inside a suburban house as in the case of femtocell deployment. The presented work can be divided into two parts. The first part considered the development of a stochastic shadowing model for indoor-to-outdoor channels that considers the complex effects of different buildings' floor plans. The



Figure 5.11: Simulated BER versus SNR in the presence of single FBS using the developed model, outer room model and LOS.

second part utilized the proposed model to evaluate the performance of a legacy network user in the proximity of one or more houses where FBS are deployed. The system performance was evaluated in terms of the outage probability and the signal to interference ratio, where a closed-form solutions were derived for several cases of interest. The analytical and simulation results demonstrated that using the proposed model will lead to more realistic results as compared to other restricted models that may lead to highly optimistic or pessimistic results.

References

- M. Mirahmadi, A. Al-Dweik, and A. Shami, "Indoor-outdoor channel characterization and its application to interference analyses of femtocells," *IEEE Trans. Wireless Commun.*, submitted for publication.
- [2] Technical Specification Group Radio Access Networks (TSG-RAN), "3G Home NodeB study item technical report (release 8)," 3GPP, Tech. Rep. TR 25.820, 2008.
- [3] —, "HNB and HNB-macro propagation models," 3GPP, Tech. Rep. WG4 #44-bis, 2007.
- [4] Y. Kim, S. Lee, and D. Hong, "Performance analysis of two-tier femtocell networks with outage constraints," *IEEE Trans. Wireless Commun.*, vol. 9, no. 9, pp. 2695–2700, Sep. 2010.
- [5] J. Y. Lee, S. J. Bae, Y. M. Kwon, and M. Y. Chung, "Interference analysis for femtocell deployment in OFDMA systems based on fractional frequency reuse," *IEEE Commun. Lett.*, vol. 15, no. 4, pp. 425–427, Apr. 2011.
- [6] V. Chandrasekhar and J. Andrews, "Uplink capacity and interference avoidance for two-tier femtocell networks," *IEEE Trans. Wireless Commun.*, vol. 8, no. 7, pp. 3498–3509, Jul. 2009.
- [7] S. Park, W. Seo, Y. Kim, S. Lim, and D. Hong, "Beam subset selection strategy for interference reduction in two-tier femtocell networks," *IEEE Trans. Wireless Commun.*, vol. 9, no. 11, pp. 3440–3449, Nov. 2010.
- [8] A. Valcarce and J. Zhang, "Empirical indoor-to-outdoor propagation model for residential areas at 0.9–3.5 ghz," *IEEE Antennas Wireless Propag. Lett.*, vol. 9, pp. 682–685, 2010.

- [9] M. Mirahmadi, A. Shami, and A. Al-Dweik, "A building architecture model for predicting femtocell interference in next-generation networks," in *IEEE Int. Conf. Commun. (ICC)*, Ottawa, Canada, Jun. 2012, pp. 5059–5063.
- [10] M. K. Simon and M.-S. Alouini, Digital communication over fading channels. Wiley-Interscience, 2005, vol. 95.
- [11] Y. Lustmann and D. Porrat, "Indoor channel spectral statistics, K-factor and reverberation distance," *IEEE Trans. Antennas Propag.*, vol. 58, no. 11, pp. 3685–3692, Nov. 2010.
- [12] J.-J. Park, M.-D. Kim, H.-K. Chung, and W. Kim, "Ricean K-factor analysis of indoor channel measurements at 3.7 GHz," in 5th Int. ICST Conf. Commun. and Networking in China (CHINACOM), Aug. 2010, pp. 1–5.
- [13] S. Atapattu, C. Tellambura, and H. Jiang, "A mixture gamma distribution to model the SNR of wireless channels," *IEEE Trans. Wireless Commun.*, vol. 10, no. 12, pp. 4193–4203, Dec. 2011.
- [14] A. Al-Dweik, B. Sharif, and C. Tsimenidis, "Accurate BER analysis of OFDM systems over static frequency-selective multipath fading channels," *IEEE Trans. Broadcast.*, vol. 57, no. 4, pp. 895–901, Dec. 2011.
- [15] M. Nakagami, "The m-distribution-a general formula of intensity distribution of rapid fading," *Statistical Method of Radio Propagation*, 1960.
- [16] Y. Maghsoodi and A. Al-Dweik, "Error-rate analysis of FHSS networks using exact envelope characteristic functions of sums of stochastic signals," *IEEE Trans. Veh. Technol.*, vol. 57, no. 2, pp. 974–985, Mar. 2008.
- [17] G. K. Karagiannidis, "A closed-form solution for the distribution of the sum of Nakagami-m random phase vectors," *IEEE Commun. Lett.*, vol. 10, no. 12, pp. 828–830, Dec. 2006.
- [18] A. Jeffrey and D. Zwillinger, Table of integrals, series, and products, 7th ed. Academic Press, 2007.

- [19] M.-S. Alouini, A. Abdi, and M. Kaveh, "Sum of gamma variates and performance of wireless communication systems over Nakagami-fading channels," *IEEE Trans. Veh. Technol.*, vol. 50, no. 6, pp. 1471–1480, Nov. 2001.
- [20] T. Tran and A. Sesay, "Sum of arbitrarily correlated gamma variates and performance of wireless communication systems over Nakagami-m fading channels," *IET Commun.*, vol. 1, no. 6, pp. 1133–1137, Dec. 2007.

Part III

Mitigating Impulsive Noise in Frequency-Selective Channels

Chapter 6 BER Reduction of OFDM Based Broadband Communication Systems over Multipath Channels with Impulsive Noise

6.1 Preamble

Orthogonal frequency division multiplexing (OFDM) is a multicarrier modulation technique that employs orthogonal subcarriers. Due to the unique features of OFDM, it is currently adopted in several wireless communication standards such as digital audio broadcasting (DAB) [2], digital video broadcasting-terrestrial (DVB-T) [3], worldwide interoperability for microwave access (WiMAX) technologies [4] and the 4G LTE-Advanced [5].

Besides its superiority for wireless transmission, OFDM has rendered itself as the dominant modulation technique for many wired technologies and standards such as the second generation digital video transmission over cable (DVB-C2) [6], the asymmetric digital subscriber line (ADSL) [7] and home networking over power line communications (PLC) [8]. For such applications, the main advantages of using OFDM are bandwidth efficiency and reduced system complexity as a result of using the fast Fourier transform (FFT) and its inverse version (IFFT).

While the multipath propagation phenomenon is a major source of disturbance for several communication systems, impulsive noise (IN) is yet another significant

A version of this chapter has been submitted for publication in [1].

source of disturbance for particular OFDM based applications such as DVB-T [9–11], PLC [12] and ADSL systems [13]. IN is typically generated by man-made electromagnetic devices, e.g. electric motors. In industrial environments, simultaneous fading and IN deteriorate the performance of communication networks significantly. In PLC networks, IN generated on the power distribution network and the multipath fading caused by reflection of the signal from irregularities in the wire are the main sources of error. Moreover, the recent adoption of OFDM systems in optical networks where optical switching generates significant IN, adds these type of networks to the list of the OFDM systems that suffer from IN.

Unlike multipath fading, adopting OFDM for IN channels might impact the performance negatively [14], because an impulsive burst may destroy all subcarriers within the OFDM symbol due to the averaging process of the FFT. Hence, using OFDM for IN channels must be accompanied with effective IN mitigation techniques [15]. In the literature, the common IN mitigation techniques used for single carrier systems such as clipping and blanking are extended to OFDM systems [12]. Approaches that are designed specifically for OFDM systems are reported in the literature as well. For example, Abedlkefi *et al.* [16] exploited the pilots embedded within OFDM signals to detect and correct the samples contaminated by IN. Zhidkov [14] proposed a frequency domain IN cancellation technique for DVB-T systems where the IN is estimated and subtracted after the FFT.

In general, most of the aforementioned techniques suffer from limited performance improvement, particularly in heavily distributed IN channels. Moreover, most of the systems reported in the literature assume that the IN bursts have a short duration that is equal to the OFDM sample period, and hence all IN samples can be considered uncorrelated [12–17]. Although such assumption is pivotal to enable analytical performance evaluation, it does not capture the bursty nature of the IN, which is confirmed by channel measurements for various applications [13,18,19], particularly for broadband communications where the OFDM sample duration is very short [8]. For example, the OFDM symbol duration as defined by the IEEE 1901 PLC standard [8] is only 5.12 μs . Hence, even very narrow bursts can affect several consecutive OFDM samples [18], or even the entire OFDM symbol. Consequently, most of the techniques that are designed based on the assumption of uncorrelated samples might not be effective in the presence of IN bursts. In [20], Al-Dweik *et. al.* proposed a new OFDM-based system to combat IN by interleaving the time-domain samples after the IFFT. The work was then extended in [21] to include multipath fading channels. However, the results in [20] and [21] were entirely obtained via Monte Carlo simulation using the assumption of ideal IN samples detection and blanking.

This chapter presents a novel technique to combat the adverse effects of multipath propagation as well as IN for OFDM based communications systems. The new system is capable of improving the bit error rate (BER) without excessive complexity increase, or spectral efficiency reduction. Closed-form formulae are derived for the signal-to-interference and noise ratio (SINR) in frequency-selective fading channels using zero forcing (ZF) and minimum mean squared error (MMSE) equalizers. Analytical and simulation results reveal that the proposed TDI system offers a remarkable BER performance improvement and can effectively and jointly combat the IN bursts and multipath fading. It is worth noting that the BER of the TDI in frequencyselective fading channels without IN is identical to precoded OFDM systems [22], [23]. However, the TDI significantly outperforms such systems in the presence of IN.

The rest of this chapter is organized as follows. The system and channel models are presented in Section II. The proposed system model is described in Section III. Section IV presents the symbol blanking scheme suggested to mitigate the IN effects. The system performance analysis in fading channels using ZF and MMSE equalizers are provided in Sections V and VI, respectively. The numerical results are given in Section VII, and finally Section VIII concludes the chapter.

Notataion: In what follows unless otherwise specified, uppercase boldface and blackboard letters, such as **H** and **H**, will denote $N \times N$ matrices, whereas lowercase boldface letters, such as **x**, will denote row or column vectors with Nelements. Samples or data symbols from the ℓ th OFDM symbol will be denoted using lower case letters such as y. Symbols with a hat such as \hat{x} will denote the estimate of x, and symbols with breve $\check{\mathbf{x}}$ denotes an interleaved sequence. Moreover, $x \sim \mathcal{N}, \mathcal{N}_c, \mathcal{U}$ and \mathcal{B} will denote that random variable x follows the normal, complex normal, uniform or Bernoulli distribution, respectively. The expectation, Hermitian
transpose and conjugation are denoted as $E[\cdot], [\cdot]^H$ and $[\cdot]^*$.

6.2 OFDM System and Channel Models

6.2.1 OFDM system model

Consider an OFDM system with N subcarriers modulated by a sequence of N complex data symbols $\mathbf{d} = [d_0, d_1, ..., d_{N-1}]^T$. The data symbols are selected uniformly from a particular constellation such as quadrature amplitude modulation (QAM). The modulation process can be implemented efficiently using N-points IFFT, where the IFFT output during the ℓ th OFDM block $\mathbf{x}(\ell) = \mathbf{F}^H \mathbf{d}(\ell)$. Matrix \mathbf{F} is the normalized FFT matrix whose elements are defined as $F_{i,k} = \kappa e^{-\omega i k}$, $\kappa = 1/\sqrt{N}$, $\omega = j2\pi/N$, $j = \sqrt{-1}$ and $\{i, k\} = 0, 1, ..., N-1$ denote the row and column numbers, respectively. Since \mathbf{F} is a unitary matrix, then $\mathbf{F}^H = \mathbf{F}^{-1}$. Consequently, the *n*th sample in \mathbf{x} can be expressed as

$$x_n(\ell) = \kappa \sum_{i=0}^{N-1} d_i(\ell) e^{\omega i n}, \quad n = 0, 1, ..., N-1.$$
(6.1)

To eliminate the inter-symbol-interference (ISI) between consecutive OFDM symbols in frequency-selective multipath fading channels, a cyclic prefix (CP) of length \bar{N} samples no less than the normalized channel delay spread (L_h) is formed by copying the last \bar{N} samples of \mathbf{x} and appending them at the front of \mathbf{x} to compose the transmission symbol with a total length $N_t = N + \bar{N}$ samples and a duration of T_t seconds. Hence, the complex baseband symbol transmitted during the ℓ th signaling period can be expressed as

$$\tilde{\mathbf{x}}(\ell) = [x_{N-\bar{N}}(\ell), x_{N-\bar{N}+1}(\ell), \dots, x_{N-1}(\ell), x_0(\ell), x_1(\ell), \dots, x_{N-1}(\ell)]^T.$$
(6.2)

Consequently, the *i*th transmitted sample $\tilde{x}_i(\ell) = x_{\langle i-\bar{N} \rangle}(\ell)$, $i = 0, 1, ..., N_t - 1$, where $\langle i \rangle \triangleq i \mod N$. Then, the sequence $\tilde{\mathbf{x}}$ is upsampled, filtered and up-converted to a radio frequency centered at f_c .

At the receiver frontend, the received signal is down-converted to baseband

and sampled at a rate $T_s = T_t/N_t$. In this work, we assume that the channel is composed of $L_h + 1$ independent multipath components each of which has a gain h_m and delay $m \times T_s$, where $m \in \{0, 1, ..., \bar{N}\}$. The channel taps are assumed to be constant over N OFDM symbols, which corresponds to quasi-static multipath fading channels. The received sequence $\tilde{\mathbf{y}} = \tilde{\mathbb{H}}(\ell)\tilde{\mathbf{x}}(\ell) + \tilde{\mathbf{z}}(\ell)$, where the channel matrix $\tilde{\mathbb{H}}$ is an $N_t \times N_t$ Toeplitz matrix with h_0 on the principal diagonal and $h_1, ..., h_{L_h}$ on the minor diagonals, respectively [24] and $\tilde{\mathbf{z}}(\ell)$ represents the overall additive noise that includes AWGN and IN. For notations' simplicity, we omit the subscript ℓ in the remaining parts of this section. Given that $L_h < \bar{N}$, the *n*th sample of $\tilde{\mathbf{y}}$ can be expressed as $\tilde{y}_n = \sum_{i=0}^{L_h} h_i x_{\langle n-i-\bar{N} \rangle} + z_n$. Subsequently, the receiver should identify and extract the sequence $\mathbf{y} = [\tilde{y}_{\bar{N}}, \tilde{y}_{\bar{N}+1}, \ldots, \tilde{y}_{N+\bar{N}-1}]$ and discard the first \bar{N} CP samples $[\tilde{y}_0, \tilde{y}_1, \ldots, \tilde{y}_{\bar{N}-1}]$. Therefore,

$$y_n = \sum_{i=0}^{L_h} h_i x_{\langle n-i \rangle} + z_{n+\bar{N}}, \quad 0 \le n \le N-1.$$
 (6.3)

In vector notation, the sequence \mathbf{y} can be expressed as

$$\mathbf{y} = \mathbb{H}\mathbf{x} + \mathbf{z} \tag{6.4}$$

where $\mathbf{z} = [z_{\bar{N}}, z_{\bar{N}+1}, \cdots, z_{\bar{N}+N-1}]^T$, and the channel matrix \mathbb{H} is circulant [24], [25].

6.2.2 Impulsive noise model

The IN is usually characterized by bursts of one or more short pulses whose amplitude g, duration T_I , and time of occurrence are all random parameters. The main models used for the amplitude distribution are the Middleton class-A [26], exponential [18] and Gaussian [27]. The common distributions of the bursts' arrival process are partitioned Markov chains (PMC) [18], Gated Bernoulli-Gaussian (GBG) [27], and Poisson [19]. The burst width distribution is usually modeled using the PMC or as the sum of two lognormal random variables [13].

To enable analytical performance evaluation over IN channels, the GBG model is widely adopted in the literature where the IN is modeled as a sequence of independent, Bernoulli distributed bursts with equal and fixed width, which is equal to the duration of a single time-domain OFDM sample [15, 16, 27]. Hence, the overall noise samples at the receiver frontend $z_n = w_n + \rho_n g_n$, where the AWGN $w_n \sim \mathcal{N}_c (0, \sigma_w^2)$ and $\sigma_w^2 = E \left[|w_n|^2 \right] = N_0/2$. The IN component $\rho_n g_n$ is modeled as a gated Gaussian process, where $\rho_n \sim \mathcal{B} \{0, 1\}$ where $p = P\{\rho_n = 1\}$, and $g_n \sim \mathcal{N}_c (0, \sigma_g^2)$ where $\sigma_g^2 \gg \sigma_w^2$ [15].

In this work, the GBG model is adopted to represent the IN, which is modeled as a sequence of independent bursts each of which consists of $1 \leq \varkappa \leq N + \bar{N}$ pulses. In general, the burst width \varkappa is a random variable whose distribution can be modeled as a uniform or lognormal distribution, or by using the PMC [13]. Therefore, the overall received noise $\mathbf{z} = \mathbf{w} + \rho \mathbf{b} \mathbf{g}$, where $\mathbf{z} = [z_0, z_1, ..., z_{N+\bar{N}-1}]$. The vector \mathbf{b} is used to specify the position as well as the width of the IN burst with respect to the OFDM symbol. Given that n_0 is the index of the first OFDM sample that is affected by a noise burst, then the elements of \mathbf{b} are

$$b_n = \begin{cases} 1, & n_0 \le n < n_0 + \varkappa \\ 0, & \text{otherwise} \end{cases}, \tag{6.5}$$

where $n_0 \sim \mathcal{U}\left[0, N + \overline{N} - \varkappa\right]$. Therefore, the location of the noise burst will be random and uniformly distributed over the OFDM symbol period. Thus, any IN burst can affect a maximum of one OFDM symbol.

6.3 Proposed System Model

The proposed system is similar to the conventional OFDM system described in Section II. However as depicted in Fig. 6.1, after the IFFT, N OFDM symbols are interleaved

using a simple row/column interleaver,

$$write \begin{cases} x_0(0) & x_0(1) & \cdots & x_0(N-1) \\ x_1(0) & x_1(1) & \cdots & x_1(N-1) \\ \vdots & \vdots & & \\ x_{N-1}(0) & x_{N-1}(1) & \cdots & x_{N-1}(N-1) \end{cases}$$
(6.6)

It is worth noting that using block interleavers as represented in (6.6) simplifies the analysis, but the delay of such interleavers is equal to N OFDM symbols. Alternatively, convolutional interleavers can offer the same BER performance with only N/2 symbols delay.

The interleaver produces N interleaved symbols $\{\breve{\mathbf{x}}(0), \breve{\mathbf{x}}(1), ..., \breve{\mathbf{x}}(N-1)\}$ where

$$\mathbf{\breve{x}}(\ell) = \begin{bmatrix} \breve{x}_{\ell}(0) & \breve{x}_{\ell}(1) & \cdots & \breve{x}_{\ell}(N-1) \end{bmatrix}^{T}.$$
(6.7)

The CP samples are formed by copying the last \bar{N} samples of $\breve{\mathbf{x}}$ and appending them at the front of $\breve{\mathbf{x}}$ to compose the transmission symbol. Hence, the complex baseband symbol transmitted during the ℓ th signaling period can be expressed as

$$\tilde{\mathbf{x}}(\ell) = [\breve{x}_{\ell}(N-\bar{N}), \, \breve{x}_{\ell}(N-\bar{N}+1), \dots, \, \breve{x}_{\ell}(N-1), \, \breve{x}_{\ell}(0), \, \breve{x}_{\ell}(1), \dots, \, \breve{x}_{\ell}(N-1)]^{T}.$$
(6.8)

Consequently, the *i*th transmitted sample is $\tilde{x}_i(\ell) = \check{x}_\ell(\langle i - \bar{N} \rangle)$, $i = 0, 1, ..., N_t - 1$. The remaining transmission and reception processes are similar to conventional OFDM systems. Therefore, the received sequence $\tilde{\mathbf{y}}(\ell) = \tilde{\mathbb{H}}(\ell)\tilde{\mathbf{x}}(\ell) + \tilde{\mathbf{z}}(\ell)$, where $\tilde{y}_n = \sum_{i=0}^{L_h} h_i \check{x}_\ell(\langle n - i - \bar{N} \rangle) + z_n$. Subsequently, after discarding the first \bar{N} CP samples, the remaining samples can be expressed as

$$\breve{y}_n = \sum_{i=0}^{L_h} h_i \,\breve{x}_\ell(\langle n-i\rangle) + z_{n+\bar{N}}, \quad 0 \le n \le N-1.$$
(6.9)

In vector notation $\mathbf{\breve{y}} = \mathbb{H}\mathbf{\breve{x}} + \mathbf{z}$.

Chapter 6: BER Reduction of OFDM Based Broadband Communication Systems over Multipath Channels with Impulsive Noise 134



Figure 6.1: The proposed TDI system in fading channels.

6.4 TDI Symbol Blanking

To simplify the discussion, assume initially that the channel matrix \mathbb{H} is the identity matrix. Therefore, the received N samples, after removing the CP, can be written as

$$\breve{\mathbf{y}}(\ell) = \breve{\mathbf{x}}(\ell) + \mathbf{w}(\ell) + \rho(\ell)\mathbf{b}(\ell)\mathbf{g}(\ell).$$
(6.10)

Since the deinterleaving process interleaves the vectors $\mathbf{w}(\ell)$, $\mathbf{b}(\ell)$ and $\mathbf{g}(\ell)$, it yields

$$\mathbf{y}(\ell) = \mathbf{x}(\ell) + \breve{\mathbf{w}}(\ell) + \breve{\rho}(\ell)\breve{\mathbf{b}}(\ell)\breve{\mathbf{g}}(\ell)$$
(6.11)

where $\mathbf{\check{w}}$ and $\mathbf{\check{g}}$ are the interleaved AWGN and IN vectors, which have the same statistical properties as \mathbf{w} and \mathbf{g} . The burst and sample gating factors $\check{\rho}$ and $\mathbf{\check{b}}$ have different properties from the original ρ and \mathbf{b} . For example, given that $\rho(0) = 1$ and $\varkappa = N_t$, then $\check{\rho}(\ell) = b_0(\ell) = 1 \forall \ell \in \{1, 2, ...N\}$, *i.e.*, the first sample of every OFDM symbol after deinterleaving will be affected by an IN pulse, while all other samples in all symbols will be IN free. Consequently, the FFT output can be expressed as

$$r_k(\ell) = \kappa \sum_{n=0}^{N-1} y_n(\ell) e^{-\omega nk}$$
(6.12)

$$= \kappa \sum_{n=0}^{N-1} \left[x_n(\ell) + \breve{w}_n(\ell) + \breve{\rho}(\ell) \breve{b}_n(\ell) \breve{g}_n(\ell) \right] e^{-\omega nk}, \quad k = 0, 1, ..., N$$
(6113)

which can be reduced to

$$r_k(\ell) = d_k(\ell) + \psi_k(\ell) + u_k(\ell)$$
(6.14)

where the FFT of the AWGN $\psi_k \sim \mathcal{N}_c(0, \sigma_w^2)$. The last term in (6.14) represents the FFT of the IN

$$u_k(\ell) = \kappa \breve{\rho}(\ell) \sum_{n=0}^{N-1} \breve{b}_n(\ell) \breve{g}_n(\ell) e^{-\omega nk}$$

where $u_k \sim \mathcal{N}_c(0, \sigma_u^2)$, and

$$\sigma_u^2 = \kappa^2 \check{\rho}(\ell) \sigma_g^2 \sum_{n=0}^{N-1} \check{b}_n(\ell)$$

= $\kappa^2 \check{\rho}(\ell) \sigma_g^2 \check{\varkappa}(\ell)$ (6.15)

where $\breve{\varkappa} \neq \varkappa$ is the number of nonzero elements in the vector $\breve{\mathbf{b}}(\ell)$, *i.e.*, the Hamming weight of $\breve{\mathbf{b}}(\ell)$.

It can be concluded from (6.14) and (6.15) that the deinterleaving process spreads the IN burst over most OFDM symbols within the deinterleaved block. The FFT process applied after the deinterleaving averages the IN pulses over all subcarriers within a given OFDM symbol, which may cause the loss of up to \varkappa OFDM symbols because $\sigma_q^2 \gg \sigma_w^2$.

An efficient solution to mitigate the effects of impulse noise is to apply blanking [12, 14], where the received samples with high amplitudes are set to zero. The contaminated samples are detected and suppressed by comparing the received samples' values with a particular threshold \mathcal{T}_1 . Therefore, the output of the blanking nonlinearity is $\breve{q}_n = \breve{y}_n \forall |\breve{y}_n| \leq \mathcal{T}_1$ and 0 otherwise. We refer to this approach as sample-by-sample blanking. After blanking and deinterleaving, the *n*th sample of the FFT input $q_n = y_n \forall |y_n| \leq \mathcal{T}_1$ and 0 otherwise, where $y_n = x_n + w_n + \rho b_n g_n$. Therefore, the FFT output $\mathbf{s}(\ell) = \mathbf{Fq}(\ell)$ and the *k*th subcarrier can be written as

$$s_{k}(\ell) = \kappa \sum_{n=0}^{N-1} q_{n}(\ell) e^{-\omega nk}, \ k = 0, \ 1, \ ..., \ N-1$$
$$= \kappa \sum_{n \in \mathcal{C}} y_{n}(\ell) e^{-\omega nk}$$
(6.16)

where $C = \{n \in \{0, 1, ..., N-1\} | q_n \neq 0\}$. The average error probability can be expressed as

$$P_e = \sum_{\mathcal{C}} P(e|\mathcal{C})P(\mathcal{C}).$$
(6.17)

However, evaluating $P(\mathcal{C})$ is quite difficult because it is a function of the bursts probability vector $\boldsymbol{\rho}$, the bursts location vector \mathbf{n}_0 , the bursts width vector $\boldsymbol{\varkappa}$, and the threshold \mathcal{T}_1 . Furthermore, it depends on the number and location of the blanked information samples.

The sample blanking threshold \mathcal{T}_1 should be selected to minimize the BER, and it is a function of several variables such as signal-to-noise ratio (SNR), signal-to-IN ratio (SIR) and \varkappa [12]. It is worth noting that the blanking process is not ideal in the sense that it will not necessarily blank all IN samples, and may blank some information samples, which is due to the high peak-to-average-power-ratio (PAPR) problem inherent in OFDM systems [28]. In addition, the sample-by-sample blanking is not feasible in frequency-selective channels due to the vast amplitude fluctuations that a signal may experience in such channels [16].

A possible remedy for these problems is to identify the contaminated symbols first, and then identify and blank the corrupted samples within that symbol. A simple technique to identify the contaminated symbols is to introduce an additional threshold \mathcal{T}_2 for the blanking process, where \mathcal{T}_2 denotes the number of samples with amplitudes larger than \mathcal{T}_1 . Then, all samples that have $|y_n| \geq \mathcal{T}_1$ will be blanked in a given received symbol, if and only if, the number of samples that will be blanked is larger than \mathcal{T}_2 . Such blanking policy minimizes the chances to blank IN-free information samples from symbols other than the contaminated ones, and hence can improve the BER.

In a frequency-selective channel, despite the fact that detecting samples that are hit by IN, based on a threshold-based mechanism is impossible, detecting corrupted symbols based on the described technique is fairly easy and accurate. Hence, in frquency-selective channels, we simply blank the entire contaminated OFDM symbol. This approach is called symbol blanking for the rest of the chapter. The time diversity provided by the TDI interleaving mechanism allows symbol blanking without any significant loss.

Based on the widely used assumption that $\sigma_g^2 \gg \sigma_w^2$, the symbol blanking process is expected to be highly accurate. By assuming a perfect burst detection process, the average probability of error over a block of N symbols can be expressed as

$$P_e = \sum_{\epsilon=0}^{N-1} P(e|\epsilon)P(\epsilon)$$
(6.18)

where ϵ denotes the total number of IN bursts per block of N OFDM symbols, which has a binomial PDF

$$P(\epsilon = i) = \binom{N}{i} p^{i} (1-p)^{N-i}.$$
(6.19)

It is worth noting the symbol blanking substantially relaxes the system sensitivity to \mathcal{T}_1 , which is difficult to estimate accurately in conventional OFDM systems [17,29].

6.5 System Performance in AWGN Channels

In the AWGN channel, the error probability $P(e|\epsilon)$ can be derived by rewriting (6.16) as

$$s_k(\ell|\epsilon) = \sum_{n=\epsilon}^{N-1} y_n(\ell) e^{-\omega nk}$$
$$= \beta d_k + \kappa^2 \sum_{\substack{i=0\\i\neq k}}^{N-1} d_i(\ell) \alpha_{i,k} e^{-\frac{\omega}{2}(\epsilon+1)(i-k)} + \psi_{w\epsilon}$$
(6.20)

where $\alpha_{i,k} = \frac{\sin\left[\pi\epsilon\kappa^2(k-i)\right]}{\sin\left[\pi\kappa^2(i-k)\right]}$, $\beta = \frac{N-\epsilon}{N}$, $\psi_{w\epsilon}$ is the FFT of the AWGN given that ϵ samples are blanked. Hence, $\psi_{w\epsilon} \sim \mathcal{N}\left(0, \sigma_{w\epsilon}^2\right)$ where $\sigma_{w\epsilon}^2 = \beta^2 \sigma_w^2$.

It can be noted from (6.20) that the blanked samples attenuate the desired signal amplitude and introduces inter-carrier interference (ICI), which is given in the second term of (6.20). For large N, the central limit theorem can be invoked and

Chapter 6: BER Reduction of OFDM Based Broadband Communication Systems over Multipath Channels with Impulsive Noise 138

$$ICI \sim \mathcal{N}\left(0, \sigma_{ICI,k}^2\right)$$
, where

$$\sigma_{ICI,k}^{2} = \kappa^{4} \sigma_{d}^{2} \sum_{i=0}^{N-1} \alpha_{i,k}^{2}, \, i \neq k$$
(6.21)

By noting that all subcarriers will experience stochastically the same SINR, we can assume, without loss of generality, that k = 0 in (6.21). Moreover, since the ICI and noise component are mutually independent, the SINR conditioned on ϵ can be written as

$$\operatorname{SINR}|_{\epsilon} = \frac{\sigma_d^2 \beta^2}{\sigma_{w\epsilon}^2 + \sigma_{ICI}^2} \\ = \frac{1}{\frac{1}{\frac{1}{\gamma} + \frac{\kappa^4}{\beta^2} \sum_{i=1}^{N-1} \alpha_{i,k}^2}}$$
(6.22)

where $\gamma = \sigma_d^2 / \sigma_Z^2$. Therefore, by assuming perfect burst detection and symbol blanking, the conditional BER for QPSK modulated subcarriers can be expressed as

$$P(e|\epsilon) = Q\left(\sqrt{\text{SINR}\,|_{\epsilon}}\right). \tag{6.23}$$

6.6 System Performance in Fading Channels using ZF Equalizer

As it can be noted from (6.18), the probability of error is computed as the average of all possible symbol blanking scenarios. A particular case of interest is the one where $\epsilon = 0$, which corresponds to an IN-free frequency-selective fading channel. Therefore, we first consider the case of $\epsilon = 0$, then the results are generalized for $\epsilon > 0$. The TDI system block diagram is given in Fig. 6.1 for notation clarification.

6.6.1 No samples blanked, $\epsilon = 0$

The receiver design can be performed by noting that $\mathbb{H} = \mathbf{F}^H \mathbf{H} \mathbf{F}$ [30], where

$$\mathbf{H} = \text{diag}([H_0, H_1, \cdots, H_{N-1}]), H_k = \sum_{m=0}^{L_h} h_m e^{-\omega m k}$$

Thus (6.4) can be written as

$$\breve{\mathbf{y}} = \mathbf{F}^H \mathbf{H} \mathbf{F} \breve{\mathbf{x}} + \mathbf{w}. \tag{6.24}$$

Therefore, the ZF equalizer output can be written as $\mathbf{\breve{s}} = \mathbf{F}^{H}\mathbf{H}^{-1}\mathbf{F}\mathbf{\breve{y}} = \mathbf{\breve{x}} + \mathbb{V}\mathbf{w}$, where $\mathbb{V} \triangleq \mathbf{F}^{H}\mathbf{H}^{-1}\mathbf{F}$. However, since \mathbf{H}^{-1} is a diagonal matrix,

$$\mathbf{H}^{-1} = \text{diag}\left(\left[\frac{1}{H_0}, \frac{1}{H_1}, \cdots, \frac{1}{H_{N-1}}\right]\right)$$
(6.25)

and matrix \mathbb{V} is circulant, the first row of \mathbb{V} , using the Matlab notations, is given by

$$\mathbb{V}(1,:) = \kappa^2 \left[\sum_{k=0}^{N-1} \frac{1}{H_k}, \sum_{k=0}^{N-1} \frac{e^{-\omega k}}{H_k}, \cdots, \sum_{k=0}^{N-1} \frac{e^{-(N-1)\omega k}}{H_k} \right].$$
(6.26)

It can be noted from (6.26) that each element in \mathbb{V} is composed of a mixture of all elements of \mathbf{H}^{-1} . Specifically, each row in \mathbb{V} consists of the FFT of the vector $\left[\frac{1}{H_0}, \frac{1}{H_1}, \cdots, \frac{1}{H_{N-1}}\right]$. Consequently, the term $\mathbb{V}\mathbf{w} \triangleq \mathbf{\Phi}$ is given by

$$\breve{\boldsymbol{\Phi}} = \kappa \left[\begin{array}{ccc} N-1 \\ \sum \limits_{i=0}^{N-1} w_i \Delta_i & \sum \limits_{i=0}^{N-1} w_i \Delta_{\langle i-1 \rangle} & \sum \limits_{i=0}^{N-1} w_i \Delta_{\langle i-2 \rangle} & \cdots & \sum \limits_{i=0}^{N-1} w_i \Delta_{\langle i+1 \rangle} \end{array} \right]^T \quad (6.27)$$

where $\Delta_i = \sum_{k=0}^{N-1} \frac{e^{-\omega k i}}{H_k}$. After deinterleaving, the ℓ th OFDM symbol $\mathbf{s}(\ell) = \mathbf{x}(\ell) + \mathbf{\Phi}(\ell)$, where $\mathbf{x}(\ell) = \mathbf{F}^H \mathbf{d}(\ell)$ and $\mathbf{\Phi}$ is formed by equalizing and deinterleaving the additive

noise,

$$\Phi(\ell) = \kappa \left[\sum_{i=0}^{N-1} w_i(0) \Delta_{\langle -\ell+i \rangle}(0) \sum_{i=0}^{N-1} w_i(1) \Delta_{\langle -\ell+i \rangle}(1) \cdots \sum_{i=0}^{N-1} w_i(N-1) \Delta_{\langle -\ell+i \rangle}(N-1) \right]^T. \quad (6.28)$$

The FFT is then applied to extract the decision variables $\mathbf{r}(\ell) = \mathbf{d}(\ell) + \mathbf{F} \mathbf{\Phi}(\ell)$.

As an example, for the case of $\ell=0$, the noise component can be described as

$$\mathbf{F}\Phi(0) = \kappa^2 \left[\sum_{j,i=0}^{N-1} w_i(j)\Delta_i(j) \sum_{j,i=0}^{N-1} w_i(j)\Delta_i(j)\mathrm{e}^{-\omega j} \cdots \sum_{j,i=0}^{N-1} w_i(j)\Delta_i(j)\mathrm{e}^{-(N-1)\omega j} \right]^T$$
(6.29)

Since all samples are identically distributed, without loss of generality, we consider the first subcarrier of the first OFDM symbol, which is given by

$$r_0(0) = d_0(0) + \kappa^2 \sum_{j=0}^{N-1} \sum_{i=0}^{N-1} w_i(j) \Delta_i(j).$$
(6.30)

Substituting the value of Δ_i and noting that $H_k(\ell) = H_k \ \forall \ell$ for the considered quasi-static channel, then (6.30) can be written as

$$r_0(0) = d_0(0) + \kappa^2 \sum_{j=0}^{N-1} \sum_{i=0}^{N-1} \sum_{k=0}^{N-1} \frac{e^{-\omega ki}}{H_k} w_i(j).$$
(6.31)

By comparing (6.31) to the standard OFDM, which has

$$r_0^{std} = d_0 + \frac{\kappa^2}{H_0} \sum_{i=0}^{N-1} w_i.$$
(6.32)

It can be noted that the channel frequency response at each subcarrier in the TDI system is a mixture of the channel frequency responses H_i of the entire interleaving block length, which is caused by the deinterleaving and the last FFT operations.

The conditional SNR for a given channel matrix \mathbf{H} can be expressed as

SNR
$$|_{\mathbf{H}} = \frac{E\left\{|d_0|^2\right\}}{E\left\{\left|\kappa^2 \sum_{i=0}^{N-1} \sum_{j=0}^{N-1} w_i(j)\Delta_i\right|^2\right\}}.$$
 (6.33)

Since $E\{w_i(j)w_v^*(k)\} = 0$ for $j \neq k$ or $i \neq v$, the denominator of (6.33) can be simplified as follows,

$$E\left\{\left|\kappa^{2}\sum_{i=0}^{N-1}\sum_{j=0}^{N-1}w_{i}(j)\Delta_{i}\right|^{2}\right\} = \kappa^{4}\sum_{i=0}^{N-1}\sum_{i'=0}^{N-1}\Delta_{i}\Delta_{i'}^{*}\sum_{j=0}^{N-1}\sum_{j'=0}^{N-1}E\left\{w_{i}(j)w_{i'}^{*}(j')\right\}$$
$$= \kappa^{2}\sigma_{w}^{2}\sum_{i=0}^{N-1}|\Delta_{i}|^{2}.$$
(6.34)

It is worth noting that (6.34) can be obtained using Parseval's theorem as well,

$$E\left\{ \left| \sigma_w^2 \sum_{i=0}^{N-1} \sum_{j=0}^{N-1} z_i(j) \Delta_i \right|^2 \right\} = \kappa^2 \sigma_w^2 \sum_{k=0}^{N-1} \frac{1}{|H_k|^2}.$$
 (6.35)

Therefore, the SNR using a ZF equalizer is given by

SNR
$$|_{\mathbf{H}} = \frac{\sigma_d^2}{\kappa^2 \sigma_w^2 \sum_{k=0}^{N-1} \frac{1}{|H_k|^2}}$$
 (6.36)

where $E\left\{|d_0|^2\right\} = \sigma_d^2$. By denoting $\varphi^{-1} = \kappa^2 \sum_{k=0}^{N-1} \frac{1}{|H_k|^2}$, then

$$\operatorname{SNR}|_{\mathbf{H}} = \frac{\gamma}{\varphi^{-1}} = \varphi\gamma \tag{6.37}$$

where $\varphi^{-1} = 1/\varphi$ and $\gamma = \sigma_d^2/\sigma_w^2$. It is interesting to note that (6.36) and [23, Eq. (61)] are identical, and (6.36) is similar to [31, Eq. 17] where the block length M = N, [22, Eq. 24] for the case of $L_i = 1$, and with [32, Eq. 2] for $t_{k,i} = 1$. Furthermore, it can be noted from (6.36) that if any subcarrier goes through a deep

fade, *i.e.* $\exists k, |H_k|^2 \to 0$, then $1/|H_k|^2 \to \infty$, and hence SNR $\to 0$. Therefore, ZF equalizer is expected to offer poor BER in frequency-selective fading channels even without IN.

Given that the PDF $f_{\varphi}(\varphi)$ is known, the BER can be expressed as

$$P_{ZF}(\gamma) = \int_{0}^{\infty} Q\left(\sqrt{\varphi\gamma}\right) f_{\varphi}(\varphi) \, d\varphi.$$
(6.38)

Note that φ is composed of the sum of N correlated random variables each of which has the form $|H_k|^{-2}$. If the channel frequency response parameters $\{H_0, ..., H_{N-1}\}$ are i.i.d. RV, then evaluating $f_{\varphi}(\varphi)$ might be feasible for particular scenarios. For example, if we assume that $H_i \sim \mathcal{N}_c(0,1)$, then $|H_i|^2$ has a Chi-square χ^2 PDF with two degrees of freedom, and $|H_i|^{-2}$ has an inverse- χ^2 PDF. Since the inverse χ^2 PDF can be expressed in terms of the inverse Gamma PDF, the PDF of the sum $\sum_i |H_i|^{-2}$ can be evaluated as described in [33]. However, since H_k are correlated, evaluating $f_{\varphi}(\varphi)$ analytically is difficult. Hence, semi-analytical solutions can be incorporated [22, 31, 32]. In addition, based on the work of Wang and Giannakis [34], which states that systems with instantaneous SNR similar to (6.37) will not have diversity gain if $\lim_{\gamma \to 0^+} f_{\varphi}(\varphi) = c > 0$. McCloud [31] demonstrated that $f_{\varphi}(\varphi)$ is bounded away from zero as $\varphi \to 0^+$. This result proves that the proposed interleaved system with ZF equalizer has diversity gain, but without quantifying this gain. Consequently, Monte Carlo simulations can be invoked to evaluate the system performance. Besides, it can be clearly noticed from (6.36) that SNR $|_{\mathbf{H}} \rightarrow 0$ if any of the subcarriers goes into a deep fade. Consequently, the performance improvement will be apparent only at $\gamma \gg 1.$

In this analysis, we keep the cyclic prefix for the TDI system, in spite of the fact that the interleaver destroys the channel orthogonality. The cyclic prefix here acts merely as a time guard band and can be replaced with zeros for saving power.

6.6.2 Samples blanked, $\epsilon > 0$

In the process of OFDM symbol blanking, a sequence of blanked and non-blanked symbols is generated. Therefore, the input of the deinterleaver can be written as

$$\mathbf{\breve{s}}(\ell) = \begin{cases} 0 \times \mathbf{\breve{y}}(\ell), & \text{blanked symbol} \\ \mathbf{\breve{x}}(\ell) + \mathbf{\breve{\Phi}}(\ell), & \text{otherwise} \end{cases}$$

where $\check{\Phi}(\ell)$ is defined in (6.27). The deinterleaving process rearranges the N OFDM symbols back to their original order before interleaving, which yields $\mathbf{s}(\ell) = \mathbb{B}\mathbf{F}^H \mathbf{d}(\ell) + \mathbb{B}\Phi(\ell)$, where \mathbb{B} is a matrix that specifies the blanking process. If no symbol is blanked, then \mathbb{B} is the identity matrix. Each of the samples to be blanked is nulled by setting the corresponding main diagonal element to zero. The received signal is then achieved by computing the FFT of \mathbf{s} ,

$$\mathbf{r}(\ell) = \mathbf{F} \mathbb{B} \mathbf{F}^H \mathbf{d}(\ell) + \mathbf{F} \mathbb{B} \mathbf{\Phi}(\ell).$$
(6.39)

Using the Appendix, the first term in (6.39), $\mathbf{v} \triangleq \mathbf{F} \mathbb{B} \mathbf{F}^H \mathbf{d}(\ell)$ can be calculated for a particular \mathbb{B} . Thus, the *k*th sample of \mathbf{v} is given by,

$$v_k(\ell|\epsilon) = \beta d_k + \kappa^2 \sum_{\substack{i=0\\i\neq k}}^{N-1} d_i \frac{\sin\left[\pi\epsilon\kappa^2(k-i)\right]}{\sin\left[\pi\kappa^2(i-k)\right]} e^{-j\frac{\pi}{N}(\epsilon+1)(i-k)}$$
(6.40)

where $\beta = \frac{N-\epsilon}{N}$ and $\epsilon = N - tr(\mathbb{B})$ represents the number of blanked symbols and $tr(\cdot)$ denotes the trace.

The second term in (6.39), can be calculated for the $\ell = 0$ case as follows,

$$\mathbf{F}\mathbb{B}\Phi(0) = \kappa^{3} \left[\sum_{j,i,k=0}^{N-1} b_{j} \frac{w_{i}(j)}{H_{k}(j)} e^{-\omega(ik+0 j)} \sum_{j,i,k=0}^{N-1} b_{j} \frac{w_{i}(j)}{H_{k}(j)} e^{-\omega(ik+j)} \cdots \sum_{j,i,k=0}^{N-1} b_{j} \frac{w_{i}(j)}{H_{k}(j)} e^{-\omega(ik+Nj)} \right]^{T}$$

where $b_j = \mathbb{B}(j, j)$. The assumption that $\ell = 0$ is used for notational simplicity.

Following the same approach described in previous subsection we obtain

$$r_{k} = \beta d_{k} + \kappa^{2} \sum_{\substack{i=0\\i\neq k}}^{N-1} d_{i} \frac{\sin\left[\pi \epsilon \kappa^{2}(k-i)\right]}{\sin\left[\pi \kappa^{2}(i-k)\right]} e^{-j\frac{\pi}{N}(\epsilon+1)(i-k)} + \kappa^{3} \sum_{j=0}^{N-1} b_{j} \sum_{i=0}^{N-1} w_{i}(j) \sum_{k=0}^{N-1} \frac{e^{-\omega ik}}{H_{k}(j)}$$
(6.41)

where the second and third terms in (6.41) represent the ICI and additive noise, respectively. As depicted in the Appendix, the noise variance is given by

$$\sigma_{w\epsilon}^{2} = E \left\{ \left| \kappa^{3} \sum_{j=0}^{N-1} b_{j} \sum_{i=0}^{N-1} w_{i}(j) \sum_{k=0}^{N-1} \frac{e^{-wik}}{H_{k}(j)} \right|^{2} \right\}$$
$$= \beta \kappa^{2} \sigma_{w}^{2} \sum_{k=0}^{N-1} \frac{1}{|H_{k}|^{2}}$$
(6.42)

 $\sigma_{w\epsilon}^2 = \beta^2 \sigma_w^2$ represents the variance of the AWGN given that ϵ samples are blanked, and the ICI variance is

$$\sigma_{ICI,k}^{2} = \kappa^{4} \sigma_{d}^{2} \sum_{\substack{i=0\\i\neq k}}^{N-1} \frac{\sin^{2} \left[\pi \epsilon \kappa^{2} (k-i)\right]}{\sin^{2} \left[\pi \kappa^{2} (i-k)\right]}.$$
(6.43)

Therefore, the SINR for the kth subcarrier can be calculated as

SINR
$$|_{\mathbf{H},\epsilon,k} = \beta^2 \frac{\sigma_d^2}{\sigma_{w\epsilon}^2 + \sigma_{ICI,k}^2}.$$
 (6.44)

Since all subcarriers will experience the same SINR, the subcarrier index k can be set to zero and dropped from (6.45) without loss of generality. Moreover, substituting (6.43) into (6.45) and simplifying the results gives

SINR
$$|_{\mathbf{H},\epsilon} = \left[\frac{\kappa^2}{\gamma} \sum_{i=0}^{N-1} \frac{1}{|H_i|^2} + \frac{\kappa^4}{\beta^2} \sum_{i=1}^{N-1} \frac{\sin^2(\pi\epsilon\kappa^2 i)}{\sin^2(\pi\kappa^2 i)}\right]^{-1}.$$
 (6.45)

The BER can be calculated by substituting (6.45) into the well known QPSK error

formula,

$$P(e|\epsilon, \mathbf{H}) = Q\left(\sqrt{\text{SINR}\,|_{\mathbf{H},\epsilon}}\right). \tag{6.46}$$

Therefore,

$$P(e|\mathbf{H}) = \sum_{\epsilon=0}^{N-1} Q\left(\sqrt{\mathrm{SINR}\,|_{\mathbf{H},\epsilon}}\right) \binom{N}{i} p^{\epsilon} (1-p)^{N-\epsilon}.$$
(6.47)

The conditional BER given in (6.45) can be used to compute the average BER semianalytically by generating a large number of realizations for **H**.

6.7 System Performance in Fading Channels using MMSE Equalizer

6.7.1 No samples blanked, $\epsilon = 0$

The typical approach to mitigate the performance degradation of ZF equalizers is to use MMSE equalizers, where the output signal of such equalizers can be calculated as $\mathbf{\breve{s}} = [\mathbb{H}^*\mathbb{H} + \xi \mathbf{I}]^{-1} \mathbb{H}^*\mathbf{\breve{y}}$, where $\xi = \gamma^{-1}$ for optimum error performance. Considering that fact that $\mathbb{H} = \mathbf{F}^H \mathbf{H} \mathbf{F}$ and noting that the input samples to the equalizer are interleaved, the equalizer output can be written as

$$\mathbf{\breve{s}} = \mathbf{F}^{H} \left[\mathbf{H}^{*} \mathbf{H} + \xi \mathbf{I} \right]^{-1} \mathbf{H}^{*} \mathbf{F} \mathbf{\breve{y}}.$$
(6.48)

By noting that $\breve{\mathbf{y}} = \mathbf{F}^H \mathbf{H} \mathbf{F} \breve{\mathbf{x}} + w$, then (6.48) is simplified to

$$\mathbf{\breve{s}} = \mathbf{F}^{H} \left[|\mathbf{H}|^{2} + \xi \mathbf{I} \right]^{-1} |\mathbf{H}|^{2} \mathbf{F} \mathbf{\breve{x}} + \mathbf{F}^{H} \left[|\mathbf{H}|^{2} + \xi \mathbf{I} \right]^{-1} \mathbf{H}^{*} \mathbf{F} \mathbf{w}.$$
(6.49)

Because \mathbf{H} , $\left[|\mathbf{H}|^2 + \xi \mathbf{I}\right]^{-1} |\mathbf{H}|^2 \triangleq \boldsymbol{\lambda}$ and $\left[|\mathbf{H}|^2 + \xi \mathbf{I}\right]^{-1} \mathbf{H}^* \triangleq \bar{\boldsymbol{\lambda}}$ are all $N \times N$ diagonal matrices, the equalizer output $\mathbf{\breve{s}}$ can be written as

$$\breve{\mathbf{s}} = \mathbf{F}^H \boldsymbol{\lambda} \, \mathbf{F} \breve{\mathbf{x}} + \mathbf{F}^H \bar{\boldsymbol{\lambda}} \, \mathbf{F} \mathbf{w} \tag{6.50}$$

where

$$\lambda_k = \frac{|H_k|^2}{|H_k|^2 + \xi}, \quad \bar{\lambda}_k = \frac{H_k^*}{|H_k|^2 + \xi}$$
(6.51)

Let $\mathbb{A} \triangleq \mathbf{F}^H \mathbf{\lambda} \mathbf{F}$ and $\mathbb{A}' \triangleq \mathbf{F}^H \bar{\mathbf{\lambda}} \mathbf{F}$ where \mathbb{A} and \mathbb{A}' are circulant matrices with their first rows defined as

$$\mathbb{A}(1,:) = \kappa^2 \left[\sum_{k=0}^{N-1} \lambda_k \sum_{k=0}^{N-1} \lambda_k e^{-\omega k} \sum_{k=0}^{N-1} \lambda_k e^{-\omega 2k} \cdots \sum_{k=0}^{N-1} \lambda_k e^{-\omega (N-1)k} \right]$$

$$(6.52)$$

and

$$\mathbb{A}'(1,:) = \kappa^2 \left[\sum_{k=0}^{N-1} \bar{\lambda}_k \sum_{k=0}^{N-1} \bar{\lambda}_k e^{-\omega k} \sum_{k=0}^{N-1} \bar{\lambda}_k e^{-\omega 2k} \cdots \sum_{k=0}^{N-1} \bar{\lambda}_k e^{-\omega (N-1)k} \right]$$

$$(6.53)$$

Thus, the MMSE equalizer output in (6.50) can be reduced to

$$\mathbf{\breve{s}} = \mathbb{A}\mathbf{\breve{x}} + \mathbb{A}'\mathbf{w}.$$
(6.54)

It is clear that when $\xi = 0$, then $\mathbb{A} = \mathbf{I}_N$ and $\mathbb{A}' = \mathbf{H}^{-1}$, which implies that the MMSE equalizer becomes ZF equalizer. For $\xi > 0$, $\mathbf{\breve{s}}$ will have some sort of inter-symbol interference (ISI) because \mathbb{A} is not a diagonal matrix. Consequently, a compromise has to be made between the degradation resulted from the ISI and noiseenhancement problem. Obviously, setting $\xi = 1/\gamma$ yields the conventional MMSE equalizer, which is widely used in the literature [35].

Following the same approach used for the ZF equalizer, we define $\breve{\Psi} \triangleq \mathbb{A}'w$ similar to $\breve{\Phi}$ except that $1/H_k$ is replaced by $\bar{\lambda}_k$. Therefore,

$$\mathbf{\breve{\Psi}}(\ell) = \kappa^2 \left[\sum_{i,k=0}^{N-1} w_i(\ell) \bar{\lambda}_k \mathrm{e}^{-\omega k \langle i-0 \rangle} \sum_{i,k=0}^{N-1} w_i(\ell) \bar{\lambda}_k \mathrm{e}^{-\omega k \langle i-1 \rangle} \cdots \sum_{i,k=0}^{N-1} w_i(\ell) \bar{\lambda}_k \mathrm{e}^{-\omega k \langle i-N+1 \rangle} \right]^T. \quad (6.55)$$

By denoting $\bar{\Lambda}_i = \kappa \sum_{k=0}^{N-1} \bar{\lambda}_k e^{-\omega k i}$, which is the is the FFT of $\bar{\lambda}$ ($\bar{\Lambda} = \mathcal{F}\{\bar{\lambda}\}$), then

(6.55) can be written as

$$\mathbf{\check{\Psi}}(\ell) = \kappa \left[\sum_{i=0}^{N-1} w_i(\ell) \bar{\Lambda}_i \sum_{i=0}^{N-1} w_i(\ell) \bar{\Lambda}_{\langle i-1 \rangle} \sum_{i=0}^{N-1} w_i(\ell) \bar{\Lambda}_{\langle i-2 \rangle} \cdots \sum_{i=0}^{N-1} w_i(\ell) \bar{\Lambda}_{\langle i-N+1 \rangle} \right]^T. \quad (6.56)$$

The symbol index ℓ is added in (6.55) to denote the noise sample taken at ℓ th OFDM symbol, which is needed to represent the deinterleaved noise. The first term in (6.54) can be written as

$$\breve{\boldsymbol{\Theta}}(\ell) \triangleq \mathbb{A}\breve{\mathbf{x}} = \kappa^2 \left[\sum_{i,k=0}^{N-1} x_{\ell}(i)\lambda_k \mathrm{e}^{-\omega k \langle i-0 \rangle} \sum_{i,k=0}^{N-1} x_{\ell}(i)\lambda_k \mathrm{e}^{-\omega k \langle i-1 \rangle_N} \\ \cdots \sum_{i,k=0}^{N-1} x_{\ell}(i)\lambda_k \mathrm{e}^{-\omega k \langle i-N+1 \rangle_N} \right].$$
(6.57)

By denoting $\Lambda_i = \kappa \sum_{k=0}^{N-1} \lambda_k e^{-\omega k i}$, which is the FFT of λ ($\Lambda = \mathcal{F}{\lambda}$), then (6.57) can be written as

$$\check{\boldsymbol{\Theta}}(\ell) = \kappa \left[\begin{array}{cc} N-1 \\ \sum \\ i=0 \end{array}^{N-1} x_{\ell}(i) \Lambda_{\langle i=0 \rangle} & \sum \\ i=0 \end{array}^{N-1} x_{\ell}(i) \Lambda_{\langle i=1 \rangle_{N}} & \cdots & \sum \\ i=0 \end{array}^{N-1} x_{\ell}(i) \Lambda_{\langle i=N+1 \rangle_{N}} \end{array} \right]^{T}.$$
(6.58)

The MMSE output is

$$\mathbf{\breve{s}}(\ell) = \mathbf{\breve{\Theta}}(\ell) + \mathbf{\breve{\Psi}}(\ell), \tag{6.59}$$

where

$$\breve{s}_n(\ell) = \kappa \sum_{i=0}^{N-1} x_\ell(i) \Lambda_{\langle i-n \rangle} + \kappa \sum_{i=0}^{N-1} w_i(\ell) \bar{\Lambda}_{\langle i-n \rangle}.$$
(6.60)

The next step is to deinterleave $\mathbf{\breve{s}}(\ell)$, which can be calculated by substituting (6.55) and (6.57) into (6.59). The result is the desired deinterleaved signal $\mathbf{s}(\ell) = \mathbf{\Theta}(\ell) + \Psi(\ell)$, where

$$s_n(\ell) = \kappa \sum_{i=0}^{N-1} x_n(i) \Lambda_{\langle i-\ell \rangle} + \kappa \sum_{i=0}^{N-1} w_i(n) \bar{\Lambda}_{\langle i-\ell \rangle}.$$
 (6.61)

The final step is to apply the FFT to get the N decision variables $\mathbf{r}(\ell) = \mathbf{Fs}(\ell) = \mathbf{F\Theta}(\ell) + \mathbf{F\Psi}(\ell)$, which are then applied to the demodulator to get the information symbols. After some algebraic manipulations, $\mathbf{F\Theta}(\ell)$ and $\mathbf{F\Psi}(\ell)$ can be written as

$$\mathbf{F}\boldsymbol{\Theta}(\ell) = \kappa^2 \left[\sum_{m,i=0}^{N-1} x_m(i)\Lambda_{\langle i-\ell \rangle} \quad \sum_{m,i=0}^{N-1} x_m(i)\Lambda_{\langle i-\ell \rangle} \mathrm{e}^{-\omega m} \\ \cdots \quad \sum_{m,i=0}^{N-1} x_m(i)\Lambda_{\langle i-\ell \rangle} \mathrm{e}^{-\omega(N-1)m} \right]^T \quad (6.62)$$

and

$$\mathbf{F}\Psi(\ell) = \kappa^2 \left[\sum_{m,i=0}^{N-1} w_i(m) \bar{\Lambda}_{\langle i-\ell \rangle} \quad \sum_{m,i=0}^{N-1} w_i(m) \bar{\Lambda}_{\langle i-\ell \rangle} \mathrm{e}^{-\omega m} \\ \cdots \quad \sum_{m,i=0}^{N-1} w_i(m) \bar{\Lambda}_{\langle i-\ell \rangle} \mathrm{e}^{-\omega(N-1)m} \right]^T. \quad (6.63)$$

Therefore, the FFT kth pin output is given by

$$r_{k}(\ell) = \kappa^{2} \sum_{m=0}^{N-1} \sum_{i=0}^{N-1} x_{m}(i) \Lambda_{\langle i-\ell \rangle} e^{-\omega km} + \kappa^{2} \sum_{m=0}^{N-1} \sum_{i=0}^{N-1} w_{i}(m) \bar{\Lambda}_{\langle i-\ell \rangle} e^{-\omega km}$$
$$= \kappa \sum_{i=0}^{N-1} \Lambda_{\langle i-\ell \rangle} d_{k}(i) + \kappa^{2} \sum_{i=0}^{N-1} \sum_{m=0}^{N-1} \bar{\Lambda}_{\langle i-\ell \rangle} w_{i}(m) e^{-\omega km}.$$
(6.64)

Since all the elements of \mathbf{r} are identically distributed, for simplicity, we consider the first FFT-pin output of the first OFDM symbol,

$$r_0(0) = \kappa d_0(0)\Lambda_0 + \kappa \sum_{i=1}^{N-1} \Lambda_i d_0(i) + \kappa^2 \sum_{i=0}^{N-1} \sum_{m=0}^{N-1} \bar{\Lambda}_i w_i(m).$$
(6.65)

The first term in (6.65) shows the desired information symbol, whereas the ISI is characterized by the second term and finally, the third term gives the noise component of the received signal. The SINR for a given channel matrix can be expressed as

SINR
$$|_{\mathbf{H},k=0} = \frac{E\left\{|\kappa d_0(0)\Lambda_0|^2\right\}}{E\left\{\left|\kappa \sum_{i=1}^{N-1} \Lambda_i d_0(i) + \kappa^2 \sum_{i=0}^{N-1} \sum_{m=0}^{N-1} \bar{\Lambda}_i w_i(m)\right|^2\right\}}.$$
 (6.66)

Since the noise and ISI terms are independent, it is possible to rewrite (6.66) as

SINR
$$|_{\mathbf{H},k=0} = \frac{E\left\{ |\kappa d_0(0)\Lambda_0|^2 \right\}}{E\left\{ \left| \kappa \sum_{i=1}^{N-1} \Lambda_i d_0(i) \right|^2 \right\} + E\left\{ \left| \kappa^2 \sum_{i=0}^{N-1} \sum_{m=0}^{N-1} \bar{\Lambda}_i w_i(m) \right|^2 \right\}}.$$
(6.67)

The expected value of the desired signal can be reduced to

$$E\left\{ |\kappa d_0(0)\Lambda_0|^2 \right\} = \kappa^2 E\left\{ |d_0(0)|^2 \right\} |\Lambda_0|^2$$

= $\kappa^2 \sigma_d^2 |\Lambda_0|^2$. (6.68)

The denominator of (6.67) is composed of two parts, \mathcal{F}_1 and \mathcal{F}_2 . Since $E\left\{d_0(i)d_0(i')\right\} = 0$ for $i \neq i'$, and $E\left\{|d_0(i)|^2\right\} = \sigma_d^2$, then \mathcal{F}_1 and \mathcal{F}_2 are reduced to,

$$F_1 \triangleq E\left\{ \left| \kappa \sum_{i=1}^{N-1} \Lambda_i d_0(i) \right|^2 \right\} = \kappa^2 \sigma_d^2 \sum_{i=1}^{N-1} |\Lambda_i|^2.$$
(6.69)

And similarly,

$$F_{2} \triangleq \kappa^{4} E \left\{ \left| \sum_{i=0}^{N-1} \sum_{m=0}^{N-1} \bar{\Lambda}_{i} w_{i}(m) \right|^{2} \right\} = \kappa^{2} \sigma_{w}^{2} \sum_{i=0}^{N-1} \left| \bar{\Lambda}_{i} \right|^{2}.$$
(6.70)

Following the same approach used for the ZF equalizer, and inserting (6.69) and (6.70) into (6.67) gives

SINR
$$|_{\mathbf{H},k=0} = \frac{\sigma_d^2 |\Lambda_0|^2}{\sigma_d^2 \sum_{i=1}^{N-1} |\Lambda_i|^2 + \sigma_w^2 \sum_{i=0}^{N-1} |\bar{\Lambda}_i|^2}.$$
 (6.71)

By substituting the values of Λ_0 , Λ_i and $\bar{\Lambda}_i$ into (6.71), and using the fact that $\sum_{i=0}^{N-1} \frac{e^{-\omega i(k-n)}}{H_k H_n^*} = 0, \forall k \neq n$, and noting that all subcarriers will experience the same SINR, then SINR at the FFT output can be expressed as

$$\operatorname{SINR}|_{\mathbf{H}} = \gamma \frac{\sum_{k=0}^{N-1} \lambda_k}{\sum_{k=0}^{N-1} \frac{\lambda_k}{|H_k|^2}}.$$
(6.72)

Similar to the ZF case, the SINR of the TDI given in (6.72) is equal to SINR of the

WHT-OFDM system [22], [23, Eq. (68)]. Consequently, the two systems will have identical BER performance in fading channels. The BER of the proposed system with MMSE can be calculated semi-analytically as

$$P_{MMSE}(\gamma) = \frac{1}{L} \sum_{i=1}^{L} Q\left(\sqrt{\text{SINR}\,|_{\mathbf{H}_i}}\right)$$
(6.73)

where \mathbf{H}_i is the *i*th realization of \mathbf{H} and L is the total number of realizations.

6.7.2 Samples blanked, $\epsilon > 0$

Similar to the ZF case, some OFDM symbols are blanked if they are declared as contaminated by the IN bursts. Therefore, the equalizer output (6.54) can be written as

$$\mathbf{\breve{s}}(\ell) = \begin{cases} 0 \times \mathbf{\breve{y}}, & \text{blanked symbol} \\ \mathbf{\breve{\Theta}}(\ell) + \mathbf{\breve{\Psi}}(\ell), & \text{otherwise} \end{cases}$$
(6.74)

where $\breve{\Theta}$ and $\breve{\Psi}$ are defined in (6.57), and (6.55), respectively. The deinterleaver output $\mathbf{s}(\ell) = \mathbb{B}\Theta(\ell) + \mathbb{B}\Psi(\ell)$ and the FFT output $\mathbf{r}(\ell) = \mathbf{Fs}(\ell) = \mathbf{FB}\Theta(\ell) + \mathbf{FB}\Psi(\ell)$, where

$$r_{k}(\ell) = \underbrace{\kappa^{2} \sum_{n=0}^{N-1} b_{n} \sum_{i=0}^{N-1} x_{n}(i) \Lambda_{\langle i-\ell \rangle_{N}} e^{-\omega nk}}_{v_{k}(\ell)} + \underbrace{\kappa^{2} \sum_{n=0}^{N-1} b_{n} \sum_{i=0}^{N-1} w_{i}(n) \bar{\Lambda}_{\langle i-\ell \rangle_{N}} e^{-\omega nk}}_{u_{k}(\ell)}.$$
(6.75)

The first term in (6.75) $v_k(\ell)$ can be written as

$$v_{k}(\ell) = \kappa^{2} \sum_{i=0}^{N-1} \Lambda_{\langle i-\ell \rangle_{N}} \sum_{n=0}^{N-1} b_{n} x_{n}(i) e^{-\omega nk}$$

= $\kappa \sum_{i=0}^{N-1} \Lambda_{\langle i-\ell \rangle_{N}} \left[\beta \ d_{k}(i) + \kappa^{2} \sum_{\substack{k=0\\k \neq n}}^{N-1} d_{k}(i) \frac{\sin \left[\pi \epsilon \kappa^{2}(n-k)\right]}{\sin \left[\pi \kappa^{2}(k-n)\right]} e^{-\frac{\omega}{2}(\epsilon+1)(k-n)} \right].$
(6.76)

To simplify the calculations, we consider $r_0(0)$, which is equal to

$$r_0(0) = \kappa \beta \Lambda_0 d_0(0) + \kappa \beta \sum_{i=1}^{N-1} \Lambda_i d_0(i) + \kappa^3 \sum_{i=0}^{N-1} \Lambda_i \sum_{k=1}^{N-1} d_k(i) \frac{\sin\left[\pi \epsilon \kappa^2 k\right]}{\sin\left[\pi \kappa^2 k\right]} e^{-j\frac{\pi}{N}(\epsilon+1)k}.$$
(6.77)

In (6.77), the first term includes the desired signal and the second term is the interference caused by the MMSE equalizer, whose power is denoted as $\sigma_{I,MMSE}^2$. The last term is the interference caused by losing the blanked ϵ samples from the OFDM symbol, which has a power of $\sigma_{I,Blank}^2$. The equalization and blanking interference power can be calculated as

$$\sigma_{I,MMSE}^{2} = E \left\{ \left| \kappa \beta \sum_{i=1}^{N-1} \Lambda_{i} d_{0}(i) \right|^{2} \right\}$$
$$= \kappa^{2} \beta^{2} \sigma_{d}^{2} \sum_{i=1}^{N-1} |\Lambda_{i}|^{2}.$$
(6.78)

and

$$\sigma_{I,Blank}^{2} = E \left\{ \left| \kappa^{3} \sum_{i=0}^{N-1} \Lambda_{i} \sum_{k=1}^{N-1} d_{k}(i) \frac{\sin\left[\pi \epsilon \kappa^{2} k\right]}{\sin\left[\pi \kappa^{2} k\right]} e^{-\frac{\omega}{2}(\epsilon+1)k} \right|^{2} \right\}$$
$$= \kappa^{6} \sigma_{d}^{2} \alpha_{\epsilon} \sum_{i=0}^{N-1} |\Lambda_{i}|^{2}, \qquad (6.79)$$

where $\alpha_{\epsilon} \triangleq \sum_{k=1}^{N-1} \frac{\sin^2(\pi \epsilon \kappa^2 k)}{\sin^2(\pi \kappa^2 k)}$. The second term in (6.75), $u_k(\ell)$, describes the noise component of the received signal,

$$u_k(\ell) = \kappa^2 \sum_{n=0}^{N-1} b_n \sum_{i=0}^{N-1} w_i(n) \bar{\Lambda}_{\langle i-\ell \rangle_N} e^{-\omega nk}$$
$$= \kappa^2 \sum_{i=0}^{N-1} \bar{\Lambda}_{\langle i-\ell \rangle_N} \sum_{n=0}^{N-1} b_n w_i(n) e^{-\omega nk}.$$

The variance of the noise component is

$$\sigma_{u\epsilon}^{2} = E \left\{ \left| \kappa^{2} \sum_{i=0}^{N-1} \bar{\Lambda}_{\langle i-\ell \rangle_{N}} \sum_{n=0}^{N-1} b_{n} w_{i}(n) \mathrm{e}^{-\omega n k} \right|^{2} \right\}$$
$$= \kappa^{2} \sigma_{w\epsilon}^{2} \sum_{i=0}^{N-1} |\bar{\Lambda}_{i}|^{2}. \tag{6.80}$$

Note that the second summation in (6.80) is just the FFT of the noise samples with ϵ samples missing. Consequently, the conditional SINR is calculated as

SINR
$$|_{\mathbf{H},\epsilon} = \kappa^2 \beta^2 \frac{\Lambda_0^2 \sigma_d^2}{\sigma_{w\epsilon}^2 + \sigma_{I,MMSE}^2 + \sigma_{I,Blank}^2}.$$
 (6.81)

After some mathematical manipulations (6.81) can be written as

SINR
$$|_{\mathbf{H},\epsilon} = \frac{\sum_k \lambda_k}{\frac{1}{\gamma} \sum_k \frac{\lambda_k}{|H_k|^2} + \frac{\kappa^2}{\beta^2} \alpha_\epsilon \frac{\sum_k \lambda_k^2}{\sum_k \lambda_k}}, \quad k = 0, 1, \dots N - 1.$$
 (6.82)

The BER can be computed semi-analytically as described in (6.47).

6.8 Numerical Results

Monte Carlo simulations are used to evaluate the performance of the proposed system over frequency-selective multipath fading channels. The OFDM system considered



Figure 6.2: BER versus the threshold \mathcal{T}_1 used for sample blanking at different SNRs.

in this chapter has N = 128 and $\bar{N} = 16$. All data symbols are QPSK modulated with symbol rate of 16.7 kbps. The channel is assumed to be time-invariant throughout the duration of each interleaving block. In addition, full channel state information and perfect synchronization are assumed throughout this work. Each simulation run consists of 2.56×10^6 independent OFDM symbols. The multipath fading channel model considered in this work corresponds to a Rayleigh frequencyselective channel with normalized delays of [0, 1, 2, 3, 4] samples and average gains [0.35, 0.25, 0.18, 0.12, 0.1]. The channel coefficients are changed randomly in each N OFDM symbol. The IN is modeled as GBG process, where the bursts position is uniformly distributed within the OFDM symbol period. Unless it is specified otherwise, the burst gating factor probability p = 0.01, the burst width $\varkappa = (N + \bar{N})/2$, the signal-to-IN ratio SIR = -20 dB and SNR $\triangleq \gamma$.



Figure 6.3: BER versus the threshold \mathcal{T}_1 for different values of \mathcal{T}_2 and SNR.

6.8.1 AWGN channels

First, the performance of the system in the AWGN channel is studied. Fig. 6.2 presents the BER versus the threshold \mathcal{T}_1 for sample based blanking, where the results show that \mathcal{T}_1 is almost independent of the SNR, particularly at high SNRs. However, the BER is highly dependent on \mathcal{T}_1 , which implies that the signal power should be estimated accurately to estimate and set \mathcal{T}_1 to minimize the BER. The results in this work are obtained using $\mathcal{T}_1=3$, unless it is specified otherwise.

For symbol blanking, the optimum values of \mathcal{T}_1 and \mathcal{T}_2 are required. However, it is worth noting that there are several combinations of \mathcal{T}_1 and \mathcal{T}_2 that provide minimum BER. However, the general trend is that \mathcal{T}_1 and \mathcal{T}_2 are inversely proportional as shown in Fig. 6.3. As it can be noted from this figure, the BER is almost independent for a wide range of \mathcal{T}_1 , which relaxes the threshold selection process.

Fig. 6.4 shows the BER of the TDI and conventional OFDM with blind and



Figure 6.4: BER performance of variants of the TDI system versus standard OFDM.

non-blind sample and symbol blanking. The non-blind blanking is performed by assuming that the receiver has perfect impulsive noise state information (NSI), and hence only the contaminated samples will be blanked. As it can be noted from the figure, conventional OFDM with NSI and blind sample blanking suffers from a severe BER degradation, which is interpreted as an error floor at $\sim 10^{-3}$. On the contrary, the TDI degradation is less than 2 dB as compared to the IN-free (AWGN) case at BER of 10^{-5} . Furthermore, the BER of the proposed symbol blanking is very close to the sample based blanking with NSI where the degradation is less than 1 dB.

6.8.2 Fading channels

To demonstrate the effect of TDI on the decision variables \mathbf{r} at the FFT output, the magnitude of the noise term in each subcarrier is presented for a particular channel realization as shown in Fig. 6.5. The aggregate noise and interference (AGNI) output



Figure 6.5: Noise enhancement of the TDI and OFDM systems with ZF and MMSE equalizers for SNR = 15 dB.

at the kth subcarrier is compute as $|r_k - d_k|^2$. The AGNI is selected because it clearly reveals the noise enhancement problem caused by the frequency-selective fading. As it can be observed from the figure, the channel has two nulls at subcarriers 50 to 56 and 64 to 69. As expected, the division by the relevant H_k in conventional OFDM will cause severe noise enhancement for the subcarriers with deep fading as shown in Fig. 6.5. The ZF equalizer and the last FFT mixes the channel frequency response (CFR) of all subcarriers, which mainly will deteriorate the BER performance as the enhanced noise is distributed over all subcarriers. Finally, the MMSE does not suffer from any noticeable noise enhancement problem due to the bias added to H_k before the division process, which proves that the CFR mixing process is advantageous and will cause BER improvement.

Fig. 6.6 presents the effective SINR of the TDI and OFDM systems versus the SNR γ , which measures the SNR degradation caused by both, fading and blanking.



Figure 6.6: Effective SINR of TDI and OFDM for different values of ϵ .

The immunity of the TDI to fading can be observed from the $\epsilon = 0$ case where the SINR of the TDI is about 5 dB higher than the conventional OFDM. It is evident that OFDM suffers from losing some samples, where $\epsilon > 0$, more than TDI which leads to the increase in the SINR gap to approximately 7 and 7.5 dB for $\epsilon = 2$ and 1, respectively. For $\epsilon \geq 3$, the effective SINR becomes very low, which decreases the SINR difference to about 4 dB. In other words, despite OFDM that is degraded even with $\epsilon = 1$, TDI can successfully resist the adverse effect of losing samples up to $\epsilon = 3$, where its slope becomes equal to that of OFDM.

The BER performance of IN-free ($\epsilon = 0$) TDI, WHT-OFDM and conventional OFDM systems using ZF and MMSE equalizers is given in Fig. 6.7. As it can be noted from the figure, the lowest BER is achieved by the TDI/WHT using MMSE, whereas the TDI/WHT ZF gives the worst BER. The OFDM inherent fading-resistance gives it a mediocre performance. Since OFDM transmits the signal in orthogonal narrow



Figure 6.7: BER of the TDI and conventional OFDM using ZF and MMSE equalizers.

frequency band channels, both equalizers give the same BER; a result which is expected. Moreover, the simulation results in this figure corroborate with the analysis, which indicates that the TDI offers frequency diversity gain that is equivalent to the WHT-OFDM using MMSE. As expected, the ZF suffers from noise enhancement problem which can be observed in all ZF equalizers.

Fig. 6.8 shows the BER versus \mathcal{T}_1 using different values of \mathcal{T}_2 and SNR. It can be noted from the figure that there is no unique optimum $(\mathcal{T}_1, \mathcal{T}_2)$ that should be used. Moreover, once \mathcal{T}_2 is selected, \mathcal{T}_1 can be selected without a major concern about its accuracy due to the BER plateau shown in Fig. 6.8. Therefore, the thresholds selection process is relaxed as compared to the sample-by-sample blanking. Besides, it is interesting to note that analytical BER can be achieved regardless of the SNR.

Fig. 6.9 shows the joint effect of IN and fading, and compares the BER of TDI, OFDM and WHT-OFDM using ZF equalization. The performance of OFDM



Figure 6.8: BER versus the threshold \mathcal{T}_1 for different values of \mathcal{T}_2 and SNR in fading channels.

without IN is shown in the figure as the baseline. The blind blanking is performed by comparing the received sample to the optimum thresholds, while non-blind is performed by assuming that the receiver has perfect IN state information (NSI). In general, OFDM systems offer poor performance in IN channels even when blanking is invoked and the perfect NSI is available. The simulation results show that OFDM exhibits an error floor greater than 10^{-3} , which persists even with NSI. Similar to the conventional OFDM, the WHT-OFDM system does not offer any noticeable immunity against IN. Although the TDI-ZF does not offer BER improvement in frequencyselective fading channels, it managed to mitigate the IN reasonably, where the BER converges to the OFDM BER with $\epsilon = 0$. Furthermore, the blind symbol blanking applied for TDI offers BER that is approximately equal to the ideal blanking with NSI. In all other cases, a severe BER degradation is observed.

The BER of TDI-MMSE is compared to OFDM and WHT-OFDM in Fig.



Figure 6.9: BER using ZF equalizer in the presence of IN.

6.10, where the results clearly show that the TDI significantly outperforms the other considered systems. In addition to the high robustness in frequency-selective fading channels, the TDI can effectively combat the degradation caused by IN bursts with less than 1 dB degradation from the IN-free case at BER of 10^{-6} . Both, the OFDM and WHT-OFDM exhibit high errors at SNR of about 35 dB. However, the WHT-OFDM slightly outperforms the conventional OFDM due to its robustness to fading. Comparing the BER for the blind and NSI blanking in TDI shows that blind symbol blanking can be performed with high accuracy. This figure also shows that the semi-analytical and simulation results are almost identical.

The effect of the IN bursts arrival probability is given in Fig. 6.11. It can be noted from the figure that the TDI BER degradation for $p \leq 0.02$ is negligible, while an error floor is observed at $BER \approx 10^{-4}$ for p = 0.05. It is worth noting that $P(\epsilon \geq 1)|_{p=0.02} \approx 0.92$, which corresponds to a severe IN channel where every block



Figure 6.10: BER using MMSE equalizer in the presence of IN.

of N OFDM symbols will be mainly hit by one more IN bursts. For the p = 0.05 and 0.1 cases, $P(\epsilon \ge 3)|_{p=0.05} \approx 0.96$ and $P(\epsilon \ge 3)|_{p=0.1} \approx 1$, respectively. Therefore, the TDI can offer a superior BER performance even in severe IN channels where p = 0.05. The p = 0.1 case is too extreme, and hence it will be difficult to have reliable communications in such scenarios.

It is worth mentioning that all of the results demonstrated in this section, assume some kind of IN blanking technique. For the case of sample blanking with full NSI, it is obvious that modifying the IN amplitude does not change the results, since all of the corrupted samples are ideally detected and blanked. However, increasing the amplitude of the IN will help the blind sample blanking method to correctly detect the corrupted samples. This will result in closing the BER gap between blind and full NSI cases. In the proposed symbol blanking method, since the BER performance of blind and full NSI cases matches, increasing the IN amplitude does not change



Figure 6.11: BER of the TDI for different gating probability p.

the performance. On the other hand, decreasing the amplitude obviously makes the detection harder for the blind case. Nevertheless, through adjusting the combination of \mathcal{T}_1 and \mathcal{T}_2 , and sacrificing the capability of detecting short IN bursts, it might be possible to achieve the theoretical BER in some cases.

6.9 Chapter Summary

This work presented a novel technique for joint mitigation of impulsive noise (IN) and multipath fading effects in broadband communication systems with OFDM modulation. The proposed TDI technique, interleaves the samples of several OFDM symbols after the IFFT and deinterleaves them before the final FFT stage. The performance of the TDI in presence of IN has been analyzed in multipath fading channels where ZF and MMSE equalizers are employed to compensate the frequency-selectivity effects. The performance of the proposed system is evaluated in terms of BER where closed-form formulae are derived for the SINR using ZF and MMSE equalizers. The obtained analytical and simulation results show that the TDI can effectively combat the effects of heavily distributed IN and the frequency-selectivity of the channel. The BER improvement obtained in frequency-selective channels is due to the inherent frequency diversity of the TDI, which is equivalent to Walsh-Hadamard precoded OFDM systems. Moreover, the TDI has time diversity, which was effectively used to reduce the BER degradation caused by the IN.

References

- M. Mirahmadi, A. Al-Dweik, and A. Shami, "Low complexity BER reduction technique for OFDM systems over impulsive noise channels," *IEEE Trans. Commun.*, submitted for publication.
- [2] Radio broadcasting systems; Digital Audio Broadcasting (DAB) to mobile, portable and fixed receivers, ETS Std. 300 401, 1995.
- [3] Digital video broadcasting (DVB); Framing structure, channel coding and modulation for digital terrestrial television, ETSI Std. EN 300 744 v1.6.1, 2008.
- [4] IEEE Standard for Local and metropolitan area networks Part 16: Air Interface for Broadband Wireless Access Systems Amendment 3: Advanced Air Interface, IEEE Std. 802.16m, 2011.
- [5] LTE; Evolved Universal Terrestrial Radio Access (E-UTRA) and Evolved Universal Terrestrial Radio Access Network (E-UTRAN); Overall description, 3GPP Std. TS 36.300, 2011.
- [6] Digital Video Broadcasting (DVB); Frame structure channel coding and modulation for a second generation digital transmission system for cable systems (DVB-C2), ETSI Std. EN 302 769, 2010.
- [7] Asymmetric digital subscriber line transceivers 2 (ADSL2) Extended bandwidth (ADSL2plus), ITU-T Std. G.992.5, 2009.
- [8] IEEE Standard for Broadband over Power Line Networks: Medium Access Control and Physical Layer Specifications, IEEE Std. P.1901, 2010.

- [9] A. Hazmi, J. Rinne, and M. Renfors, "An enhanced impulse burst cancellation method using pilots and soft bits in OFDM based systems," in *IEEE 5th Work-shop Signal Proc. Advances in Wireless Commun.*, Lisbon, Portugal, Jul. 2004, pp. 373–376.
- [10] A. Hazmi, J. Rinne, M. Renfors, J. Vesma, T. Auranen, P. Talmola, and J. Henriksson, "Mitigation techniques for high power and long duration interference in DVB-T/H systems," in *IEEE 18th Int. Symp. Personal, Indoor and Mobile Radio Commun. (PIMRC)*, Athens, Greece, Sep. 2007, pp. 1–5.
- [11] A. Hazmi, J. Rinne, T. Kuusisto, and M. Renfors, "Performance evaluation of symbol synchronization in OFDM systems over impulsive noisy channels," in *IEEE 59th Veh. Tech. Conf. (VTC)*, vol. 3, Milan, Italy, 2004, pp. 1782–1786.
- [12] S. V. Zhidkov, "Analysis and comparison of several simple impulsive noise mitigation schemes for OFDM receivers," *IEEE Trans. Commun.*, vol. 56, no. 1, pp. 5–9, Jan. 2008.
- [13] I. Mann, S. McLaughlin, W. Henkel, R. Kirkby, and T. Kessler, "Impulse generation with appropriate amplitude, length, inter-arrival, and spectral characteristics," *IEEE J. Sel. Areas Commun.*, vol. 20, no. 5, pp. 901–912, Jun. 2002.
- [14] S. V. Zhidkov, "Impulsive noise suppression in OFDM-based communication systems," *IEEE Trans. Consum. Electron.*, vol. 49, no. 4, pp. 944–948, Nov. 2003.
- [15] R. Pighi, M. Franceschini, G. Ferrari, and R. Raheli, "Fundamental performance limits of communications systems impaired by impulse noise," *IEEE Trans. Commun.*, vol. 57, no. 1, pp. 171–182, Jan. 2009.
- [16] F. Abdelkefi, P. Duhamel, and F. Alberge, "Impulsive noise cancellation in multicarrier transmission," *IEEE Trans. Commun.*, vol. 53, no. 1, pp. 94–106, Jan. 2005.
- [17] S. V. Zhidkov, "Performance analysis and optimization of OFDM receiver with blanking nonlinearity in impulsive noise environment," *IEEE Trans. Veh. Technol.*, vol. 55, no. 1, pp. 234–242, Jan. 2006.
- [18] M. Zimmermann and K. Dostert, "Analysis and modeling of impulsive noise in broad-band powerline communications," *IEEE Trans. Electromagn. Compat.*, vol. 44, no. 1, pp. 249–258, Feb. 2002.
- [19] Y. Ma, P. So, and E. Gunawan, "Performance analysis of OFDM systems for broadband power line communications under impulsive noise and multipath effects," *IEEE Trans. Power Del.*, vol. 20, no. 2, pp. 674–682, Apr. 2005.
- [20] A. Al-Dweik, A. Hazmi, B. Sharif, and C. Tsimenidis, "Efficient interleaving technique for OFDM system over impulsive noise channels," in *IEEE 21st Int. Symp. Personal Indoor and Mobile Radio Commun. (PIMRC)*, Istanbul, Turkey, Sep. 2010, pp. 167–171.
- [21] S. Nayyef, A. Al-Dweik, A. Hazmi, B. Sharif, and C. Tsimenidis, "An efficient technique for OFDM systems over fading channels impaired by impulsive noise," in *IEEE Veh. Tech. Conf. (VTC)*, Quebec City, Canada, Sep. 2012, pp. 1–5.
- [22] S. Wang, S. Zhu, and G. Zhang, "A Walsh-Hadamard coded spectral efficient full frequency diversity OFDM system," *IEEE Trans. Commun.*, vol. 58, no. 1, pp. 28–34, Jan. 2010.
- [23] M. S. Ahmed, S. Boussakta, B. S. Sharif, and C. C. Tsimenidis, "OFDM based on low complexity transform to increase multipath resilience and reduce PAPR," *IEEE Trans. Signal Process.*, vol. 59, no. 12, pp. 5994–6007, Dec. 2011.
- [24] Y. G. Li and G. L. Stuber, Orthogonal frequency division multiplexing for wireless communications. Springer, 2006.
- [25] A. Al-Dweik, S. Younis, A. Hazmi, C. Tsimenidis, and B. Sharif, "Efficient OFDM symbol timing estimator using power difference measurements," *IEEE Trans. Veh. Commun.*, vol. 61, no. 2, pp. 509–520, Feb. 2012.

- [26] D. Middleton, "Canonical and quasi-canonical probability models of class a interference," *IEEE Trans. Electromagn. Compat.*, no. 2, pp. 76–106, May 1983.
- [27] M. Ghosh, "Analysis of the effect of impulse noise on multicarrier and single carrier QAM systems," *IEEE Trans. Commun.*, vol. 44, no. 2, pp. 145–147, Feb. 1996.
- [28] E. Al-Dalakta, A. Al-Dweik, A. Hazmi, C. Tsimenidis, and B. Sharif, "PAPR reduction scheme using maximum cross correlation," *IEEE Commun. Lett.*, vol. 16, pp. 2032–2035, Dec. 2012.
- [29] C.-H. Yih, "Iterative interference cancellation for OFDM signals with blanking nonlinearity in impulsive noise channels," *IEEE Signal Process. Lett.*, vol. 19, no. 3, pp. 147–150, Mar. 2012.
- [30] H. Eghbali, S. Muhaidat, and N. Al-Dhahir, "A novel receiver design for singlecarrier frequency domain equalization in broadband wireless networks with amplify-and-forward relaying," *IEEE Trans. Wireless Commun.*, vol. 10, no. 3, pp. 721–727, Jan. 2011.
- [31] M. L. McCloud, "Analysis and design of short block OFDM spreading matrices for use on multipath fading channels," *IEEE Trans. Commun.*, vol. 53, no. 4, pp. 656–665, Apr. 2005.
- [32] Y.-P. Lin and S.-M. Phoong, "BER minimized OFDM systems with channel independent precoders," *IEEE Trans. Signal Process.*, vol. 51, no. 9, pp. 2369– 2380, 2003.
- [33] V. Witkovský, "Computing the distribution of a linear combination of inverted gamma variables," *Kybernetika*, vol. 37, no. 1, pp. 79–90, 2001.
- [34] Z. Wang and G. B. Giannakis, "A simple and general parameterization quantifying performance in fading channels," *IEEE Trans. Commun.*, vol. 51, no. 8, pp. 1389–1398, Aug. 2003.

[35] Y. Jiang, M. K. Varanasi, and J. Li, "Performance analysis of ZF and MMSE equalizers for MIMO systems: an in-depth study of the high SNR regime," *IEEE Trans. Inf. Theory*, vol. 57, no. 4, pp. 2008–2026, Apr. 2011.

Chapter 7 Conclusion and Future Works

In the last decades, information and communication technology substantially changed the lifestyle of people. Inexpensive computing resources and broadband access are actively elevating the distribution of data, information, and knowledge to a level that was deemed unimaginable a while ago. Among the most influential driving forces responsible for this trend are inexpensive and high speed access networks that offer high quality services to users. In this thesis, we tried to advance the knowledge related to several parts of access networks.

7.1 Improving Quality of Service in Hybrid Optical Wireless Network

As the first step in our endeavor, improving the quality of service (QoS) in a hybrid optical-wireless network is targeted. As a matter of fact, due to the low maintenance costs of passive optical networks (PON), a large percentage of today's broadband networks employ PONs in their distribution and core networks. Observing this, we improved the QoS of PONs that are connected to wireless transmitters, such as WiMAX BSs. First, a detailed design for the integrated ONU-BS envisioned in the pioneering work of Shen *et. al.* [1] was proposed. Then, a technique to intelligently use the information available in the wireless domain to predict the incoming traffic was proposed. Finally, to take advantage of the traffic prediction, two well-known dynamic bandwidth assignment (DBA) algorithms, namely CB-IPACT and Excess Distribution (ED) [2,3], were modified with the proposed technique.

The modified and original algorithms were compared via extensive simulations. The results show that without sacrificing the throughput, the delay guarantee bound for EF flows in a hybrid EPON-WiMAX network, can be reduced from 3 ms to 1.3 ms, i.e., by more than 50%, in high load scenarios. This significant improvement comes with a slight increase in the delay in low and medium load conditions. The same trend can be seen for lower priority packets. Moreover, since all of the wireless and optical networks share the design principles utilized in the proposed technique, we envision that same results will be achieved if different wireless and/or optical technology is used.

The effect of various parameters, such as load, EPON cycle length and WiMAX frame length on the performance of the modified DBA algorithms are studied. Through the analysis, it was found that optical networks with longer cycle length than wireless frames, cannot be effectively integrated with the wireless domain via the proposed technique. In addition, the performance of the algorithms in a realistic scenario, where wireless BSs are not the only source of traffic and the distribution of traffic between the sources follows realistic conditions, was also studied. The results prove that the same improvement will be seen in high load sources, and also the proposed algorithm is applicable to real networks.

As an interesting extension of this work, one can study the effects of different wireless scheduler algorithms. Moreover, the information available in the optical part can also be utilized with the same approach used in this thesis to improve the performance of wireless scheduler. In addition, the implementation of the proposed algorithm on real ONUs and BSs using their standard interface can open a vast opportunity for improving hybrid networks.

7.2 Managing Femtocell Interference in Pre-planned Cellular Network

In the second part, we turned our attention to the problem of employing femtocells in legacy cellular network and managing the interference caused by that. We observed, as significant as it is, the complex effects of building structure on femtocell interference has never been accurately captured in the literature. Hence, a procedural building generation algorithm was developed to randomly generates realistic floor plans, which was later used to develop a database of building features. The propagation of FBSs' signal in the generated floor plans was estimated and a Gaussian-Mixture distribution was proposed to model the femtocell interference. The work also proves that in proximity of the buildings that enclose a FBS, the interference power may significantly vary depending on the position of the receiver and the transmitter.

In Chapter 5, a composite shadowing/multipath fading channel model was proposed to capture the dynamic effects of the indoor-outdoor channel. The interference caused by a femtocell signal that traversed through this channel, on a macrocell user in proximity of the house was mathematically analyzed. The proposed composite model was then used to analytically derive the received SIR, and outage probability. The numerical results show that without careful placement of the femtocells or an efficient interference mitigation technique, the outage probability is very high in the proximity of the buildings. It was also shown that increasing the number of femtocells deteriorates the outage even if the total power received from femtocells is maintained at the same level. The results also establish that if employed carefully, the building structure itself can be used to decrease the interference to some degree.

We are currently working with our collaborators to measure the effects predicted in this study. The acquired data will enable us to fine tune the proposed model. Besides, in this thesis, we focused on suburban dwellings as it is the main type of building to install femtocells in. However, the same analysis should be performed for large buildings, like office or public buildings. Moreover, the proposed model is not restricted to the one type of interference mentioned in this thesis, rather it is applicable to other types of interference as well. As an instance, the femtocell interference on the same-tier femtocell users can be studied with the proposed model. In this thesis, we calculated the effect of up to four FBSs simultaneously. The results reveal that increasing the number of femtocells results in a maximum acceptable density of femtocells which can be used as a rule of thumb by operators to design their two-tier femtocell/macrocell network.

7.3 Mitigating Impulsive Noise in Frequency-Selective Fading Channels

Chapter 6 deals with the problem of impulsive noise (IN) in OFDM-based systems. We modeled the degradation caused by simultaneous fading and impulsive noise when blanking is used. In addition, a novel time-domain interleaving (TDI) technique was proposed to mitigate the effect of IN. The analyses proved that the threshold based blanking [4] typically used in to mitigate the effect of IN is incapable of effectively suppressing IN in fading environments. Therefore, we enhanced our technique with a symbol-based blanking that removes symbols that are hit by IN without major BER degradation. The performance of this technique was analyzed in AWGN and fading channels, using ZF and MMSE equalizers.

We are also in the process of studying the proposed technique in dynamic environments. Preliminary results show that the technique can be used in slow-moving environments with high IN. In future, it is possible to suggest an industrial wireless LAN that is resilient to IN and can work appropriately in industrial conditions by adjusting the parameters in the proposed technique. Furthermore, any OFDM-based system can use similar technique to combat the effects of losing one or more samples due multiple access noise, jamming, etc. This can be a potential topic for future researches in this area.

References

- G. Shen, R. Tucker, and C.-J. Chae, "Fixed mobile convergence architectures for broadband access: Integration of EPON and WiMAX," *IEEE Commun. Mag.*, vol. 45, no. 8, pp. 44–50, Aug. 2007.
- [2] G. Kramer, B. Mukherjee, and G. Pesavento, "IPACT: A dynamic protocol for an Ethernet PON (EPON)," *IEEE Commun. Mag.*, vol. 40, no. 2, pp. 74–80, Feb. 2002.
- [3] C. Assi, Y. Ye, S. Dixit, and M. Ali, "Dynamic bandwidth allocation for qualityof-service over Ethernet PONs," *IEEE J. Sel. Areas Commun.*, vol. 21, no. 9, pp. 1467 – 1477, Nov. 2003.
- [4] S. V. Zhidkov, "Performance analysis and optimization of OFDM receiver with blanking nonlinearity in impulsive noise environment," *IEEE Trans. Veh. Technol.*, vol. 55, no. 1, pp. 234–242, Jan. 2006.

Appendices

Appendix A Blanking Distortion in OFDM Systems

Assume that $\mathbf{x} = \mathcal{F}^{-1}{\mathbf{d}}$. Therefore,

$$x_n = \kappa \sum_{i=0}^{N-1} d_i e^{\omega i n}.$$
 (A.1)

Given that ϵ samples of \mathbf{x} are zeroed, calculating $\mathbf{s} = \mathcal{F}\{\mathbf{y}\}$, where $\mathbf{y} = \mathbb{B}\mathbf{x}$, gives

$$s_k|\epsilon = \kappa \sum_{n=0}^{N-1} y_n e^{-\omega nk} = \kappa \sum_{n=\epsilon}^{N-1} x_n e^{-\omega nk}$$
(A.2)

Inserting (A.2) in (A.1), gives

$$s_{k}|\epsilon = \kappa^{2} \sum_{n=\epsilon}^{N-1} \sum_{i=0}^{N-1} d_{i}e^{\omega i n}e^{-\omega n k}$$
$$= \beta d_{k} + \kappa^{2} \sum_{n=\epsilon}^{N-1} \sum_{\substack{i=0\\i\neq k}}^{N-1} d_{i}e^{\omega n(i-k)}$$
(A.3)

where $\beta = \frac{N-\epsilon}{N}$. With the change of the summations order, it can be rewritten as

$$= \beta d_k + \kappa^2 \sum_{\substack{i=0\\i \neq k}}^{N-1} d_i \sum_{n=\epsilon}^{N-1} e^{\omega n(i-k)}.$$
 (A.4)

The last summation can be simplified using the identity

$$\sum_{n=\epsilon}^{N-1} e^{\omega n(i-k)} = \frac{e^{-2j\pi (-k-iN+kN)\kappa^2} - e^{-2j\pi (-k-\epsilon i+\epsilon k)\kappa^2}}{e^{2j\pi i\kappa^2} - e^{2j\pi k\kappa^2}}$$

and taking $e^{\frac{2j\pi k}{N}}$ as a common factor, which results in

$$\sum_{n=\epsilon}^{N-1} e^{\omega n(i-k)} = \frac{e^{2j\pi(i-k)} - e^{2j\pi\epsilon(i-k)\kappa^2}}{e^{2j\pi(i-k)\kappa^2} - 1}.$$
 (A.5)

Note that $e^{2j\pi(i-k)} = 1$. Thus

$$\sum_{n=\epsilon}^{N-1} e^{\omega n(i-k)} = \frac{1 - e^{2j\pi\epsilon(i-k)\kappa^2}}{e^{2j\pi(i-k)\kappa^2} - 1}$$
(A.6)

Using Euler formula, and after some algebric manipulations, we have

$$1 - e^{2jm} = 2\sin m \left(\sin m - j\cos m\right)$$

where $m = \frac{\pi \epsilon (i-k)}{N}$. Similarly, letting $\frac{\pi (i-k)}{N} = u$, gives $e^{2ju} - 1 = 2 \sin u \left[-\sin u + j \cos u \right]$. Therefore,

$$\sum_{n=\epsilon}^{N-1} e^{\omega n(i-k)} = \frac{\sin m}{\sin u} \frac{\sin m - j \cos m}{-\sin u + j \cos u}$$
$$= -\frac{\sin m}{\sin u} e^{-j(m+u)}$$
$$= \frac{\sin \left[\pi \epsilon \kappa^2 (k-i)\right]}{\sin \left[\pi \kappa^2 (i-k)\right]} e^{-j\pi \kappa^2 (\epsilon+1)(i-k)}$$

Finally,

$$s_k|\epsilon = \beta d_k + \kappa^2 \sum_{\substack{i=0\\i\neq k}}^{N-1} d_i \frac{\sin\left[\pi\epsilon\kappa^2(k-i)\right]}{\sin\left[\pi\kappa^2(i-k)\right]} e^{-j\pi\kappa^2(\epsilon+1)(i-k)}$$
(A.7)

The second term corresponds with the error caused by zeroing some samples of \mathbf{x} and its mean is

$$E\left\{\kappa^{2}\sum_{\substack{i=0\\i\neq k}}^{N-1} d_{i}\frac{\sin\left[\pi\epsilon\kappa^{2}(k-i)\right]}{\sin\left[\pi\kappa^{2}(i-k)\right]}e^{-j\pi\kappa^{2}(\epsilon+1)(i-k)}\right\} = 0.$$
 (A.8)

The variance of this error can be computed as

$$\sigma_k^2 = E \left\{ \left| \kappa^2 \sum_{\substack{i=0\\i\neq k}}^{N-1} d_i \frac{\sin\left[\pi\epsilon\kappa^2(k-i)\right]}{\sin\left[\pi\kappa^2(i-k)\right]} e^{-j\pi\kappa^2(\epsilon+1)(i-k)} \right|^2 \right\}$$
$$= \kappa^4 E \left\{ \sum_{\substack{i=0\\i\neq k}}^{N-1} \sum_{\substack{v=0\\v\neq k}}^{N-1} E \left\{ d_i d_v^* \right\} \frac{\sin\left[\pi\epsilon\kappa^2(k-i)\right]}{\sin\left[\pi\kappa^2(i-k)\right]} \frac{\sin\left[\pi\epsilon\kappa^2(k-v)\right]}{\sin\left[\pi\kappa^2(v-k)\right]} e^{j\pi\kappa^2(\epsilon+1)(v-i)} \right\}$$
(A.9)

Assuming that the data samples d_i are independent and identically distributed with $E\{d_id_k\} = 0 \ \forall i \neq k$ and variance σ_d^2 , then

$$\sigma_k^2 = \kappa^4 \sum_{\substack{i=0\\i\neq k}}^{N-1} E\left\{d_i d_i^*\right\} \frac{\sin\left[\pi \epsilon \kappa^2 (k-i)\right]}{\sin\left[\pi \kappa^2 (i-k)\right]} \frac{\sin\left[\pi \epsilon \kappa^2 (k-i)\right]}{\sin\left[\pi \kappa^2 (i-k)\right]} = \frac{\sigma_d^2}{N^2} \sum_{\substack{i=0\\i\neq k}}^{N-1} \frac{\sin^2\left[\pi \epsilon \kappa^2 (k-i)\right]}{\sin^2\left[\pi \kappa^2 (i-k)\right]}$$
(A.10)

Appendix B Samples of Automatically Generated Floor Plans

Several automatically generated floor plans by PBG algorithm developed in Chapter 3 are illustrated in the following figures to show the resemblance to real floor plans and also their variety.





Appendix C Curriculum Vitae

Maysam MirAhmadi

Innovation Center for Information Engineering (iCIE), Western University (formerly, The University of Western Ontario), London, Ontario, N6A 5B9, Canada.

Education

2009 - present

Doctor of Philosophy

Electrical and Computer Engineering Dept., Western University (formerly, The University of Western Ontario)

2005 - 2007

Master of Science

Electrical Engineering Dept., Amirkabir University of Technology (Tehran Polytechnic)

2001 - 2005

Bachelor of Science (Honors)

Electrical Engineering Dept., Amirkabir University of Technology (Tehran Polytechnic)

Work & Research Experiences

2009 - present

Research Assistant

Electrical and Computer Engineering Dept., Western University (formerly, The University of Western Ontario)

2007 - 2009

Technical Manager & Senior Designer

Kavoshcom R&D Group

2007

Research Assistant

Electrical Engineering Dept., Amirkabir University of Technology (Tehran Polytechnic)

2006 - 2007

Hardware Design Engineer

Kavoshcom R&D Group

Honors and Awards

- Best paper award at the 7th IEEE International Wireless Communication and Mobile Computing Conference (IEEE IWCMC'11).
- Best paper award runner-up at the IEEE International Communication Conference in 2012 (IEEE ICC'12).
- Robocup Rescue 2002 championship with Kavosh Team, Fukuoka, Japan, 2002.

Teaching Experiences

2009 - present

Teaching Assistant

Electrical and Computer Engineering Dept., Western University (formerly, The University of Western Ontario)

2007 - 2009

Lecturer

Computer Engineering Dept., Azad University, Tehran, Iran

2007

Instructor

Data Processing Company (DPCO)

2004

Laboratory Instructor

Amirkabir Univ. of Tech. (Tehran Polytechnic)

Publications

Refereed Journals

- M. Mirahmadi, A. Al-Dweik, A. Shami, "Indoor-Outdoor Channel Characterization and its Application to Interference Analyses of Femtocells", *IEEE Transaction on Wireless Communications*, accepted subject to revision, 2012.
- M. Mirahmadi, A. Shami, "Traffic-prediction-assisted Dynamic Bandwidth Assignment for Hybrid Optical Wireless Networks", *Computer Networks*, Elsevier, 2012.
- M. Mirahmadi, A. Al-Dweik, A. Shami, "BER Reduction of OFDM Based Broadband Communication Systems over Multipath Channels with Impulsive Noise", submitted to *IEEE Transaction on Communications*, 2013.
- S. Jabbehdari, A. Movaghar, M. Mirahmadi, "Fairness of Two Level Hierarchical Scheduling Scheme for High-Speed Networks by Using SFQ and E-LOSS", *International Journal of Information Technology (IJIT)*, 2004.

Refereed Conference Proceedings

- M. Mirahmadi, A. Shami, A. Al-Dweik, "A Building Architecture Model for Predicting Femtocell Interference in Next-Generation Networks", *IEEE International Communication Conference (ICC)*, 2012.
- M. Mirahmadi, A. Shami, "Traffic-prediction-assisted Dynamic Bandwidth Assignment for Hybrid Wireless Optical Networks", *IEEE International Wireless Communications and Mobile Computing Conference (IWCMC)*, 2011, Istanbul, Turkey.
- S. Jabbehdari, A. Movaghar, M. Mirahmadi, "Fairness of Two Level Hierarchical Scheduling Scheme for High-Speed Networks by Using SFQ and E-LOSS", *1st International Conference on Information Technology (ICIT)*, Istanbul, Turkey, 2004.
- S. Jabbehdari, A. Movaghar, M. Mirahmadi, "A Novel Two Level Hierarchical Scheduling Scheme by Using SFQ and E-LOSS Algorithms", *9th Annual CSI Computer Conference (CSICC)*, Tehran, Iran, 2004.

Other

- M. Mirahmadi, A. Shami, "A Novel Algorithm for Real-time Procedural Generation of Building Floor Plans", Computing Research Repository (CoRR), 2011.
- M. Mirahmadi, H. Taheri, M. Dehghan, "Maximizing Wireless Sensor Network Lifetime via Load Balancing Topology Control", M.Sc. Thesis, 2007.
- M. Mirahmadi, H. Taheri, "Design and Implementation of a Bluetooth Serial Link", B.Sc. Thesis, 2004.