Design and measurement results for cooperative spectrum sharing

Vivek Ashok Bohara

2011

Vivek, A. B. (2011). Design and measurement results for cooperative spectrum sharing.
Doctoral thesis, Nanyang Technological University, Singapore.

<https://hdl.handle.net/10356/46306>

<https://doi.org/10.32657/10356/46306>

**DESIGN AND MEASUREMENT RESULTS FOR COOPERATIVE
SPECTRUM SHARING**

VIVEK ASHOK BOHARA

School of Electrical and Electronic Engineering

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

2011

Acknowledgments

I express my sincere gratitude, regards and thanks to my supervisor, Assistant Professor See Ho Ting for his excellent guidance, useful suggestions and above all continuous encouragement throughout my research work. His unrelenting support and continued belief in my abilities is the single most important factor for the completion of this work. All these years of my PhD he has been a mentor, guardian and older brother who always stood by me, believed in me and taught me very important life lessons that goes beyond getting PhD.

I would like to thank Associate Professors Guan Yong Liang and Law Choi Look for their help, discussions and insightful suggestions during my PhD work. I am grateful to Nanyang Technological University's Start Up Grant and Assistant Professor See Ho Ting for providing me research scholarship and excellent research facilities. I would like to take this opportunity to thank Nanyang Technological University - National Instruments Wireless Research Programme under Assistant Professor See Ho Ting for the measurement equipment. The proof-of-concept demonstrations and testing of proposed algorithms wouldn't have been possible without the measurement equipment.

I am thankful to my research centre members Han Yang, Li Qiang, Harya, Liu Zilong, Thien Than Tun and others for their help and suggestions during the period of my research. I am also thankful to supporting staff in the Positioning and Wireless Technology Center, Joseph, Ooy Mei and Thida for providing all technical support and help. I express my sincere gratitude to my group of friends "12 survivors" and many others whose companionship gave me enthusiasm and kept me sane. I also thank all the people who have directly or indirectly helped in the completion of the thesis.

I would like to thank my reverent parents, dearest siblings and other family members for their boundless and unconditional love. Last but not the least, thanks to my beloved wife Swati for her support, love, care and understanding. My gratitude for them is beyond words.

Dedication

*To my parents for their blessings and encouragement
and
my wife for her love and support.*

Table of Contents

Acknowledgments	ii
Summary	viii
List of Figures	x
List of Tables	xiv
List of Abbreviations	xv
List of Symbols	xx
Chapter 1	
Introduction	1
1.1 Motivations and Objectives	1
1.2 Major contribution of the thesis	4
1.3 Organization of the thesis	6
1.4 Background and Preliminaries	7
1.4.1 Wireless Communication	7
1.4.1.1 Fading	8
1.4.2 Cooperative Communications	11
1.4.2.1 Amplify-and-forward relaying	13
1.4.2.2 Decode-and-forward relaying	13
1.4.2.2.1 Fixed DF relaying	13
1.4.2.2.2 Selective DF relaying	13
1.4.2.3 Outage performance	14
1.4.2.3.1 Outage behavior of direct transmission . .	14
1.4.2.3.2 Outage behavior of AF relaying	15
1.4.2.3.3 Outage behavior of DF relaying	16
1.4.3 Cognitive Radio Communications	18
1.4.3.1 Spectrum scarcity	18

1.4.3.2	Hierarchical spectrum sharing	20
1.4.3.2.1	Interference avoidance	21
1.4.3.2.2	Interference control	22
1.4.3.2.3	Interference mitigation	22
1.4.4	Cooperative and Cognitive Wireless Systems	24
1.4.4.1	Cooperative transmission between secondary systems	25
1.4.4.2	Cooperative transmission between primary and secondary system	26
1.4.4.3	Performance analysis of a CSS protocol	28
1.4.4.3.1	Outage performance of primary system	28
1.4.4.3.2	Outage performance of secondary system	29
1.4.5	Testbeds for Cognitive Radios	31
1.4.5.1	CORNET	33
1.4.5.2	ORBIT	34
1.4.5.3	Emulab	34
1.4.5.4	BWRC cognitive radio testbed	35
1.4.5.5	Issues with existing deployments of CR testbeds	35
1.5	Conclusions	37

Chapter 2

An Orthogonal Spectrum Sharing Scheme for Wireless Sensor Networks		38
2.1	Introduction	38
2.1.1	System Model and Protocol Description	42
2.1.1.1	System model	42
2.1.1.2	Protocol description	44
2.2	Average Received SNR for OSSS	46
2.2.1	Average received SNR of primary system with OSSS	46
2.2.1.1	Phase 1	46
2.2.1.2	Phase 2	46
2.2.2	Average received SNR of secondary system with OSSS	48
2.2.2.1	Phase 1	48
2.2.2.2	Phase 2	48
2.2.3	Channel estimation and other requirements	49
2.3	Average Received SNR for AF with Superposition Coding (AF-SC)	50
2.3.1	Average received SNR of primary system with AF-SC	50
2.3.1.1	Phase 1	50
2.3.1.2	Phase 2	52
2.3.2	Average received SNR of secondary system with AF-SC	53
2.3.2.1	Phase 1	53
2.3.2.2	Phase 2	53

2.3.3	Simulation results and discussion	56
2.4	Conclusion	57

Chapter 3

A	Testbed for Cooperative Spectrum Sharing	59
3.1	Introduction	59
3.2	Design and Implementation	60
3.2.1	Implementation	63
3.2.1.1	Implementation of primary system	63
3.2.1.2	Implementation of secondary system	63
3.2.1.3	OFDM Frame structure	64
3.3	Measurement	65
3.3.1	Protocol Flowchart	65
3.3.2	A Benchmark: PT-PR Retransmission Protocol	68
3.3.3	Measurement Set-Up	69
3.3.3.1	Path loss between the nodes	71
3.4	Measurement Results	71
3.4.1	Qualitative results	71
3.4.2	Quantitative results	73
3.4.2.1	Packet error rate measurements	73
3.4.2.2	Hourly measurements	76
3.5	Discussion and Key Lessons Learned	78
3.6	Conclusion	82

Chapter 4

Analytical Evaluation of Impact of Nonlinear HPA on an OFDM Communication System		83
4.1	Introduction	83
4.1.1	Impact of a nonlinear HPA on a two-tone signal	85
4.2	System Models for Analysis	86
4.2.1	OFDM signal model	86
4.2.2	Model for nonlinear HPA	88
4.3	Characterization of the Received Signal	89
4.4	Average Symbol Error Rate in Rayleigh Fading Channel	91
4.5	Simulation Results	92
4.5.1	Results with HPA modeled as a MPM with delay taps . . .	92
4.5.2	Results with HPA modeled as a Wiener-Hammerstein (W-H) model	96
4.6	Conclusion	102

Chapter 5

Conclusions and Future Work	103
5.1 Future work	105

Appendix A	
Derivation for Average SNR of primary system with OSSS	106
Appendix B	
Derivation of analytical results	108
B.1 Canonical Decomposition of Received Signal	108
B.1.1 Nonlinear noise component is uncorrelated with the desired data symbol	109
B.1.2 Modeling of nonlinear noise component as a zero mean Gaus- sian process	109
B.1.3 Derivation for complex attenuation component	111
B.2 A Method to Extract Coefficients for a HPA	112
Appendix C	
Author's Publications	116
Bibliography	119

Summary

Scarcity of radio spectrum and inherent inefficiency of current spectrum allocation policies have spurred much research for alternative spectrum access techniques giving birth to the notion of cognitive radios. Recently, cooperative spectrum sharing (CSS) where two wireless systems operate over the same spectrum band albeit with different priorities has been proposed as a viable framework for cognitive radio. The primary system, comprising of a primary transmitter (PT) and primary receiver (PR), supports the relaying functionality. The secondary system, comprising of a secondary transmitter (ST) and secondary receiver (SR), operates on a secondary basis with the guarantee that its operation does not affect the primary system performance. However, most of the proposed CSS protocols are interference limited and the performance of the systems are limited by the amount of interference from one system to another. Consequently, there is an inherent trade-off between the achievable performance of the primary and secondary systems. In this thesis, we try to resolve the above issue by proposing an interference-free CSS protocol known as orthogonal spectrum sharing scheme (OSSS), which alleviates the interference from the primary system to secondary system. The performance of OSSS has been demonstrated through simulation and analytical results.

Another issue related to CSS protocols is the lack of measurement results to demonstrate their performance in a realistic environment. Hence, how much performance enhancement CSS can bring in a real wireless environment is still an

open question. We try to answer this question by designing and developing a testbed for proof-of-concept demonstration and performance assessment of CSS protocols. The testbed is programmed to follow the OFDM standards in IEEE 802.11a. The performance of the testbed has been validated by obtaining both quantitative as well as qualitative results. Quantitative results are obtained by measuring the packet error rates for both primary and secondary systems whereas qualitative results are shown by utilizing a CSS protocol to successfully transmit two different images from PT to PR and ST to SR respectively. The spectrum access probability for the secondary system is also measured.

Apart from the above, we also provide a theoretical format to analytically evaluate the back-off required in a nonlinear HPA while operating an OFDM based communication system. Thus the HPA can be operated with sufficient back-off so that nonlinear distortions due to HPA will have minimal impact on the obtained measurement results. As shown later in this thesis, an improper selection of back-off has considerable impact on the end-to-end symbol error rate (SER) performance of an OFDM based communication system.

List of Figures

1.1	Spectrum allocation chart in Singapore [1]	2
1.2	Band by band average spectrum occupancy in Singapore [2]	2
1.3	Variation of signal strength vs distance	8
1.4	Different classes of small scale fading	9
1.5	BER for BPSK modulation in Rayleigh fading channel	11
1.6	Cooperative communication network	12
1.7	A two-phase relay network	12
1.8	AF and DF relaying	14
1.9	Key elements of cognitive radio	19
1.10	Cognitive radio network	20
1.11	Hierarchical spectrum sharing [50]	21
1.12	Interference avoidance	22
1.13	Interference control	23
1.14	Interference mitigation	23
1.15	Cooperative spectrum sensing network	26
1.16	Cooperative transmission between primary and secondary system .	27
2.1	OSSS: 1st transmission phase	42
2.2	OSSS: 2nd transmission phase	43
2.3	Protocol flowchart	45

2.4	AF-SC: 1st transmission phase.	51
2.5	System configuration for simulation.	54
2.6	Average received SNR of primary transmission for various values of $\frac{P_s}{\sigma^2}$ for OSSS, AF-SC and direct transmission with ARQ. Theoretical and simulation values are reported for $\overline{\text{SNR}}_p$ and $\overline{\text{SNR}}_{\text{MRC}}$, whereas only simulation values are reported for $\overline{\text{SNR}}_p^{\text{AF-SC}}$	55
2.7	Average received SNR of secondary transmission for various values of $\frac{P_s}{\sigma^2}$ for OSSS and AF-SC.	58
3.1	System model for cooperative spectrum sharing	61
3.2	NI PXIe Hardware (Transmitter and Receiver)	64
3.3	Architecture of ST and SR	65
3.4	Frame structure	66
3.5	Measurement flowchart	67
3.6	PT-PR retransmission protocol flowchart	69
3.7	Floor plan of measurement environment.	70
3.8	Measurement environment	72
3.9	Primary system transmit image	72
3.10	Secondary system transmit image	73
3.11	Image received at PR	74
3.12	Image at SR after interference cancellation	75
3.13	Packet error rate for the primary system.	76
3.14	Spectrum access probability for the secondary system.	77
3.15	Packet error rate for the secondary system.	78
3.16	Packet error rates for primary system during the different times of the day, $P_p = -2\text{dBm}$	79

3.17	Packet error rates for primary system during the different times of the day, $P_p = -8\text{dBm}$	80
3.18	Spectrum access probability for the secondary system during the different times of the day.	81
4.1	Two tone input signal. F-1 and F1 represents the two dominant fundamental frequency components of two-tone input signal whereas df represent the frequency spacing between them.	86
4.2	Response of a nonlinear HPA to two-tone input signal.	87
4.3	OFDM	88
4.4	Theoretical (solid lines) and simulation (marker points) results for symbol error rate for NPM, 16 QAM, $N = 512$ for different values of IBO (a) = 15dB (b) = 10dB (c) = 7.5dB (d) = 5dB (e) = 2dB in an AWGN channel.	94
4.5	Theoretical and simulation results for average SER for each subcarrier in MPM1 due to nonlinear noise only, 16 QAM, $N = 256$, $N_g = 64$, IBO = 3dB.	95
4.6	Theoretical and simulation results for average SER for each subcarrier in MPM2 due to nonlinear noise only, 16 QAM, $N = 256$, $N_g = 64$, IBO = 3dB.	96
4.7	Theoretical and simulation results for average SER for each subcarrier in MPM3 due to nonlinear noise only, 16 QAM, $N = 512$, $N_g = 128$, IBO = 8dB.	97

4.8	Theoretical (solid lines) and simulation (marker points) results for average SER in AWGN channel for MPM3, 16 QAM, $N = 512$, $N_g = 128$ with different values of IBO (a) = 15dB (b) = 10dB (c) = 7.5dB (d) = 5dB (e) = 2dB.	98
4.9	Theoretical (solid lines) and simulation (marker points) results for average SER in AWGN channel for a HPA with memory, 16 QAM, $N = 512$, $N_g = 128$ with different values of IBO, (a) = 6dB (b) = 10dB (c) = 13dB (d) = 15dB.	100
4.10	Theoretical (solid lines) and simulation (marker points) results for average SER in frequency selective Rayleigh fading channel for a HPA with memory, 16 QAM, $N = 512$, $N_g = 128$ with different values of IBO, (a) = 6dB (b) = 10dB (c) = 13dB (d) = 15dB. . . .	101
B.1	Block diagram of the hardware setup to extract the coefficient for a nonlinear HPA	114
B.2	Implementation error:- Theoretical vs Practical(Test-bed measurements),16QAM	115

List of Tables

1.1	Summary of types of diversity	10
1.2	Comparison of interference control, interference mitigation and in- terference avoidance spectrum sharing techniques [12]	24
1.3	Comparison of existing experimental cognitive radio testbeds	36
3.1	Physical layer parameters	63
4.1	Coefficients for MPM	93
4.2	Exponential power delay profile	99

List of Abbreviations

ACK	acknowledgement
ADD	analog-to-digital converter
AF	amplify-and-forward
AF-SC	amplify-and-forward with superposition coding
AH-DSA	ad hoc dynamic spectrum access
AM	amplitude modulation
ARQ	automatic repeat and request
AWGN	additive white Gaussian noise
BPSK	binary phase shift keying
BER	bit error rate
BWRC	Berkeley Wireless Research Center
CBS	central base station
CCTH	cooperative clear-to-help
CCTS	cooperative clear-to-send
CPS	conventional primary system
CR	cognitive radio

CRC	cyclic redundancy check
CRTS	cooperative right to send
CSI	channel state information
CSS	cooperative spectrum sharing
DF	decode-and-forward
DSA	dynamic spectrum access
DFT	discrete fourier transform
FCC	Federal Communication Commission
FDM	frequency division multiplexing
FIR	finite impulse response
FFT	fast fourier transform
FPGA	field programmable gate array
G-K	Gauss-Kronrod
GPP	general-purpose processor
GSM	global system for mobile communications
HPA	high power amplifier
HSS	hierarchical spectrum sharing
IBO	input back-off
IDA	Infocomm Development Authority of Singapore
IFFT	inverse fast fourier transform
ISI	inter-symbol interference

LOS	line-of-sight
LSB	lower side band
LTE	long term evaluation
LTI	linear time invariant
MAC	media access control
MCM	multi-carrier modulation
MEMS	micro-electro mechanical systems
MPM	memory polynomial model
MRC	maximal ratio combining
NACK	negative acknowledgement
NIC	network interface cards
NLOS	non-line-of-sight
NPM	nonlinear polynomial model
OFDM	orthogonal frequency division multiplexing
OSSS	orthogonal spectrum sharing scheme
PAPR	peak to average power ratio
PER	packet error rate
PHY	physical
PM	phase modulation
PR	primary receiver
PT	primary transmitter

QAM	quadrature amplitude modulation
QoS	quality of service
QPSK	quadrature phase shift keying
SAP	spectrum access probability
SDR	software defined radio
SDF	selective decode-and-forward
SER	symbol error rate
SFF	small foam factor
SNR	signal-to-noise ratio
SR	secondary receiver
SSPA	solid state power amplifiers
STBC	space time block code
ST	secondary transmitter
TDD	time division duplex
TWTA	traveling wave tube amplifier
UMTS	universal mobile telecommunication system
USB	upper side band
USRP	universal software radio peripheral
W-H	Wiener-Hammerstein
WiFi	wireless fidelity
WiMAX	worldwide interoperability for microwave access

WSN	wireless sensor networks
WSS	wide sense stationary uncorrelated scattering

List of Symbols

$(\cdot)^*$	conjugate
$(\cdot)^T$	transpose
$(\cdot)^H$	hermitian
$(\cdot)^R$	real part
$(\cdot)^I$	imaginary part
$E(\cdot)$	expectation
$z \sim \mathcal{CN}(\mu, \sigma^2)$	complex Gaussian random variable with mean μ and variance σ^2
N	number of subcarriers
N_g	guard interval length
n	discrete time index
k	subcarrier index
m	OFDM symbol index
B_q	number of delay samples
qth	index of the delay tap
τ_l	normalized discrete delay

L	number of multipaths
$h_l^{(m)} \sim \mathcal{CN}(0, \sigma_l^2)$	channel response of l th multipath
$(k)_N$	residue of k modulo N
A	input amplitude for the maximum amplifier output power
$x \sim \varepsilon(\lambda)$	exponential random variable x with mean $\frac{1}{\lambda}$
ν	path loss component
d_i	distance between the respective transmitters and receivers
P_p	transmit power at PT
P_s	transmit power at ST
$\overline{SNR}_d$	average received SNR between PT and PR without spectrum sharing
$\overline{SNR}_T$	target average SNR for primary system
$\overline{SNR}_{MRC}$	average received SNR for the primary system after the retransmission with MRC at PR
$\overline{SNR}_p^{AF-SC}$	average received SNR for the primary system for AF-SC
$\overline{SNR}_s^{AF-SC}$	average received SNR for the secondary system for AF-SC
$\overline{SNR}_p$	average received SNR of the primary system with OSSS
$\overline{SNR}_s$	average received SNR of secondary system with OSSS
α	power allocation factor

PE_p	packet error for primary system
PE_s	packet error for secondary system
SA	spectrum access for secondary system
τ_l	normalized discrete delay of multi-path channel
P_{AF}^{out}	outage probability for AF relaying at high SNR region
P_{DF}^{out}	outage probability for DF relaying at high SNR region
P_{SDF}^{out}	outage probability for SDF relaying at high SNR region
P_p^{out}	outage probability for primary system with CSS
P_s^{out}	outage probability for secondary system with CSS

Introduction

1.1 Motivations and Objectives

“Our nation’s wireless needs are too often governed by 1970s regulations that hinder economic progress and innovation. We need to re-think our approach to radio spectrum to bring our national policy into the wireless era and ensure that spectrum is available for entrepreneurs, innovators and first responders.”

Edward Zander (Former CEO and Chairman of the Board of Motorola)

With an exponential increase in the number of wireless applications in recent years, there is an insatiable demand for more radio spectrum. Perpetuating the problem further, most of the radio spectrum (3kHz to 300GHz) has already been allocated under the licensed band and is no longer available for new wireless systems. This is evident from a glance at Singapore’s Infocomm Development Authority (IDA) spectrum allocation chart shown in Fig. 1.1.

Although licensing the radio spectrum for exclusive usage guarantees protection against harmful interference from other radio systems, recent studies have suggested that such an approach result in inefficient and under-utilization of most of the allocated spectrum resources. As observed from Fig. 1.2 the average spectrum occupancy for the whole frequency band of study is just 4.54%. Moreover, most of the public safety and military radio systems require spectrum for occasional operation which leads to an additional amount of unused spectrum.

As a result of this inherent inefficiency of current spectrum allocation policies as well as scarcity of radio spectrum, researchers over the years have proposed

Figure 1.1. Spectrum allocation chart in Singapore [1]

Figure 1.2. Band by band average spectrum occupancy in Singapore [2]

alternative spectrum access techniques to improve spectral efficiency and capacity in radio communication, giving birth to the notion of cognitive radios [3]. Conceptually, cognitive radios are able to co-exist with the current licensed users without degrading their performance. Cognitive radios achieve this through the utilization of advanced wireless and signal processing techniques [4] to exploit opportunities in the spectrum where they are able to transmit their data without interfering with the licensed users.

Design and implementation of prototype testbeds for cognitive radio (CR) networks in general are inherently complex, costly and a time consuming affair [5], hence computer simulation has been a preferred methodology for researchers over the years. As a result, most research on CR networks have been limited to the theoretical performance evaluation and simulations. But there is growing concern that most of the current simulators make several simplifying assumptions that may or may not be valid in practice and there might be a significant gap between theoretical and measured results. To aggravate the problem further, the few testbeds that are available for CR networks are mostly concentrating on the “detect-and-avoid” interweave schemes for cognitive radio [6]-[11].

Recently, cooperative spectrum sharing (CSS) [12]-[20], which incorporates cooperative relaying to cognitive radio, has been proposed as an alternative model to the detect-and-avoid model. In CSS, two wireless systems operate over the same frequency band albeit with different priorities. The primary system, comprising of a primary transmitter (PT) and primary receiver (PR), supports the relaying functionality. The secondary system, comprising of a secondary transmitter (ST) and secondary receiver (SR), operate on a secondary basis with the guarantee that its operation does not affect the primary system performance.

It has already been shown through theoretical and simulation results that as long as certain conditions are met, CSS is able to provide spectrum access for the secondary system without degrading the performance of the primary system. However, currently there are no hardware testbeds to demonstrate the practical performance of such protocols. As a consequence, most of the analytical work on CSS cannot be validated.

Another issue related to CSS protocols is that most of them are interference limited and performance of the systems are limited by the amount of interference

acceptable from one system to another. Consequently, there is an inherent trade-off between the achievable performance of the primary and secondary systems [12], [16]-[22].

The above points motivated us to dig deeper into the paradigm of cognitive radio and propose a new CSS protocol and a proof-of-concept implementation of the above protocol on a RF testbed. Our objectives can be summarized as follows

- Investigate the various CSS protocols proposed for cognitive radio networks. Outline their respective pros and cons and discuss new methodologies to improvise them.
- Propose a new CSS protocol and compare it with the conventional CSS protocols through simulation results.
- Provide a benchmark for fellow researchers who intend to advance the state of art in experimental research on cognitive radios by designing and developing a RF testbed based on the proposed CSS protocol.
- Highlight the obtained measurement results under different conditions / scenario, and discuss the potential impact of such measurement results.
- State the key lessons learned while designing and implementing the proposed protocol on the testbed.

1.2 Major contribution of the thesis

Proposed an interference-free CSS protocol. The proposed protocol allows a secondary system to coexist with the primary system in the same spectrum band without interfering one another [23].

- V. A. Bohara, S. H. Ting, Y. Han and A. Pandharipande, “Interference free overlay cognitive radio network based on cooperative space time coding,” in *Proceedings of 5th International Conference on cognitive radio oriented wireless networks and communications, CrownCom 2010*, Cannes, France, June 2010.

We also extended the above work to wireless sensor networks (WSN). Here we showed that when the PT-PR link is weak, WSN can be used to enhance the Quality of Service (QoS) of the primary system in exchange for spectrum access for WSN [24].

- V. A. Bohara, S. H. Ting, Y. Han and A. Pandharipande “An orthogonal spectrum sharing scheme for wireless sensor networks.” *EURASIP Journal on Wireless Communications and Networking* 2011 2011:10.

Designed and implemented a CSS protocol on a NI PXIe testbed. The testbed is programmed to follow the OFDM standards in IEEE 802.11a. Measurement results in a realistic office environment were also obtained, thus proving that CSS is a practically viable approach for secondary spectrum access [25]-[29].

- V. A. Bohara and S. H. Ting, “Preliminary measurement results for cognitive spectrum sharing based on cooperative relaying,” in *Proceedings of International Conference on Wireless Communications & Signal Processing, WCSP 2010*, Suzhou, China, Oct. 2010.
- V. A. Bohara and S. H. Ting, “Design and implementation of overlay cognitive radio network on NI PXIe platform,” in *N.I. ASEAN Virtual Instrumentation Applications Contest 2010*, Sept. 2010. (Awarded the best paper in academic segment) Available: <http://digital.ni.com/worldwide/singapore.nsf/web/all/9E11D52A0EE58F2C862577C0002A28F2>
- V. A. Bohara, S. H. Ting and Y. Han, “Experimental results for cooperative spectrum sharing,” accepted to *Proceedings of IEEE Globecom*, Houston, Texas, USA, Dec. 2011.
- V. A. Bohara, S. H. Ting, “Measurement results for cognitive spectrum sharing based on cooperative relaying,” *IEEE Transactions on Wireless Communication*, vol. 10, no. 7, pp. 2052-2057, July 2011.

An analytical methodology to evaluate the impact of nonlinear high power amplifier (HPA) on an OFDM system is also proposed. The analytical results are useful in deducing the back-off required in a nonlinear HPA while obtaining

the measurement results for the above testbed. It has also been shown that an improper selection of back-off has considerable impact on the end-to-end symbol error rate (SER) performance of an OFDM based communication system [30]-[32].

- V. A. Bohara and S. H. Ting, “Theoretical analysis of OFDM signals in nonlinear polynomial models,” in *Proceedings of 6th ICICS*, Singapore, Dec. 2007.
- V. A. Bohara and S. H. Ting, “Analysis of OFDM signals in nonlinear high power amplifier with memory,” in *Proceedings of International Conference on Communications, ICC 2008*, Beijing, Peoples Republic of China, May 2008.
- V. A. Bohara and S. H. Ting, “Analytical performance of orthogonal frequency division multiplexing systems impaired by a non-linear high-power amplifier with memory,” *IET Communications*, vol. 3, no. 10, pp. 1659-1666, Oct. 2009.

1.3 Organization of the thesis

The remainder of the thesis is organized as follows. The remaining part of chapter 1 provides the necessary background knowledge about the field of wireless communication. A form of transmit diversity known as cooperative diversity that utilizes space-time cooperation between different users in a wireless network is also discussed. This is followed by a discussion on the cognitive radio and dynamic spectrum access techniques to improve the spectrum utilization of current wireless systems. It also gives insights on cooperative and cognitive techniques to improve throughput, energy and spectral efficiency of a wireless network. Furthermore, it discusses the CSS protocols proposed in existing literature, highlighting their pros and cons. Chapter 1 will also touch on the various prototype testbeds that have been proposed in literature for performance evaluation of cognitive radio networks.

In Chapter 2, an interference free CSS protocol for wireless sensor networks (WSNs) is proposed. The drawbacks of traditional interference-limited CSS approaches are highlighted and compared with the proposed scheme. Analytical and simulation results are shown to validate the proposed scheme.

Chapter 3 showcases a testbed for proof-of-concept demonstration and performance assessment of a CSS protocol. The performance of the testbed has been demonstrated by obtaining both quantitative as well as qualitative measurement results.

Chapter 4 analyzes the impact of nonlinear HPA on design and implementation of an OFDM communication system. A theoretical framework has been developed to show that the in-band distortion due to nonlinear HPA can be canonically characterized by a complex attenuation component and nonlinear noise component. A comparison between theoretical and simulated results is also presented to verify the accuracy of the analysis. Chapter 4 also demonstrates the impact of improper selection of input back-off (IBO) on the performance of an OFDM communication system. Finally, chapter 5 provides a summary of the results obtained, draws conclusions, and outlines possible directions for future work.

1.4 Background and Preliminaries

1.4.1 Wireless Communication

In simple terms, wireless communication is defined as the transfer of information over a distance without the use of electrical conductors or “wires”. From 1895, when Guglielmo Marconi opened the way for modern wireless communications, till today with the advent of pre-4G technologies such as mobile WiMAX and 3G Long Term Evolution (LTE), the paradigm of wireless communication has developed at an amazing pace.

There are two fundamental aspects of wireless communication that distinguish it from wired communication and makes it even more challenging. First is the phenomenon of fading [35], which is the time variation of the instantaneous signal strengths due to multipath fading and path loss due to distance attenuation and shadowing. Second is multi-user interference [109], since wireless systems generally operate as a multiple access system in which a number of transmitter/receivers pairs operate in the same physical channel using some form of orthogonal or non-orthogonal multiplexing. Interference can also be caused by two heterogenous wireless system wanting to operate in the same frequency band, for e.g. WiFi and

Figure 1.3. Variation of signal strength vs distance

Bluetooth in the 2.4GHz band.

1.4.1.1 Fading

Fig. 1.3 illustrates the variation of signal strength due to fading. Note that the signal strength varies more rapidly across small distances (small scale fading) but the average signal strength varies slowly with distance (large scale fading) [37]. Large-scale fading is caused by the path loss of signal as a transmitter-receiver pair moves away from each other and shadowing by obstacles. Antenna losses and filter losses also contribute to large scale fading.

Small scale fading is mainly caused by multipaths. When the receiver receives multiple copies of the same transmitted signal through different paths and at different times, the different copies may add up constructively or destructively, resulting in a large variation of signal strength. In practice, the dynamic range for small scale fading can be as large as 30dB. The three main effects of small scale fading are

- Significant changes in the signal strength over a small distance caused by random addition/subtraction of RF multipath waves (constructive/destructive interference).

Figure 1.4. Different classes of small scale fading

- Frequency dispersion (Doppler spread) due to varying Doppler shifts on different multipath signals.
- Time dispersion (temporal distortion or delay spread) caused by the multipath propagation delays

Fig. 1.4 illustrates the different classes of small scale fading.

Communication over a fading channel has relatively poorer performance due to fading. Fig. 1.5 illustrates the degradation of bit error rate (BER) for BPSK modulation in Rayleigh fading channel. When compared to the AWGN case, around 25dB degradation due to the multipath channel (at a BER of 10^{-4}) can be observed. However, the effects of fading can be substantially mitigated through the use of diversity techniques. The basic principle of diversity is to exploit the random nature of radio propagation by finding independent (or at least uncorrelated) signal paths for communication. By having more than one path to select from, the probability that all the paths are in deep fade simultaneously is low, hence both

Table 1.1. Summary of types of diversity

Diversity	Advantages	Disadvantages
Spatial	<p>Easy to design.</p> <p>No extra power or bandwidth required.</p> <p>Can be exploited even when the fading channel is neither frequency selective nor time selective.</p> <p>Number of diversity branches (L) selectable.</p>	<p>Large antenna spacing required.</p> <p>Hardware more expensive.</p>
Temporal	<p>Number of diversity branches (L) selectable.</p> <p>Implicit in coding /interleaving</p> <p>No extra physical space required.</p>	<p>L times more bandwidth necessary.</p> <p>Effective only when the fading is time-selective.</p> <p>Large buffer memory if Doppler frequency is small.</p>
Spectral	<p>Number of diversity branches (L) selectable.</p> <p>No extra physical space required.</p>	<p>L times more power and bandwidth necessary.</p> <p>Effective only when the fading is frequency-selective.</p>

the instantaneous and average SNR at the receiver can be improved [37]. Traditionally three main forms of diversity are exploited in varying degrees in wireless communication systems to compensate for the effects of fading. They are spectral diversity, temporal diversity and spatial diversity. Table 1.1 gives the various pros and cons of these diversity techniques. Spatial diversity outweighs the other two forms of diversity as it does not need any additional bandwidth and power. Moreover, spatial diversity can be used even when the channel is neither frequency selective nor time selective. However, this is at the expense of extra hardware and the corresponding implementation complexities.

Figure 1.5. BER for BPSK modulation in Rayleigh fading channel

1.4.2 Cooperative Communications

Space-time coding realizes spatial diversity by transmitting signals through multiple antennas [38],[39]. However, many wireless devices such as mobile handsets or nodes in a wireless sensor network, are limited by size, computational power or hardware complexity to implement multiple transmit antennas. In such a case, a promising approach to achieve spatial diversity is to enable single antenna users/mobiles in a multi-user environment to share their antennas and form a virtual multiple-antenna node for space-time cooperation. Such a method is formerly known as cooperative diversity [40]-[42] and the network formed due to cooperation between the various nodes is known as a cooperative communication network. In a cooperative communication network, each wireless user is assumed to transmit data as well as act as a cooperative agent (helper) for another user [43] as shown in Fig. 1.6. In the simplest form, a cooperative communication network can be realized as a relay network, in which the relay simply forwards the source information to the destination, as shown in Fig. 1.7. The overall transmission is divided into two-phases. In Phase I, the source broadcasts its signal to destination which is “overheard” by the relay. In Phase II, the relay forwards the source signal

Figure 1.6. Cooperative communication network

Figure 1.7. A two-phase relay network

to destination. After the two phases, the destination combines the signal from the relay and source. This two-phase transmission introduces diversity in time and space making the source-destination link more robust against transmission errors and thereby obtaining higher throughput and reliability.

Depending on how the relay forwards the source information to the destination, different cooperative strategies have been proposed, prominent among them are amplify-and-forward (AF) relaying and decode-and-forward (DF) relaying [43].

1.4.2.1 Amplify-and-forward relaying

In AF relaying (also referred as non-regenerative approach), the relay overhears the signal transmitted by the source to destination in Phase I. The relay then amplifies and retransmits this signal in Phase II as shown in Fig. 1.8. At the destination, the two received signals are combined and source data is regenerated. Due to AF relaying, the destination will have two copies of source signal through two independent paths, thus spatial diversity can be exploited and probability of error can be reduced. The most significant aspect of AF relaying is its simplicity. Since relay just amplifies and retransmits the source data its very easy to implement. Moreover, AF relaying performs best with bad source to relay channels [44]. However, its performance suffers due to error propagation. Since there is no error correction facility in relay, the error from source to relay channel propagates to the destination.

1.4.2.2 Decode-and-forward relaying

In DF relaying (also referred as regenerative approach), the relay attempts to decode the source signal it received in Phase I. If decoding is successful, it regenerates the source signal and transmits it in Phase II as shown in Fig. 1.8. The destination then decodes the combined signal from the source and relay. Since in DF case relay decodes and forwards the source data, there is no error propagation to destination from the source-relay channel. However, this increases the complexity of relay. Depending upon the decoding result of relay, DF relaying can be further divided into fixed DF and selective DF relaying.

1.4.2.2.1 Fixed DF relaying In fixed DF relaying relay is required to fully decode the source signal in Phase I [40], hence source-relay channel becomes the limiting factor for this scheme. As shown later in the thesis, at high SNR, fixed DF is unable to provide the diversity gain for large SNR as performance is limited by successful decoding of source signal at the relay.

1.4.2.2.2 Selective DF relaying To overcome the above limitation of fixed DF relaying, selective DF relaying has been proposed in [40]. In this case, if the relay is unable to decode the source signal in Phase I, the source retransmits

Figure 1.8. AF and DF relaying

to destination in Phase II (through direct transmission). Selective DF relaying enables the cooperating terminals to exploit full spatial diversity and overcome the shortcomings of fixed DF relaying.

1.4.2.3 Outage performance

In this section we obtain the outage performance of the communication network with and without the relay functionality. The channel over the links Source $\rightarrow$ Destination, Source $\rightarrow$ Relay and Relay $\rightarrow$ Destination is modeled as Rayleigh flat fading channel with channel coefficients h_1 , h_2 and h_3 respectively as shown in Fig. 1.8. h_i are assumed to independent and identically distributed with $h_i \sim \mathcal{CN}(0, d_i^{-\nu})$ where ν is the path loss component and d_i is the distance between the respective transmitters and receivers. We also denote $|h_i|^2 = \gamma_i$. Let s denote the signal transmitted by source with zero mean and $E\{s^*s\} = 1$. The transmit power at source and relay is denoted by P_t and P_r respectively.

1.4.2.3.1 Outage behavior of direct transmission To obtain a baseline performance for comparison, we derive the outage performance under direct trans-

mission when the relay does not exist. In this case, s is transmitted through the direct link from source to destination. Denoting the signal received by destination y_{dt} we have,

$$y_{dt} = \sqrt{P_t}h_1s + n_{dt} \quad (1.1)$$

where $n_{dt} \sim \mathcal{CN}(0, \sigma^2)$ is the AWGN at destination for direct transmission. The achievable rate between source and destination is given by $R_d = \log_2 \left(1 + \frac{P_t \gamma_1}{\sigma^2} \right)$. The outage event for target rate R is given by $R_d < R$. Thus the outage probability for the direct transmission in absence of relay can be given as

$$\begin{aligned} P_{out}^d &= \Pr\{R_d < R\} \\ &= 1 - \exp \left(- \frac{\sigma^2 \rho}{d_1^{-\nu} P_t} \right) \\ &= d_1^\nu \frac{\sigma^2 \rho}{P_t} \end{aligned} \quad (1.2)$$

where $\rho = 2^R - 1$.

1.4.2.3.2 Outage behavior of AF relaying Let y_d and y_r be the signal received by destination and relay in Phase I for AF relaying. Thus we have,

$$y_d = \sqrt{P_t}h_1s + n_d \quad (1.3)$$

$$y_r = \sqrt{P_t}h_2s + n_r, \quad (1.4)$$

where $n_d \sim \mathcal{CN}(0, \sigma^2)$ and $n_r \sim \mathcal{CN}(0, \sigma^2)$ are the AWGN at destination and relay respectively in Phase I. After reception in Phase I, relay normalizes the received signal based on its power constraint and further amplifies it to generate a signal

$$s_r = \varphi y_r \quad (1.5)$$

where $\varphi = \sqrt{\frac{P_r}{P_t \gamma_2 + \sigma^2}}$ is the power normalization factor. In Phase II, relay transmits this signal to the destination. The signal received at destination in Phase II can be written as

$$y_{rd} = h_3 s_r + n_{rd} = \sqrt{P_t} h_3 \varphi h_2 s + h_3 \varphi n_r + n_{rd} \quad (1.6)$$

where $n_{rd} \sim \mathcal{CN}(0, \sigma^2)$ is the AWGN at destination in Phase II. Destination then does maximal-ratio combining (MRC) of the signal y_d and y_{rd} for the decoding of s . The achievable rate between the source and the destination with AF relaying can be written as

$$R_{af} = \frac{1}{2} \log_2 \left(1 + \frac{P_t \gamma_1}{\sigma^2} + \frac{P_t \gamma_2 \gamma_3 \varphi^2}{\varphi^2 \gamma_3 \sigma^2 + \sigma^2} \right) \quad (1.7)$$

where $\frac{1}{2}$ accounts for the fact that the transmission of s is distributed over two-phases.

The outage event for a target rate R is given by $R_{af} < R$. If $P_t = P_r = P$, then the outage probability for AF relaying at high SNR region ($\frac{P}{\sigma^2} \gg 1$) can be approximated as [40]

$$\begin{aligned} P_{AF}^{out} &= \Pr(R_{af} < R) \\ &\approx \left(\frac{1}{2d_1^{-\nu}} \frac{d_2^{-\nu} + d_3^{-\nu}}{d_2^{-\nu} d_3^{-\nu}} \right) \left(\frac{\sigma^2 \rho_1}{P} \right)^2 \end{aligned} \quad (1.8)$$

where $\rho_1 = 2^{2R} - 1$. The $\left(\frac{\sigma^2}{P} \right)^2$ behavior in Eq. 1.8 proves that AF relaying offers diversity of 2 at high SNR region.

1.4.2.3.3 Outage behavior of DF relaying Let v_r and v_d be the signal received by relay and destination in Phase I for DF relaying. Thus we have,

$$v_r = \sqrt{P_t} h_2 s + z_r, \quad (1.9)$$

$$v_d = \sqrt{P_t} h_1 s + z_d \quad (1.10)$$

where $z_r \sim \mathcal{CN}(0, \sigma^2)$ and $z_d \sim \mathcal{CN}(0, \sigma^2)$ are the AWGN at relay and destination respectively in Phase I. After reception in Phase I, the relay attempts to decode s . If decoding is successful it regenerates the source signal to give

$$t_r = \sqrt{P_r} s. \quad (1.11)$$

The achievable rate between source and relay after transmission in Phase I is given by $R_1 = \frac{1}{2} \log_2 \left(1 + \frac{P_t \gamma_2}{\sigma^2} \right)$, where $\frac{1}{2}$ accounts for the fact that the overall transmission is divided into two-phases. In Phase II, t_r is broadcasted by relay and received by destination. The signal received at destination is given by

$$v_{rd} = h_3 t_r + z_{rd} = \sqrt{P_r} h_3 s + z_{rd} \quad (1.12)$$

where $z_{rd} \sim \mathcal{CN}(0, \sigma^2)$ is the AWGN at destination in Phase II. Signals v_d and v_{rd} are then combined using MRC for the decoding of s .

The achievable rate at destination after Phase II, conditioned on successful decoding at relay, can be written as

$$R_2 = \frac{1}{2} \log_2 \left(1 + \frac{P_t \gamma_1}{\sigma^2} + \frac{P_r \gamma_3}{\sigma^2} \right). \quad (1.13)$$

If $P_t = P_r = P$, then the outage probability for fixed DF relaying for target rate R at high SNR region ($\frac{P}{\sigma^2} \gg 1$) can be approximated as [40]

$$\begin{aligned} P_{DF}^{out} &= \Pr(R_1 > R) \Pr(R_2 < R) + \Pr(R_1 < R) \\ &\approx \frac{1}{2d_2^{-\nu}} \frac{\sigma^2 \rho_1}{P}. \end{aligned} \quad (1.14)$$

As observed from Eq. (1.14), fixed DF does not offer any diversity gain for large SNR values as the source $\rightarrow$ relay channel becomes the limiting factor in this cases, thus reducing the performance of fixed DF to that of direct transmission between the source and relay.

To overcome the limitation of decode-and-forward relaying at high SNR region, selective DF can be used. In this case, if the relay is unable to decode the source signal in Phase I, the source retransmits to destination in Phase II (through direct transmission). Hence for SDF, Eq. (1.14) can be rewritten as

$$P_{SDF}^{out} = \Pr(R_1 > R) \Pr(R_2 < R) + \Pr(R_1 < R) \Pr(R_{d2} < R). \quad (1.15)$$

where $R_{d2} = \frac{1}{2} \log_2 \left(1 + \frac{2P_t \gamma_1}{\sigma^2} \right)$. The large SNR behavior of SDF relaying can be

approximated as [40]

$$P_{SDF}^{out} \approx \left(\frac{1}{2d_1^{-\nu}} \frac{d_2^{-\nu} + d_3^{-\nu}}{d_2^{-\nu} d_3^{-\nu}} \right) \left(\frac{\sigma^2 \rho_1}{P} \right)^2. \quad (1.16)$$

which also achieves diversity of 2.

1.4.3 Cognitive Radio Communications

Like cooperation, cognition is rapidly emerging as one of the underlying paradigms that will revolutionize the future generation high performance, high efficiency wireless networks [45]. In Section 1.1 we gave some background on cognitive radio and how the current inefficient utilization of spectrum resources necessitates the use of cognitive radio. In this section we will touch this topic in detail. Before going into the depth, let us once again revisit the problem of spectrum scarcity.

1.4.3.1 Spectrum scarcity

Interference from other users is another dominant source of impairment in wireless communication systems. For example, commercial radio devices may experience unacceptable interference from other radio services that are operating near the same spectrum band. Hence, spectrum regulation is needed to protect a wireless system against interference from other wireless systems. Traditionally, spectrum regulation is achieved by licensing the radio spectrum band to a particular wireless system for their exclusive usage.

However, radio spectrum is a scarce and valuable resource. Spectrum licensing has resulted into allocation of most of the prime spectrum to traditional wireless communication systems and there are few spectrum bands left for new wireless services or operators. The gravity of the problem was witnessed in the recent auction for 3G services where mobile operators had to shell out billions of dollars for spectrum licenses. The problem is further aggravated by the unutilization or under-utilization of allocated spectrum bands by traditional wireless communication services.

The above factors have propelled the search for alternative spectrum access techniques giving rise to the paradigm of cognitive radios. In simple terms, a

Figure 1.9. Key elements of cognitive radio

cognitive radio is an intelligent transceiver that is able to sense its ambient communication environment and consequently adapt its radio parameters to provide the best possible quality of service to the unlicensed users with minimal interference to the licensed users. Fig. 1.9 highlights the key parameters of a cognitive radio.

- **Sense:** In order to avoid interference to the primary user, cognitive radio should be able to sense portions of spectrum with no or reduced primary user activity and use them for its own communication. However, it should immediately vacate the band as soon as primary user activity is detected. Sensing also enables the cognitive device to dynamically change its transmission parameters based on its current knowledge of RF environment. Sensing forms an integral part of detect-and-avoid cognitive systems.

Figure 1.10. Cognitive radio network

- **Flexible and agile:** The cognitive radio should be able to dynamically program itself according to the radio environment. The operating frequency and other radio parameters should be agile and can be reconfigured on the fly.
- **Learn and adapt:** Cognitive radio should be able to learn and adapt according to the user environment. It should be able to analyze the sensory data, recognize patterns and modify internal operational behavior based on the analysis of past and present inputs.

The ultimate objective of a cognitive radio is to make efficient utilization of the currently limited radio spectrum through the capabilities described above. In achieving the above, it should make sure not to compromise the performance of the primary user.

1.4.3.2 Hierarchical spectrum sharing

One model to realize a cognitive radio framework is through hierarchical spectrum sharing (HSS). Under this model, a secondary system (unlicensed user), comprising of secondary transmitters (ST) and secondary receivers (SR), is allowed to co-exist in the same frequency band as a primary system (licensed user), which comprises of primary transmitters (PT) and primary receivers (PR) as shown in Fig. 1.10. A higher priority is given to the primary system and the secondary system operates on

Figure 1.11. Hierarchical spectrum sharing [50]

a secondary basis with lower priority. The priority of the primary system is quantified by the constraint that secondary system will access the spectrum intelligently without adversely affecting the primary system. As illustrated in Fig. 1.11, hierarchical spectrum sharing can be broadly classified under three categories, namely interference control, interference mitigation and interference avoidance [49], [50].

1.4.3.2.1 Interference avoidance Also known as spectrum interweave, in this scenario, opportunistic spectrum access by the secondary system is only allowed when the radio spectrum allocated to the primary system is determined to be unused [3], [47]-[62]. This is the fundamental approach for most of the detect-and-avoid cognitive radio schemes. This approach maintains the orthogonality between the primary and secondary signals in time and / or frequency and hence prevents interference between primary and secondary systems. Fig. 1.12 illustrates the interference avoidance behavior of a HSS protocol.

Figure 1.12. Interference avoidance

1.4.3.2.2 Interference control The secondary systems is allowed to transmit simultaneously in the same frequency band as the primary system with a constraint that the interference seen at the primary system due to secondary system is below a designated threshold level [46], [59]-[61]. Hence the potential interference at the primary system is controlled by strictly limiting the transmission power of the secondary system. Fig. 1.13 illustrates the interference control behavior of a HSS protocol.

1.4.3.2.3 Interference mitigation The secondary system is allowed to simultaneously transmit in the same frequency band as the primary system but it is assumed that the secondary transmitter has non-causal information (provided by a genie) of the primary system message or code books [22],[49],[50],[51]-[54]. This non-causal information of the primary system helps the secondary system to mitigate the interference at the primary receiver due to secondary transmission through techniques such as Gel'fand-Pinsker (GP) binning [55] or dirty-paper coding (DPC)[56]. Fig. 1.14 illustrates the interference mitigation behavior of a HSS protocol.

Figure 1.13. Interference control

Figure 1.14. Interference mitigation

Table 2.2 summarizes the differences between the interference avoidance, interference control and interference mitigation approaches for HSS protocol.

Table 1.2. Comparison of interference control, interference mitigation and interference avoidance spectrum sharing techniques [12]

Interference control	Interference mitigation	Interference avoidance
Channel side information: ST knows the channel strengths to PR.	Codebook side information: Secondary system knows the channel gains, codebooks and the messages of primary system	Activity side information: Secondary system knows the spectral holes in space, time or frequency when the primary system is not using these holes.
Secondary system can transmit simultaneously with primary system as long interference caused is below acceptable limit.	Secondary system can transmit simultaneously with primary system; the interference at the primary receiver can be offset/mitigated by using precoding techniques such as GP binning and DPC.	Secondary system transmits only when the signal from the primary system is determined to be absent except in the cases of false spectral hole detection.

1.4.4 Cooperative and Cognitive Wireless Systems

As discussed in previous sections, cooperative communication helps to combat multi-path fading and shadowing effects in wireless channels and thereby increasing the reliability and throughput of wireless communication networks. On the other hand, cognitive radios promote the efficient utilization of scarce radio spectrum thereby increasing the capacity and spectral efficiency of wireless communication networks. It is quite obvious that cognitive and cooperative principles are complementary to each other and thus it appears reasonable to exploit this natural synergy by applying them jointly. Power, energy, spectral efficiency and diversity are key resources that can be traded in different ways to achieve a desired level of performance. Heterogenous systems combining cooperative and cognitive technologies form a promising resource-trading framework for wireless communication

networks [45].

In the context of cognitive radios, cooperative transmission can give rise to the following two different scenarios [15].

1.4.4.1 Cooperative transmission between secondary systems

In this scenario, a secondary node acts as a relay for the secondary system. This technique aims to increase reliability and overall throughput for the secondary system. All the considerations that are valid for cooperative communication are also applicable here. The only variation is that the secondary system needs to continuously monitor the channel to detect any possible transmission from the primary system.

One of the most important research areas of this scenario is the class of cooperative spectrum sensing. In order to make spectrum sensing robust to severe fading environments and increase the probability of detection of the primary system's signal several authors have proposed cooperation among secondary nodes [57]- [69]. Cooperative sensing exploits the benefits of spatial diversity among the sensing nodes to improve the probability of detection. In addition, cooperative sensing can also solve hidden primary user problem and it can decrease sensing time [66],[69]. Fig. 1.15 shows a typical cooperative spectrum sensing network. In this case, the fusion center makes sensing decisions by combining observations / decisions of local sensors. The results presented in [62] show that cooperative sensing can deliver tremendous gains even with a small to moderate number of secondary nodes, as long as the nodes are far enough apart from each other. Moreover, by exploiting cooperation among secondary users it is possible to achieve a target system-level probability of detection in the case when each CR faces an "SNR Wall" [58] below which it is unable to reliably detect a primary system's signal. It has also been shown in [69] that cooperative spectrum sensing is most effective when the cooperative cognitive radios observe statically independent fading or shadowing environment.

However, cooperation increases the amount of control information needed in the network. Moreover, delays in cooperation puts additional challenges in design of a cooperative sensing network [70]. Hence there is a compromise between the

Figure 1.15. Cooperative spectrum sensing network

amount of delay and signalling overhead that is acceptable in a network and the performance achieved by cooperative sensing.

1.4.4.2 Cooperative transmission between primary and secondary system

In this scenario, the primary system is assumed to possess the relay functionality and the secondary system acts as a relay for the primary system. In fact, the interference mitigation technique discussed earlier can be considered as a part of such a configuration. One of the most important application of cooperative transmission between primary and secondary systems is the class of cooperative spectrum sharing (CSS). Suppose the primary system wants to maintain a predefined target rate, R_t , between PT and PR for effective communication. Consider a scenario in which the actual transmission rate between PT and PR drops below R_t . PT will seek cooperation from neighboring terminals to enhance its transmission performance by broadcasting a cooperative right-to-send (CRTS) message which also indicates R_t for the primary system. PR responds to CRTS by transmitting a cooperative clear-to-send (CCTS) message. Upon overhearing CRTS and CCTS, ST estimates the channel gains of PT-ST and PR-ST links, and decides whether the R_t requirement for the primary system can be met if it serves as a relay for

Figure 1.16. Cooperative transmission between primary and secondary system

the primary system. If yes, ST responds by sending a cooperative clear to help (CCTH) message to PT and PR, and the primary system correspondingly switches to a two-phase relaying transmission mode, with ST as the relay terminal. However, if ST is not able to assist the primary system to achieve R_t , it will simply remain silent and the primary system hence retains its original direct transmission from PT to PR. There are two fundamental justifications for such a configuration.

- By helping primary users complete their transmissions faster will in turn lead to more transmission opportunities for secondary systems [14].
- It has been proven in [16]-[20] that by receiving assistance from the secondary system to relay its information, overall QoS (measured in terms of throughput and reliability) of the primary system can be improved.

1.4.4.3 Performance analysis of a CSS protocol

The system configuration for a CSS protocol¹ is shown in Fig. 1.16. The channel between all the links i.e. PT-PR, PT-ST, PT-SR, ST-PR and ST-SR are modeled as Rayleigh flat fading with channel coefficients h_1, h_2, h_3, h_4 and h_5 respectively, thus $h_i \sim \mathcal{CN}(0, d_i^{-\nu})$, $i = 1, 2, 3, 4, 5$ where ν is the path loss component and d_i is the normalized distance between the respective transmitters and receivers. This normalization is done with respect to the distance between PT and PR, thus $d_1 = 1$. The instantaneous channel gain of each link is denoted by $\gamma_i = |h_i|^2$. The primary and secondary signals are denoted by x_p and x_s respectively, have zero mean and $E[x_p^* x_p] = 1$, $E[x_s^* x_s] = 1$. We denote the transmit power at PT and ST as P_p and P_s respectively.

1.4.4.3.1 Outage performance of primary system Let x_p be the primary signal transmitted by PT in Phase I. Denoting the signals received by PR, ST, and SR in Phase I as y_{11} , y_{21} , and y_{31} respectively, we have

$$y_{i1} = \sqrt{P_p} h_i x_p + n_{i1} \quad (1.17)$$

where $i = 1, 2, 3$ and $n_{i1} \sim \mathcal{CN}(0, \sigma^2)$ is the additive white Gaussian noise (AWGN) at the respective receivers for Phase I. The achievable rate between PT and ST is thus given as $R_{p1} = \frac{1}{2} \log_2 \left(1 + \frac{P_p \gamma_2}{\sigma^2} \right)$, where $\frac{1}{2}$ accounts for the fact that the overall transmission is divided into two-phases. After reception in Phase I, ST attempts to decode x_p . If the decoding is successful, ST regenerates x_p . A composite signal z_s is created by superimposing the regenerated primary signal x_p with power αP_s and the secondary signal x_s with power $(1 - \alpha) P_s$. Thus

$$z_s = \sqrt{\alpha P_s} x_p + \sqrt{(1 - \alpha) P_s} x_s. \quad (1.18)$$

In Phase II, z_s is broadcasted by ST and received by PR and SR. The signal received at PR is given by

$$y_{12} = (\sqrt{\alpha P_s} h_4) x_p + (\sqrt{(1 - \alpha) P_s} h_4) x_s + n_{12} \quad (1.19)$$

¹Please note that performance analysis shown here is for DF based CSS protocol, for AF based CSS protocol please refer to [18]

where $n_{12} \sim \mathcal{CN}(0, \sigma^2)$ is the AWGN at PR in Phase II. Signals y_{11} and y_{12} are then combined at PR using MRC for the decoding of x_p .

Thus achievable rate between PT and PR, conditioned on successful decoding at ST, can be written as

$$R_{p2} = \frac{1}{2} \log_2 \left(1 + \frac{P_p \gamma_1}{\sigma^2} + \frac{P_s \alpha \gamma_4}{P_s (1 - \alpha) \gamma_4 + \sigma^2} \right). \quad (1.20)$$

On the other hand, if ST fails to decode x_p in Phase I, it will simply remain silent in Phase II. In this case, PR can still decode for x_p through the direct link from PT to PR² and the achievable rate between PT and PR is given by $R_D = \log_2 \left(1 + \frac{P_p \gamma_1}{\sigma^2} \right)$. The outage probability of primary signal transmission with target rate R_t is thus given by

$$P_P^{out} = \Pr(R_{p1} > R_t) \Pr(R_{p2} < R_t) + \Pr(R_{p1} < R_t) \Pr\left(\frac{1}{2} R_D < R_t\right). \quad (1.21)$$

After some manipulation [16], Eq. (1.21) can be approximated as

$$P_P^{out} \approx \begin{cases} P_{p1}^{out} & 0 \leq \alpha < \bar{\alpha} \\ P_{p2}^{out} & \bar{\alpha} \leq \alpha \leq 1 \end{cases} \quad (1.22)$$

where $P_{p1}^{out} = 1 - \exp\left(-\frac{\sigma^2}{P_p} \left((d_2^\nu + 1)\vartheta - \frac{\alpha}{1-\alpha}\right)\right) - \exp\left(-\frac{\vartheta \sigma^2}{P_p}\right) + \exp\left(-\frac{\sigma^2}{P_p} \vartheta \left((d_2^\nu + 1)\right)\right)$ and $P_{p2}^{out} = 1 - \exp\left(-d_2^\nu \frac{\sigma^2}{P_p} \vartheta\right) - \exp\left(-\frac{\sigma^2 \vartheta}{P_p}\right) + \exp\left(-\frac{\sigma^2}{P_p} \vartheta (d_2^\nu + 1)\right)$ where $\bar{\alpha} = \frac{\vartheta}{1+\vartheta}$ and $\vartheta = 2^{2R_t} - 1$.

1.4.4.3.2 Outage performance of secondary system After the reception of y_{31} in Phase I, SR attempts to decode x_p , and stores the decoding result if it succeeds. The achievable rate between PT and SR is given by $R_{1s} = \frac{1}{2} \log_2 \left(1 + \right.$

²Please note that this direct link refers to signal transmitted from PT to PR in Phase I. There will be no retransmission from PT to PR in Phase II.

$\frac{P_p \gamma_3}{\sigma^2}$). In Phase II, the signal received at SR is

$$y_{22} = (\sqrt{\alpha P_s} h_5) x_p + (\sqrt{(1-\alpha) P_s} h_5) x_s + n_{22} \quad (1.23)$$

where $n_{22} \sim \mathcal{CN}(0, \sigma^2)$ is the AWGN at SR in Phase II. Assuming successful decoding at SR in Phase I, the interference component $\sqrt{\alpha P_s} h_5 x_p$ can be easily cancelled out from (1.23) to obtain $y'_{22} = (\sqrt{(1-\alpha) P_s} h_5) x_s + n_{22}$. The achievable rate between ST and SR, conditioned on successful decoding of x_p at both ST and SR in Phase I, is given as $R_{2s} = \frac{1}{2} \log_2 \left(1 + \frac{(1-\alpha) P_s \gamma_5}{\sigma^2} \right)$.

Thus the outage probability of the secondary transmission with target rate R_s is given by [16]

$$\begin{aligned} P_s^{out} &= 1 - \Pr(R_{1p} > R_t) \Pr(R_{1s} > R_t) \Pr(R_{2s} > R_s) \\ &= 1 - \exp \left(- \left(\frac{\sigma^2 (d_2^\nu + d_3^\nu) \vartheta}{P_p} + \frac{\sigma^2 (d_5^\nu) \vartheta_1}{P_s (1-\alpha)} \right) \right) \end{aligned} \quad (1.24)$$

where $\vartheta_1 = 2^{2R_s} - 1$.

The following points are worth noting about the CSS protocol discussed above.

- It has been proven in [16] that as long as ST is located within a critical radius from PT, there exists a value of α above which P_p^{out} will be less than or equal to the case without spectrum sharing. Hence, spectrum access for secondary system is possible without compromising on the performance of the primary system.
- P_p^{out} is independent of P_s and d_5 for $P_s \gg \sigma^2$, while P_s^{out} decreases with increasing P_s and decreasing d_5 . Thus for a given value of α , it is possible to improve the outage performance of secondary system, without compromising the performance of primary system, just by increasing P_s or decreasing d_5 .

Although CSS schemes are promising approaches for secondary spectrum access, there are some challenges associated with it [12].

1. Most of the CSS approaches are based on superposition coding [71], in which ST divides its power between the relay transmission of primary signal and

the secondary signal. Therefore, there is an inherent trade-off between the achievable performance of the primary and secondary systems [12], [16]-[20], [21]. In other words, the performance of secondary system is limited by the amount of interference acceptable at PR from ST.

2. Some CSS approaches [21],[53] assume that ST has non-causal knowledge of the primary user's codebooks and its messages, thus making this technique practically unrealizable.

1.4.5 Testbeds for Cognitive Radios

Majority of the research studies that evaluate cognitive radio performance are based on theoretical frameworks or computer simulations. Although analytical approaches verified with computer simulations can give useful insights into the performance of a specific protocol, they make several oversimplifying assumptions that may or may not be valid in realistic environments. Moreover, most of the computer simulations are limited by the inherent inability to accurately model the interference and random nature of wireless medium. Factors such as the above introduce significant gap between simulation and experimental results and may cause a considerable behavioral difference between simulated and the real system. Therefore, in order to augment the development of cognitive radio and fully understand some of the specific design issues, proposed schemes for cognitive radio should be verified and demonstrated in real world environment through experimental platforms and testbeds.

Such a platform/testbed will serve many purposes. The following lists some of them

- It will help to bridge the gap between simulation and reality.
- A successful demonstration of benefits of cognitive radio in realistic scenarios will alleviate the fears of spectrum regulators and might help them to move forward with the regulatory framework needed to open up the spectrum for shared usage.
- A practical demonstration will also address the concerns of primary users regarding the interference from the unlicensed / secondary users.

- It will help to understand how a cognitive system behaves in realistic scenarios leading to a deeper understanding of cognitive radios and better exploitation of a cognitive system.
- It will highlight the strength and weakness of proposed protocols / algorithms that is very difficult to contrive or gauge in simulations.

In the paradigm of wireless communication, testbeds are not a novel concept. Several testbeds have been used predominantly in the past to demonstrate the viability of a proposed technique or algorithm and have contributed significantly in the advancement of wireless communication research. However, cognitive radio research is a highly interdisciplinary research drawing expertise from signal processing, networking, machine learning and other engineering and computer science disciplines and hence requires new testbed capabilities. Moreover, it also involves complex interaction between the various layers of the protocol stack that need to be addressed jointly. The following summarizes some of the requirements of a cognitive radio testbed³

- Software reconfigurability: The testbed should be software reconfigurable so that different physical / media access control (MAC) layer functionalities can be controlled through software. For example, most of the physical layer functions like modulation, demodulation, detection, coding, interleaving , deinterleaving etc can be implemented in software, and can be reconfigured according to user requirement.
- Cross layer support: The testbed should be able to integrate the functionalities of different layers of protocol stack through real-time protocol implementation on the hardware.
- Adaptability: The testbed should be able to adapt itself in different communication environment, so that realistic measurements can be taken in controlled but different environments.

³Please note that requirements listed in [5] and [74] are in terms of spectrum sensing capabilities of the testbed, in this section we highlight some of the general requirements of a cognitive radio testbed. Some of the important points have been taken from [5] and [74].

- High speed analog to digital converters (ADCs) and digital to analog converters (DACs): Bandwidth and dynamic range of the ADC (ADC determines the sampling rate of the incoming signal) and DAC (DAC determines the analog bandwidth of the transmitted signal) and the processing capability of the CPU limits the complexity of the resource allocation algorithms that can be implemented in real time on the testbeds. Hence the testbed should have sufficiently high resolution ADC's and DAC's in order to support computationally intensive and complex algorithms.
- Multiple radio support: The testbed should be able to support multiple radios, which can be used as primary or secondary users.

In the following we describe some of the key features of existing testbeds for cognitive radios.

1.4.5.1 CORNET

Cognitive radio network testbed (CORNET) [11] is a highly reconfigurable cognitive radio testbed developed in Virginia Tech Institute that consists of 48 wireless software defined radio nodes spread over the four floors of the campus building. Each of the 48 nodes consists of a general-purpose processing (GPP) platform and a flexible RF front end with an field programmable gate array (FPGA) onboard. The GPP consists of an Intel Xeon processor whereas the RF front end consists of Ettus Research Universal Software Radio Peripheral 2 (USRP2) [77]. All the wireless nodes are also equipped with custom made daughterboard capable of spanning the frequency range 100MHz to 4GHz using variable instantaneous bandwidth from 10KHz to 20MHz. The nodes are also endowed with 14-bit, 100 MSamples /s ADC's and 16bit, 400 MSamples /s DAC's. Gigabit Ethernet interface is used as communication link between the GPP and the USRP2. The two key aspects of this testbed is the flexibility offered at PHY as well as at MAC layer and support for GNU [75] and OSSIE (SCA based SDR platform)[76] based software frameworks. This testbed is currently being used to identify the network operations that might affect the scalability of ad hoc dynamic spectrum access (AH-DSA) networks.

1.4.5.2 ORBIT

Open access research testbed for Next-Generation Wireless Networks (ORBIT) [78], [79] is an open access radio grid wireless network testbed developed at WIN-LAB, Rutgers University. It consists of 400 ORBIT Radio Nodes laid out in a 20×20 grid with each node separated from the adjacent nodes by 1m. Each ORBIT Radio Node consists of a small form factor (SFF) PC with a 1 GHz VIA C3 processor, 512 MB of RAM, 20 GB of local disk and several 802.11 a/b/g network cards. The ORBIT testbed allows both controlled experimentation on emulator as well as outdoor field measurements in realistic settings. Experimenters can install their own network layer protocols or application software to construct a specific networking or application scenario for study or develop their own applications, protocol stacks, MAC layer modifications on the testbed. The ORBIT Measurement Framework & Library (OML) is responsible for logging the experimental results for offline analysis. However, the experiments on the ORBIT testbeds is primarily focussed on higher network layers with very limited research on the PHY layer.

1.4.5.3 Emulab

EMULAB [80] is the web-accessible, reconfigurable network testbed with the primary installation run by the Flux Group at the University of Utah and with more than two dozen sites all around the world. It consist of more than 300 nodes (till date) with varying hardware configuration. Most of the test nodes are powered with 2.4 GHz 64-bit Quad Core Xeon E5530 “Nehalem” processor and 12 GB 1066 MHz DDR2 RAM (6 x 2GB modules). Some of the nodes in Emulab are also equipped with USRP. The software platform used for USRP devices is GNU Radio. Emulab requires the user to start an experiment by creating a network simulator (NS) file. By creating an NS file, user can describe the network topologies and configure the USRP board. For the duration of an experiment, a user has exclusive and essentially complete control over the devices that are allocated to him or her. At the end of an experiment, the user releases the devices back to the testbed, and the testbed must reclaim them so that they can be usefully and safely allocated to another user in the future. One of the ongoing cognitive radio experiments on this testbed is the utilization of distributed sensing technique for

estimation of background noise floor and spectrum sensing to detect the presence of any active users in the spectrum.

1.4.5.4 BWRC cognitive radio testbed

The Berkeley Wireless Research Center (BWRC) cognitive radio testbed [81], [82] is based on BEE2 (Berkeley Emulation Engine) which is a generic, multi-purpose, multi-FPGA emulation engine. It also consists of highly reconfigurable radio modems that are designed to operate in an unlicensed 2.4GHz industrial, scientific and medical (ISM) band. The radios and BEE2 are connected with the help of optical fibre cable that can extend up to a distance of one-third of a mile. The BEE2 consists of 5 Xilinx Vertex-2 Pro 70 FPGAs in a single compute module. Four FPGAs are used for computation and one for control. All computation FPGAs are connected to the control FPGA via 20Gbps links. Each FPGA embeds a PowerPC 405 core, which minimizes the latency and maximizes the data throughput between the microprocessor and reconfigurable logic. Each FPGA is also connected to external 4GB DDR memory which can be used for logging experimental data over long period of time. The analog/baseband processing is implemented with 14-bit 128 MHz DACs, 12-bit 64 MHz ADCs, and 32 MHz wide baseband filters. The software design flow is built using Matlab/ Simulink from Mathworks and Xilinx system generator, which are used to map high level design and state machine representation to FPGA configurations. BEE2-enhanced Xilinx Platform studio (BAPS) directly converts the Simulink design to FPGA configuration.

This testbed is currently being used to evaluate the performance and limitations of spectrum sensing algorithms proposed for detection of primary user signals [74]. For this case, implementation and experimental results have been presented for wideband energy detectors and cyclostationary feature detectors.

Table 1.3 shows the comparison between various testbeds used for CR deployment.

1.4.5.5 Issues with existing deployments of CR testbeds

Although the above experimental testbeds have contributed greatly to the better understanding of the performance of cognitive radios in realistic scenarios, further

Table 1.3. Comparison of existing experimental cognitive radio testbeds

Testbed Name	Location	Hardware	Software	Past & Current CR experiments
BWRC cognitive radio testbed	Univ. of Berkeley	FPGA's	Matlab / Simulink and Xilinx System generator	Spectrum sensing algorithms implementation and experiments.
EMULAB	University of Utah	USRP	GNU	Distributed spectrum sensing techniques
CORNET	Virginia Tech University	GPP and USRP2	GNU and OSSIE	Scalability of AF-DSA networks
ORBIT	WINLAB, Rutgers University	USRP and USRP2	GNU	MAC layer adaptation in cognitive radio networks

work needs to be done to demonstrate the viability of cognitive radios to solve the current spectrum scarcity problem. Moreover, there are some major issues related to the current deployments of cognitive radios.

- Most of the existing experiments on cognitive radios are taking a unidirectional approach and concentrating on spectrum sensing methods for the detection of spectrum holes (detect-and-avoid interweave approach) as shown in Table 1.3. The need of the hour is to broaden this approach, and think of alternative ways to design and implement a cognitive system. Although spectrum sensing techniques are efficient methods to realize a cognitive radio system, there is always a concern of interfering with the primary system. None of the protocols or scheme that have been proposed for spectrum sensing can provide an accuracy of 100% (probability of detection of primary signal is 1). The problem gets worse when the primary signal is very weak and its SNR is even below the noise floor. Moreover, the performance of a cognitive system designed for spectrum sensing is limited by the speed of ADC (which determines how large a bandwidth can be sampled instantaneously).
- Another issue related with some of the current deployments is inaccessible PHY layer. Some of the CR deployments use commercially available 802.11 network cards to model the PHY layer, hence time sensitive functions cannot

be implemented properly.

1.5 Conclusions

In this chapter, we presented the motivation behind this thesis. We also highlighted some of the major contribution of this thesis. Furthermore, we provided some background knowledge about wireless communication and showed how fading and interference degrades the performance of a wireless communication system. We also touched upon some of the predominant cooperative techniques like AF and DF relaying and their significance in alleviating the problem of fading. We also obtained the outage probability performance of AF and DF relaying in Rayleigh flat fading channel.

We further showed how spectrum licensing and under-utilization of the allocated spectrum has resulted into spectrum scarcity and how cognitive radio can be an effective technology to solve this spectrum scarcity problem. HSS was discussed as a possible framework for cognitive radio wherein two wireless system can operate in same frequency band albeit with different priorities. The pros and cons of interference avoidance, interference control and interference mitigation approaches that are commonly used to realize a HSS protocol were also discussed. The usage of CSS as an alternative to above approaches was investigated. The performance of a conventional CSS protocol was quantified by obtaining the outage probability of primary and secondary systems in Rayleigh flat fading channel.

Apart from above, we listed some of the requirements of cognitive radio testbeds and gave a brief overview of the experimental testbeds that have been proposed. We also highlighted some of the major issues with the current testbeds.

An Orthogonal Spectrum Sharing Scheme for Wireless Sensor Networks

2.1 Introduction

In Chapter 1, we gave the theoretical background related to CSS protocols. We showed how CSS protocols can be utilized by secondary system to gain spectrum access from the primary system. However, most of proposed CSS protocols are interference limited hence the performance of the systems are limited by the amount of interference acceptable from one system to another. Consequently, there is a inherent trade-off between the achievable performance of each system. In this chapter, we try to resolve the above issue by proposing an interference-free CSS protocol, henceforth called as orthogonal spectrum sharing scheme (OSSS), which alleviates the interference from the primary system to secondary system and vice versa. In the following, the performance of OSSS has been demonstrated by taking wireless sensor networks (WSNs)[83]-[85] as an example.

Recently WSNs are being increasingly deployed all over the world at an accelerated pace. This has been made practically feasible by significant advances in microelectro-mechanical systems (MEMS) technology, radio communications and digital electronics [84]. A typical WSN consists of spatially distributed sensor

nodes deployed in an ad-hoc manner which collects data and pass on to a central base station (CBS) via a radio link. The CBS can be a PC, data server, dedicated monitoring device, or any other gateway to a higher data rate device. WSN are used for various applications including military surveillance, habitat monitoring, object tracking, traffic monitoring etc.

Most of the sensor nodes are autonomous and send data over the radio link only when required. Furthermore, there is an increasing trend of deploying WSN in urban areas as part of the infrastructure to support smart building initiatives and power meter readings for smart grids, to name a few. However, radio spectrum in urban areas are generally extremely crowded and thus it is not possible nor economically viable to allocate a dedicated radio spectrum band to a WSN.

Factors such as the above have spurred the demand for alternative spectrum access techniques for WSNs [47], [87], [90]. This demand has been further compounded by the inefficient usage of the licensed bands by the incumbent (primary) systems. Researchers over the years have proposed dynamic spectrum access (DSA) techniques to utilize the spectrum more efficiently by allowing a secondary system (for example a WSN) to co-exist in the same frequency band as a primary system and opportunistically access the licensed bands. However, most of these techniques are interference limited, and the performance of the systems are limited by the amount of interference acceptable by one system from another.

In this chapter, by taking the above factors into consideration, we propose an OSSS which allows a WSN to gain spectrum access along with a primary system without causing any interference to one another. As a result, the performance of primary system is not limited by the interference from WSN and vice versa. In the proposed scheme, a WSN, henceforth known as secondary system, is assumed to be a single-hop network with every sensor node being able to directly communicate with every other node. Secondary transmitters (STs) are spatially distributed sensor nodes that cooperatively monitor their physical environmental conditions and send an update to their CBS, which for simplicity will be denoted as SR. ST can communicate with each other in real time and the communication link between them can be formed by using a radio, laser, infrared or an optical media depending upon the availability [84]. This inter-node communication helps in status monitoring of the STs and also avoids duplication of data at SR. Moreover,

it also keeps all STs well informed of the latest information being sent to SR.

Under the proposed framework, the secondary system operates in the same frequency band as an incumbent primary system, which comprises of PT and PR. A higher priority is given to the primary system and the secondary system operates on a lower priority with a constraint that its operation does not affect the performance of primary system. For ease of analysis, we limit ourselves to two ST nodes, ST(1) and ST(2) and denote them as a ST cluster or simply **ST** wherever necessary. Do note that due to inter-node communication, ST(1) and ST(2) has access to the same sensor information that is to be sent to SR.

Cooperation techniques to enhance the performance of a communication system in terms of diversity, coverage extension, etc, have been discussed extensively in Chapter 1. Control signalling for practical cooperation schemes have also been proposed in [91]-[94]. In our proposed scheme, we presume that the primary system is an advanced system with a relaying functionality, like IEEE 802.16j [95], and it employs a practical handshake mechanism for cooperative relaying [94].

Consider a scenario in which the average signal to noise ratio (SNR) between PT and PR drops below a particular threshold. PT will seek cooperation from neighboring terminals to enhance its transmission performance by broadcasting a CRTS message which also indicates the target average SNR, $\overline{\text{SNR}}_T$, for the primary system. PR responds to CRTS by transmitting a CCTS message. Upon overhearing CRTS and CCTS, **ST** decides¹ whether $\overline{\text{SNR}}_T$ can be met if it serves as an amplify-and-forward (AF) relay² for the primary system. If yes, ST(2) responds by sending a CCTH message to PT and PR, and the primary system correspondingly switches to a two-phase AF relaying transmission mode, with ST(1) acting as the primary relay. However, if **ST** is not able to assist the primary system to achieve $\overline{\text{SNR}}_T$, it will simply remain silent.

Once **ST** is confirmed as a relay, secondary spectrum access is achieved by adopting the following two-phase transmission protocol. The system models for the 1st and 2nd phase are shown in Fig. 2.1 and Fig. 2.2 respectively. In the 1st phase, the primary signal transmitted by PT to PR is overheard by ST(1) and SR.

¹It should be noted that whether **ST** is able to assist PT or not, is a probabilistic event due to the random fading channels.

²If ST uses decode-and-forward instead of amplify-and-forward the protocol reduces to the one proposed in [23]

Simultaneously in the same phase, ST(2) transmits the secondary signal which is received by SR as well as PR. At ST(1), the primary signal received in the 1st phase is amplified according to its power constraint.

In the 2nd phase, ST(1) and ST(2) transmit the amplified primary signal and secondary signal respectively. At PR, the received signals after the two-phase transmission are multiplied by an orthogonalization vector to cancel out the interference due to secondary signal and retrieve the primary signal. The secondary signal is retrieved at SR in the same way.

The main advantages of OSSS is summarized as follows.

- The most important attribute of OSSS is that it is not interference limited because of the orthogonality between the received primary and secondary signals. As a result, the performance of primary and secondary system is not limited by the interference from ST and PT respectively.
- The secondary system is able to achieve spectrum access as long as it is willing to increase its transmit power such that $\overline{\text{SNR}}_{\text{T}}$ is met. This ability to trade-off transmit power with spectrum access opportunity is an attractive feature for WSNs as it allows the sensor nodes to maintain its Quality of Service (QoS), such as delay constraints.
- OSSS also ensures that the performance of the WSNs can be maintained regardless of the availability of spectrum holes, thus making this scheme extremely attractive in dense urban areas where spectrum holes are hard to obtain.
- The existence of the secondary system is transparent to the primary system as it is the onus of ST to “disguise” itself as a primary relay in exchange for the chance to access the spectrum.

As a basic requirement for OSSS, we assume that the necessary CSI needed at SR can be obtained through standard pilot symbol-aided channel estimation methods [97]-[99]. We analyze OSSS by deriving the closed-form expressions for average SNR of the primary system. For comparison, we also consider an interference limited scheme where **ST** uses AF with superposition coding (AF-SC) [18]. We

Figure 2.1. OSSS: 1st transmission phase

show that for the same $\overline{\text{SNR}}_T$ requested by the primary system, OSSS can achieve a much higher performance for the secondary system than AF-SC.

2.1.1 System Model and Protocol Description

2.1.1.1 System model

The system model under consideration for the 1st and 2nd transmission phase is shown in Fig. 2.1 and Fig. 2.2 respectively. The channel between all the links i.e. PT-PR, PT-ST(1), ST(1)-PR, ST(2)-PR, ST(1)-SR, ST(2)-SR and PT-SR

Figure 2.2. OSSS: 2nd transmission phase

are modeled as Rayleigh flat fading with channel coefficients $h_1, h_2, h_3, h_4, h_5, h_6$ and h_7 respectively, thus $h_i \sim \mathcal{CN}(0, d_i^{-\nu})$, $i = 1, 2, 3, 4, 5, 6, 7$ where ν is the path loss component and d_i is the distance between the respective transmitters and receivers. Thus all the links between the terminals can be characterized by the set of parameters $\{h_i, d_i\}$ as shown in Fig. 2.1 and Fig. 2.2. The instantaneous channel gain of each link is denoted by $\gamma_i = |h_i|^2$. The primary and secondary signals are denoted by x_p and x_s respectively, have zero mean and $\mathbb{E}[x_p^* x_p] = 1$, $\mathbb{E}[x_s^* x_s] = 1$. We denote the transmit power at PT and ST as P_p and P_s respectively.

2.1.1.2 Protocol description

The protocol flowchart for the transmission from PT is shown in Fig. 2.3. In the situation where only the primary system is operating, i.e there is no spectrum sharing, the average received SNR between PT and PR is given by

$$\overline{\text{SNR}}_d = \text{E} \left[\frac{P_p \gamma_1}{\sigma^2} \right] = \frac{P_p}{d_1^\nu \sigma^2} \quad (2.1)$$

where σ^2 is the variance of additive white Gaussian noise (AWGN) at PR. The following steps illustrate the control signalling involved.

1. PT obtains $\overline{\text{SNR}}_d$ from PR through conventional channel quality feedback mechanism [100] and checks whether $\overline{\text{SNR}}_d < \overline{\text{SNR}}_T$. If yes, go to step 2. Otherwise continue with the ongoing transmission.
2. PT checks whether a retransmission of the same signal as part of an ARQ protocol will assist in achieving $\overline{\text{SNR}}_T$, i.e.

$$\overline{\text{SNR}}_{\text{MRC}} \geq \overline{\text{SNR}}_T \quad (2.2)$$

where $\overline{\text{SNR}}_{\text{MRC}} = \frac{2P_p}{d_1^\nu \sigma^2}$ is the average received SNR for the primary system after the retransmission with maximum ratio combining (MRC) at PR. If yes, PT proceeds with ARQ protocol. Otherwise, go to step 3.

3. PT transmits CRTS which indicates $\overline{\text{SNR}}_T$ required by the primary system and PR responds by sending CCTS.
4. Upon overhearing CRTS and CCTS from PT and PR respectively, **ST** will decide whether it is able to assist the primary system in achieving $\overline{\text{SNR}}_T$ by calculating $\overline{\text{SNR}}_p$, which is the achievable average received SNR of the primary system with OSSS. If $\overline{\text{SNR}}_p \geq \overline{\text{SNR}}_T$, then ST(2) will broadcast CCTH, and the primary system correspondingly switches to the two-phase OSSS protocol. Otherwise, **ST** will simply remain silent.

Figure 2.3. Protocol flowchart

2.2 Average Received SNR for OSSS

2.2.1 Average received SNR of primary system with OSSS

2.2.1.1 Phase 1

In the 1st transmission phase, as shown in Fig. 2.1, the primary signal x_p is transmitted by PT and secondary signal x_s is transmitted by ST(2) simultaneously. Denoting the signals received by PR, SR and ST(1) as $y_{pr}^{(1)}$, $y_{sr}^{(1)}$ and y_{st} respectively, we have ³,

$$y_{pr}^{(1)} = \sqrt{P_p}h_1x_p + \sqrt{P_s}h_4x_s + n_{11}, \quad (2.3)$$

$$y_{sr}^{(1)} = \sqrt{P_p}h_7x_p + \sqrt{P_s}h_6x_s + n_{12}, \quad (2.4)$$

$$y_{st} = \sqrt{P_p}h_2x_p + n_{13}. \quad (2.5)$$

Here $n_{1j} \sim \mathcal{CN}(0, \sigma^2)$, $j = 1, 2, 3$ is the AWGN at the respective receivers in the 1st transmission phase.

2.2.1.2 Phase 2

Let $z_s^{(1)}$ and $z_s^{(2)}$ be the transmitted signals from ST(1) and ST(2) during the 2nd phase respectively. The transmitted signal vector in the 2nd phase from **ST** can then be written as

$$\mathbf{z}_s = \begin{bmatrix} g & 0 \\ 0 & \sqrt{\frac{P_s}{2}} \end{bmatrix} \mathbf{x}_{st} \quad (2.6)$$

where $\mathbf{z}_s = \begin{bmatrix} z_s^{(1)} & z_s^{(2)} \end{bmatrix}^T$, $\mathbf{x}_{st} = \begin{bmatrix} y_{st} & x_s \end{bmatrix}^T$ and $g = \sqrt{\frac{P_s}{2(P_p\gamma_2 + \sigma^2)}}$. The signal received at PR in the 2nd phase is thus,

$$y_{pr}^{(2)} = \mathbf{h}_p \mathbf{z}_s + n_{21} \quad (2.7)$$

³Please note that ST(1) and ST(2) continuously update each other of the information that needs to be send to the SR. Thus, in the 1st phase, even if ST(1) receives the signal x_s from ST(2), it has *a priori* knowledge of x_s so it can be cancelled out easily from the received signal at ST(1).

where $\mathbf{h}_p = \begin{bmatrix} h_3 & h_4 \end{bmatrix}$ and $n_{21} \sim \mathcal{CN}(0, \sigma^2)$ is the AWGN. From (2.6) and (2.7), we obtain,

$$\begin{aligned} y_{pr}^{(2)} &= \sqrt{\frac{P_s}{2}} h_4 x_s + g h_3 y_{st} + n_{21} \\ &= \sqrt{\frac{P_s}{2}} h_4 x_s + g h_3 \sqrt{P_p} h_2 x_p + n_3. \end{aligned} \quad (2.8)$$

where $n_3 = g h_3 n_{13} + n_{21}$. Thus the signal at PR after the two-phase transmission can be written as

$$\mathbf{y}_p = \mathbf{H}_p \mathbf{x} + \mathbf{n}_p \quad (2.9)$$

where $\mathbf{y}_p = \begin{bmatrix} y_{pr}^{(1)} & y_{pr}^{(2)} \end{bmatrix}^T$, $\mathbf{x} = \begin{bmatrix} x_p & x_s \end{bmatrix}^T$, $\mathbf{n}_p = \begin{bmatrix} n_{11} & n_3 \end{bmatrix}^T$ and

$$\mathbf{H}_p = \begin{bmatrix} \sqrt{P_p} h_1 & \sqrt{P_s} h_4 \\ \sqrt{P_p} g h_3 h_2 & \sqrt{P_s/2} h_4 \end{bmatrix}. \quad (2.10)$$

Multiplying the orthogonalization vector $\mathbf{w}_p = \begin{bmatrix} \sqrt{1/2} & -1 \end{bmatrix}$ to $\mathbf{y}_p$ we obtain,

$$\mathbf{w}_p \mathbf{y}_p = \left(\sqrt{\frac{P_p}{2}} h_1 - g \sqrt{P_p} h_2 h_3 \right) x_p + \sqrt{1/2} n_{11} - n_3. \quad (2.11)$$

It is clear that the secondary signal x_s has been completely removed. Thus the signal received at PR experiences no interference from the secondary transmission. The instantaneous received SNR at PR after the two-phase transmission is given by

$$\begin{aligned} \text{SNR}_p &= \frac{\left| \sqrt{P_p/2} h_1 - \sqrt{P_p} g h_2 h_3 \right|^2}{\mathbb{E} \left[\left| \sqrt{1/2} n_{11} - n_3 \right|^2 \right]} \\ &= \frac{P_p \left\{ \gamma_1 + 2g^2 \gamma_2 \gamma_3 - 2\sqrt{2} g \text{Re}(h_2 h_3 h_1) \right\}}{(2g^2 \gamma_3 + 3) \sigma^2}. \end{aligned} \quad (2.12)$$

The average received SNR at PR for the primary transmission can be derived as

$$\begin{aligned} \overline{\text{SNR}}_p &= \mathbb{E}[\text{SNR}_p] \\ &= \frac{d_3^\nu P_p^2 \left\{ 3d_3^\nu P_p - d_2^\nu P_s - d_2^\nu P_s \left[\ln \left(\frac{3d_3^\nu P_p}{d_2^\nu P_s} \right) \right] \right\}}{d_1^\nu (3d_3^\nu P_p - d_2^\nu P_s)^2 \sigma^2} \\ &\quad + \frac{P_p P_s \left\{ 9d_3^{\nu^2} P_p^2 - P_s^2 (d_2^\nu)^2 - 6d_3^\nu P_p P_s d_2^\nu \left[\ln \left(\frac{3d_3^\nu P_p}{d_2^\nu P_s} \right) \right] \right\}}{(3d_3^\nu P_p - P_s d_2^\nu)^3 \sigma^2}. \end{aligned} \quad (2.13)$$

Please refer to Appendix A for the derivation.

2.2.2 Average received SNR of secondary system with OSSS

2.2.2.1 Phase 1

In the 1st transmission phase, the signal received at SR is $y_{sr}^{(1)}$ which is given in (2.4).

2.2.2.2 Phase 2

The signal received at SR in the 2nd phase is

$$y_{sr}^{(2)} = \mathbf{h}_s \mathbf{z}_s + n_{22} \quad (2.14)$$

where $\mathbf{h}_s = \begin{bmatrix} h_5 & h_6 \end{bmatrix}$ and $n_{22} \sim \mathcal{CN}(0, \sigma^2)$ is the AWGN. Substituting (2.6) into (2.14) we obtain

$$y_{sr}^{(2)} = \sqrt{\frac{P_s}{2}} h_6 x_s + g \sqrt{P_p} h_5 h_2 x_p + n_4. \quad (2.15)$$

where $n_4 = g h_5 n_{13} + n_{22}$. Thus the signal at SR after the two-phase transmission can be written as

$$\mathbf{y}_s = \mathbf{H}_s \mathbf{x} + \mathbf{n}_s \quad (2.16)$$

where $\mathbf{y}_s = \begin{bmatrix} y_{sr}^{(1)} & y_{sr}^{(2)} \end{bmatrix}^T$, $\mathbf{n}_s = \begin{bmatrix} n_{12} & n_4 \end{bmatrix}^T$ and

$$\mathbf{H}_s = \begin{bmatrix} \sqrt{P_p}h_7 & \sqrt{P_s}h_6 \\ \sqrt{P_p}gh_5h_2 & \sqrt{P_s/2}h_6 \end{bmatrix}. \quad (2.17)$$

Multiplying $\mathbf{y}_s$ with an orthogonalization vector $\mathbf{w}_s = \begin{bmatrix} gh_5h_2 & -h_7 \end{bmatrix}$, we obtain,

$$\mathbf{w}_s\mathbf{y}_s = \left(g\sqrt{P_s}h_6h_5h_2 - \sqrt{\frac{P_s}{2}}h_6h_7 \right) x_s + gh_5h_2n_{12} - h_7n_4. \quad (2.18)$$

It is clear from (2.18) that the primary signal x_p has been completely removed. Therefore SR does not experience any interference from the primary transmission. The channel estimate h_7 and h_5h_2 required at SR for the orthogonalization vector w_s can be obtained from the pilot-aided channel estimation procedures detailed in Section 2.2.3. The instantaneous received SNR at SR after the two-phase transmission can be obtained as

$$\begin{aligned} \text{SNR}_s &= \frac{\left| g\sqrt{P_s}h_6h_5h_2 - h_6\sqrt{P_s/2}h_7 \right|^2}{\text{E} \left[\left| gh_5h_2n_{12} - h_7n_4 \right|^2 \right]} \\ &= \frac{P_s \left(\gamma_6\gamma_7 + 2g^2\gamma_6\gamma_5\gamma_2 - 2\sqrt{2}g\text{Re}(h_2h_5h_7)\gamma_6 \right)}{(g^2\gamma_5\gamma_2 + g^2\gamma_5\gamma_7 + \gamma_7)2\sigma^2}. \end{aligned} \quad (2.19)$$

The average received SNR at SR, $\overline{\text{SNR}}_s = \text{E}[\text{SNR}_s]$, is intractable and we will analyze it numerically.

2.2.3 Channel estimation and other requirements

For the various transmitting and receiving terminals in OSSS, we assume that channel estimation can be done through the pilot symbols in the control frames (CRTS, CCTS and CCTH) and data frames originating from PT and **ST**. With the help of pilot symbols in the CRTS frame, SR is able to estimate h_7 . Similarly, the product channel for PT-ST(1)-SR (the relay channel from PT to PR), i.e. h_2h_5

can be estimated at SR in the 2nd phase from PT's pilot symbols since ST(1) is an AF relay [99]. Moreover, the flag indicating the switch from conventional decoding to OSSS decoding at PR can be incorporated in CCTH.

2.3 Average Received SNR for AF with Superposition Coding (AF-SC)

In this section we discuss and derive the average SNR for AF-SC protocol. The control signalling involved is exactly the same as OSSS which is given in Section 2.1.1.2.

2.3.1 Average received SNR of primary system with AF-SC

2.3.1.1 Phase 1

The system model for the 1st transmission phase of AF-SC is shown in Fig. 2.4. In this phase, both ST(1) and ST(2) overhears the signal transmitted from PT⁴. The channel coefficient between PT-ST(2) is denoted by h_8 where $h_8 \sim \mathcal{CN}(0, d_8^{-\nu})$ and $\gamma_8 = |h_8|^2$. Denoting the signals received by PR, **ST** and SR as $s_{pr}^{(1)}$, $\mathbf{s}_{st}$ and $s_{sr}^{(1)}$ respectively, we have

$$s_{pr}^{(1)} = \sqrt{P_p} h_1 x_p + \eta_{11}, \quad (2.20)$$

$$\mathbf{s}_{st} = \sqrt{P_p} \begin{bmatrix} h_2 \\ h_8 \end{bmatrix} x_p + \begin{bmatrix} \eta_2 \\ \eta_8 \end{bmatrix}, \quad (2.21)$$

$$s_{sr}^{(1)} = \sqrt{P_p} h_7 x_p + \eta_{14}, \quad (2.22)$$

⁴If there is only one ST node, then AF-SC reduces to the spectrum sharing scheme proposed in [18]

Figure 2.4. AF-SC: 1st transmission phase.

where $\mathbf{s}_{st} = \begin{bmatrix} s_{st}^{(2)} & s_{st}^{(8)} \end{bmatrix}^T$. $s_{st}^{(2)}$ and $s_{st}^{(8)}$ are the signal received by ST(1) and ST(2) respectively, and $\eta_{11}, \eta_2, \eta_8, \eta_{14}$ are the AWGN with variance σ^2 at the respective receivers. **ST** will then select the received signal with a higher received power, i.e. $s_{st}^{(\tau_{opt})} = \sqrt{P_p} h_{\tau_{opt}} x_p + \eta_{\tau_{opt}}$, $\tau_{opt} \in \{2, 8\}$ where

$$\tau_{opt} = \arg \max_{\tau \in \{2, 8\}} \left(|s_{st}^{(\tau)}|^2 \right). \quad (2.23)$$

As a result, selection diversity is achieved at **ST** in the 1st phase. After performing selection, **ST** normalizes the received primary signal based on its power constraint

and further amplifies it with the power allocation factor α where $0 \leq \alpha \leq 1$. The remaining power $(1 - \alpha)$ is assigned to the secondary signal. Thus the signal vector regenerated from $\mathbf{ST}$ can be written as

$$\mathbf{v}_{st} = \mathbf{V} \mathbf{x}_{af} \quad (2.24)$$

where $\mathbf{v}_{st} = \begin{bmatrix} v_{st}^{(1)} & v_{st}^{(2)} \end{bmatrix}^T$ is the transmit vector from $\mathbf{ST}$, and $v_{st}^{(1)}$, $v_{st}^{(2)}$ are the signals from ST(1) and ST(2) respectively,⁵

$$\mathbf{V} = \begin{bmatrix} \kappa\sqrt{\alpha} & 0 \\ 0 & \sqrt{(1-\alpha)P_s} \end{bmatrix},$$

$$\mathbf{x}_{af} = \begin{bmatrix} s_{st}^{(\tau_{opt})} & x_s \end{bmatrix}^T \text{ and the power normalization factor is given by } \kappa = \sqrt{\frac{P_s}{(P_p\gamma_{\tau_{opt}} + \sigma^2)}}.$$

2.3.1.2 Phase 2

The system model for the 2nd transmission phase of AF-SC is the same as OSSS as shown in Fig. 2.2. In this phase, the signal received by PR is given by

$$s_{pr}^{(2)} = \mathbf{h}_{af} \mathbf{v}_{st} + \eta_{21} \quad (2.25)$$

where $\mathbf{h}_{af} = \begin{bmatrix} h_3 & h_4 \end{bmatrix}$ and $\eta_{21} \sim \mathcal{CN}(0, \sigma^2)$ is the AWGN. After substituting (2.24) in (2.25) we obtain,

$$s_{pr}^{(2)} = \left(\sqrt{P_p} \alpha \kappa h_{\tau_{opt}} h_3 \right) x_p + \left(\sqrt{P_s} (1 - \alpha) h_4 \right) x_s + \sqrt{\alpha} \kappa h_3 \eta_{\tau_{opt}} + \eta_{21}. \quad (2.26)$$

Unlike OSSS, $s_{pr}^{(2)}$ also contains interference from the secondary signal. This interference limits the achievable performance of primary system in AF-SC. The signals $s_{pr}^{(1)}$ and $s_{pr}^{(2)}$ are then combined at PR using MRC for decoding of x_p . The SNR

⁵We may consider other choices such as $\mathbf{V} = \begin{bmatrix} \kappa\sqrt{\frac{\alpha}{2}} & \sqrt{\frac{(1-\alpha)P_s}{2}} \\ \kappa\sqrt{\frac{\alpha}{2}} & \sqrt{\frac{(1-\alpha)P_s}{2}} \end{bmatrix}$ or $\mathbf{V} = \begin{bmatrix} \kappa\sqrt{\alpha} & \sqrt{(1-\alpha)P_s} \\ 0 & 0 \end{bmatrix}$. Though not given in the paper, simulation results show that the $\mathbf{V}$ we used in (2.24) achieves the best performance among the three.

after MRC is given by

$$\text{SNR}_{\text{p}}^{\text{AF-SC}} = \frac{P_p \gamma_1}{\sigma^2} + \frac{P_p \gamma_{\tau_{\text{opt}}} \gamma_3 \kappa^2 \alpha}{P_s (1 - \alpha) \gamma_4 + \alpha \kappa^2 \gamma_3 \sigma^2 + \sigma^2}. \quad (2.27)$$

The average received SNR at PR, $\overline{\text{SNR}}_{\text{p}}^{\text{AF-SC}}$ for AF-SC is intractable and we will analyze it numerically. All channel estimates required for PR can be obtained from the pilot symbol-aided channel estimation procedures detailed in Section 2.2.3.

2.3.2 Average received SNR of secondary system with AF-SC

2.3.2.1 Phase 1

The signal received at SR in the 1st transmission phase is given by

$$s_{sr}^{(1)} = \sqrt{P_p} h_7 x_p + \eta_{13} \quad (2.28)$$

where $\eta_{13} \sim \mathcal{CN}(0, \sigma^2)$ is the AWGN. At SR, an estimate of x_p is obtained using (2.28) as

$$\hat{x}_p = \frac{s_{sr}^{(1)}}{\sqrt{P_p} h_7} = x_p + \frac{\eta_{13}}{\sqrt{P_p} h_7}. \quad (2.29)$$

2.3.2.2 Phase 2

The signal received at SR in the 2nd transmission phase is

$$s_{sr}^{(2)} = \mathbf{h}_{af}^s \mathbf{v}_{st} + \eta_{22} \quad (2.30)$$

where $\mathbf{h}_{af}^s = \begin{bmatrix} h_5 & h_6 \end{bmatrix}$ and $\eta_{22} \sim \mathcal{CN}(0, \sigma^2)$ is the AWGN. Substituting (2.24) in (2.30) we obtain

$$s_{sr}^{(2)} = \left(\sqrt{P_p} \alpha \kappa h_{\tau_{\text{opt}}} h_5 \right) x_p + \left(\sqrt{P_s (1 - \alpha)} h_6 \right) x_s + \sqrt{\alpha \kappa} h_5 \eta_{\tau_{\text{opt}}} + \eta_{22}. \quad (2.31)$$

The estimate $\hat{x}_p$ in (2.29) is used to cancel out the interference component $(\sqrt{P_p}\alpha\kappa h_{\tau_{opt}}h_5)x_p$ from $s_{sr}^{(2)}$, to obtain

$$\widehat{s_{sr}^{(2)}} = \left(\sqrt{P_s(1-\alpha)}h_6 \right) x_s - \frac{\sqrt{\alpha}\kappa h_{\tau_{opt}}h_5\eta_{13}}{h_7} + \sqrt{\alpha}\kappa h_5\eta_{\tau_{opt}} + \eta_{22} \quad (2.32)$$

The channel estimates $h_{\tau_{opt}}h_5$ and $h_{\tau_{opt}}$ required at SR can be obtained from the pilot symbol-aided channel estimation procedures detailed in Section 3.2.3 and [18]. Therefore, the SNR at SR can be obtained as

$$\text{SNR}_s^{\text{AF-SC}} = \frac{P_s(1-\alpha)\gamma_6\gamma_7}{\alpha\kappa^2(\gamma_{\tau_{opt}} + \gamma_7)\gamma_5\sigma^2 + \gamma_7\sigma^2}. \quad (2.33)$$

The average received SNR at SR, $\overline{\text{SNR}}_s^{\text{AF-SC}}$ is intractable and we will analyze it numerically.

Figure 2.5. System configuration for simulation.

Figure 2.6. Average received SNR of primary transmission for various values of $\frac{P_s}{\sigma^2}$ for OSSS, AF-SC and direct transmission with ARQ. Theoretical and simulation values are reported for $\overline{\text{SNR}}_p$ and $\overline{\text{SNR}}_{\text{MRC}}$, whereas only simulation values are reported for $\overline{\text{SNR}}_p^{\text{AF-SC}}$.

2.3.3 Simulation results and discussion

For ease of exposition, PT, SR, **ST** and PR are assumed to be collinear and the distance between ST(1) and ST(2) is assumed to be much smaller than the distance between the other system nodes, thus $d_2 \approx d_8$, $d_3 \approx d_4$ and $d_5 \approx d_6$. The position of PT, SR, **ST** and PR are fixed to (0,0), (0.25,0), (0.5,0) and (1,0) respectively as shown in Fig. 2.5. The path loss component is chosen to be $\nu = 4$. Thus all the radio links between PT, PR, **ST** and SR can be characterized by their respective positions on the straight line.

Fig. 2.6 shows the average SNR performance of primary system for OSSS, $\overline{\text{SNR}}_p$ with respect to $\frac{P_s}{\sigma^2}$ for different values of $\frac{P_p}{\sigma^2}$. The corresponding plot for secondary system, $\overline{\text{SNR}}_s$, is shown in Fig. 2.7. For comparison purposes, we have also plotted the results for $\overline{\text{SNR}}_{\text{MRC}}$ which is the average received SNR of primary system for direct transmission with ARQ. $\overline{\text{SNR}}_{\text{MRC}}$ will be a useful benchmark for comparison as $\overline{\text{SNR}}_{\text{MRC}}$ shows the performance of primary system with retransmission in the absence of any secondary system. Good agreement between the simulation and theoretical results for $\overline{\text{SNR}}_p$ and $\overline{\text{SNR}}_{\text{MRC}}$ in Fig. 2.6 validates the analytical results obtained in this paper.

From Fig. 2.6 and Fig. 2.7, it can be observed that the performance of primary as well as secondary system for OSSS improves with an increase in $\frac{P_s}{\sigma^2}$ for a given value of $\frac{P_p}{\sigma^2}$. This proves that the secondary transmission does not interfere with the primary transmission; in fact it contributes to the performance of the primary transmission. Moreover, it also shows that an increase in secondary transmission power P_s benefits both the primary as well as secondary systems. Another observation that can be made from Fig. 2.6 is that when the primary system is interested in improving its QoS (e.g. $\overline{\text{SNR}}_T > 13\text{dB}$ at $\frac{P_p}{\sigma^2} = 10\text{dB}$ or $\overline{\text{SNR}}_T > 23\text{dB}$ at $\frac{P_p}{\sigma^2} = 20\text{dB}$), it can always request the help of **ST** to improve its QoS while at the same time allowing spectrum access by the secondary system. QoS improvement of up to 8dB can be achieved by the primary system in the case of OSSS with respect to $\overline{\text{SNR}}_{\text{MRC}}$ at $\frac{P_s}{\sigma^2} = 40\text{dB}$ for both $\frac{P_p}{\sigma^2} = 10\text{dB}$ and $\frac{P_p}{\sigma^2} = 20\text{dB}$. From Fig. 2.6, we can also conclude that if QoS requirement for the primary system is set too high (e.g. $\overline{\text{SNR}}_T > 22\text{dB}$ at $\frac{P_p}{\sigma^2} = 10\text{dB}$), $\overline{\text{SNR}}_p < \overline{\text{SNR}}_T$ and secondary spectrum access is not possible. This limitation is due to the noise amplification at ST(1) in the AF relaying. Thus when $\overline{\text{SNR}}_T$ requirement is reasonable, secondary system

is always able to achieve spectrum access as long as it is willing to increase its transmit power such that $\overline{\text{SNR}}_T$ is met.

Fig. 2.6 and Fig. 2.7 respectively show $\overline{\text{SNR}}_p^{\text{AF-SC}}$ and $\overline{\text{SNR}}_s^{\text{AF-SC}}$ for AF-SC at $\alpha = 0.5$ and $\alpha = 0.9$. From the two figures it can be easily deduced that there is a trade-off between the performance of primary and secondary systems, and the performance of one system is limited by the interference from the other system. As we increase the value of α , the performance of primary system improves whereas the performance of secondary system deteriorates and vice versa. In AF-SC, the performance of primary system is limited by the interference from the secondary system as well as amplified noise in the 2nd phase. From Fig. 2.6, at $\frac{P_p}{\sigma^2} = 20\text{dB}$ $\overline{\text{SNR}}_p^{\text{AF-SC}} < \overline{\text{SNR}}_{\text{MRC}}$ for all values of $\frac{P_s}{\sigma^2}$ even with $\alpha = 0.9$. Thus there is no possibility of spectrum access for the secondary system in this case. Furthermore, for $\alpha = 0.9$ at $\frac{P_p}{\sigma^2} = 10\text{dB}$, AF-SC achieves the closest possible performance to OSSS for the primary system, but OSSS outperforms AF-SC by a large margin for the secondary transmission as can be observed from Fig. 2.7.

2.4 Conclusion

In this chapter, we proposed a two-phase orthogonal spectrum sharing scheme (OSSS) based on cooperative amplify-and-forward relaying for a WSN (a.k.a secondary system) to achieve spectrum access along with a primary system. We showed that by using the proposed scheme, the two systems can co-exist in the same frequency band without causing any interference to one another. Moreover, when the PT-PR link is weak, WSN can be used to enhance the QoS of the primary system. We further showed that in OSSS, WSN is always able to achieve spectrum access as long as it is willing to increase its transmit power such that SNR_T is met. We analyzed the performance of OSSS by obtaining closed form expressions for the average SNR of the primary system. In order to validate its efficiency, we also analyzed an interference limited scheme (AF-SC) and compared it with OSSS. Simulation results showed that performance of OSSS can always be better than AF-SC for both the primary system and WSN.

Figure 2.7. Average received SNR of secondary transmission for various values of $\frac{P_s}{\sigma^2}$ for OSSS and AF-SC.

A Testbed for Cooperative Spectrum Sharing

3.1 Introduction

In Chapter 1, we gave theoretical background related to CSS protocols whereas in Chapter 2 we proposed an interference free CSS protocol, namely OSSS, which mitigates the interference from primary system to secondary system and vice versa. As discussed in Chapter 1, most of the research devoted to CSS and cognitive radio in general, are related to theoretical performance evaluation and simulation, hence its very difficult to gauge their performance in realistic environment. To aggravate the problem even further, the few testbeds that are available are mainly concentrating on “detect-and-avoid” interweave schemes for cognitive radio [6] - [11].

In this chapter, as a initial proof-of-concept demonstration and performance assessment of CSS protocols, we have implemented the protocol proposed in Chapter 1, Subsection 1.4.4.3 on a software reconfigurable National Instrument (NI)PXIe platform [105]. Field trials were conducted in an indoor office environment to obtain quantitative as well as qualitative results. The quantitative results are obtained by measuring the packet error rates (PER) for both primary and secondary systems during different times of the day. The secondary spectrum access probability, which represents the probability that ST is able to assist PT and thus gain

access to the spectrum, is also measured. Qualitative results are obtained by using the above protocol to transmit two different images from PT to PR and ST to SR respectively. As a benchmark for comparison, we consider a scenario without the secondary system whereby PT will adopt a simple ARQ retransmission protocol with MRC if decoding at PR fails. For the rest of this chapter, we will call this benchmark as the PT-PR retransmission protocol.

The measurement results proved that the proposed spectrum sharing protocol is practically viable and ST is able to achieve secondary spectrum access without degrading the PER of the primary system when compared to the PT-PR retransmission protocol. To the best of our knowledge, this chapter presents the first publicly available measurement results for CSS protocols.

Apart from the above, following are some other properties and contributions of this prototype testbed

- Unlike some of the past wireless testbeds on cognitive networks that utilize off-the-shelf wireless cards or open source software and hardware platforms [5],[75], [104], this testbed is the state-of-art development with all the protocols and algorithms required for implementation developed in-house.
- This testbed is based on software reconfigurable National Instruments PXIe platform hence most of the parameters are user configurable. It also has the flexibility to implement PHY layer and build PHY/MAC cross-layer protocols.
- It uses modular design and graphical programming environment to develop sophisticated algorithms, hence the implementation is user friendly and easily reproducible anywhere else.
- The advantages of cooperative spectrum sharing at different environment and settings can be practically validated.

3.2 Design and Implementation

The testbed consists of 4 radio nodes, namely PT, PR, ST and SR. Each of these nodes is implemented on the software programmable National Instruments (NI)

Figure 3.1. System model for cooperative spectrum sharing

PXIe platform. The physical layer modem is programmed to follow the OFDM standards in IEEE 802.11a [106] and the key parameters are given in Table 3.1. All the nodes are transceivers operating in time division duplex (TDD) mode, except for SR which only needs to receive data. The nodes are time synchronized through their respective CPU clocks. This time synchronization is necessary for the execution of the two-phase cooperative relaying protocol.

Let x_p be the primary signal transmitted by PT in Phase I. Denoting the signals received by PR, ST, and SR in Phase I as y_{11} , y_{21} , and y_{31} respectively, we have

$$y_{i1} = \sqrt{P_p} h_i x_p + n_{i1} \quad (3.1)$$

where $i = 1, 2, 3$. P_p is the transmit power at PT, h_i is the channel coefficient between the respective transmitters and receivers, and $n_{i1} \sim \mathcal{CN}(0, \sigma^2)$ is the additive white Gaussian noise (AWGN) at the respective receivers for Phase I. After recep-

tion in Phase I, ST attempts to decode x_p . If the decoding is successful (successful decoding is checked by CRC) and ST is able to assist PT¹, ST regenerates x_p . A composite signal z_s is created by superimposing the regenerated primary signal x_p with power αP_s and the secondary signal x_s with power $(1 - \alpha)P_s$ where P_s is the total transmit power at ST. Thus

$$z_s = \sqrt{\alpha P_s} x_p + \sqrt{(1 - \alpha) P_s} x_s. \quad (3.2)$$

As a proof of concept and for ease of implementation, α will not be adaptively chosen as in [16]. Based on the analysis in [16], we have heuristically chosen the value of α to be 0.75, which offers a good compromise between the performance of primary and secondary system. As shown in the measurement results later, even with a non-adaptive α , the proposed spectrum sharing protocol still outperform the PT-PR retransmission protocol. In Phase II, z_s is broadcasted by ST and received by PR and SR. The signal received at PR is given by

$$y_{12} = (\sqrt{\alpha P_s} h_4) x_p + (\sqrt{(1 - \alpha) P_s} h_4) x_s + n_{12} \quad (3.3)$$

where h_4 is the channel coefficient between ST and PR, and $n_{12} \sim \mathcal{CN}(0, \sigma^2)$ is the AWGN at PR in Phase II. Signals y_{11} and y_{12} are then combined at PR using MRC for the decoding of x_p . At SR, the primary signal received in Phase I is downconverted to baseband samples and a soft estimate $\hat{x}_p = \frac{y_{31}}{\sqrt{P_p} h_3}$ is generated and stored in memory. In Phase II, the signal received at SR is

$$y_{22} = (\sqrt{\alpha P_s} h_5) x_p + (\sqrt{(1 - \alpha) P_s} h_5) x_s + n_{22} \quad (3.4)$$

where h_5 is the channel coefficient between ST and SR and $n_{22} \sim \mathcal{CN}(0, \sigma^2)$ is the AWGN at SR in Phase II. The estimate $\hat{x}_p$ stored in memory at SR in Phase I is then used to cancel out the interference component $\sqrt{\alpha P_s} h_5 x_p$ from (3.4) to obtain $y'_{22} = y_{22} - (\sqrt{\alpha P_s} h_5) \hat{x}_p$.

¹On the other hand, if ST fails to decode x_p in Phase I, it will simply remain silent in Phase II.

Table 3.1. Physical layer parameters

Modulation	QPSK
IFFT / FFT length	64
RF bandwidth	20MHz
Subcarrier spacing	312.5kHz
OFDM symbol duration	3.2usec
Guard interval	0.8usec
Total frame duration	48usec
OFDM symbols per frame	8
Data bits per frame	368
Carrier frequency	1.4GHz
Channel code	Convolutional code (7, [133 171])
Code rate	1/2

3.2.1 Implementation

3.2.1.1 Implementation of primary system

The architecture for PT and PR is similar. Each primary node (PT or PR) consists of a basic transmitter-receiver structure. The hardware set-up is shown in Fig. 3.2.

All digital baseband signal processing is programmed into the NI PXIe-8130 controller. In the transmitter, the digital baseband signals are first sent to NI PXIe-5450 which acts as the DAC, before RF upconversion by NI PXIe-5611. Similarly, in the receiver, RF downconversion is performed by NI PXIe-5601, followed by NI PXIe-5622 which acts as the ADC. Besides the physical layer signal generation and transmission, a simple MAC layer consisting of ACK and NACK control messages is also implemented to emulate the CRTS and CCTS messages.

3.2.1.2 Implementation of secondary system

The basic architecture for the secondary system is similar to that for the primary system, with the exception that ST has to operate as a DF relay and generate the composite signal (Fig. 3.3).

Figure 3.2. NI PXIe Hardware (Transmitter and Receiver)

3.2.1.3 OFDM Frame structure

The frame structures for the data and control frames (NACK/ACK) are given in Fig. 3.4.

A typical OFDM frame originating from PT and ST consists of a short preamble, long preamble, data header and data. The short preamble and long preamble are used for timing synchronization and channel estimation respectively [106]. A 28 bit data header is appended at the start of the data to distinguish between PT and ST frames. Similarly, NACK and ACK headers are used to distinguish between NACK and ACK frames respectively².

²NACK/ACK messages are encoded with 14 times repetition coding and transmitted with a power of 2dBm which is 2dB higher than the maximum transmit power used for data transmission to ensure their virtually error free detection at the respective receivers.

Figure 3.3. Architecture of ST and SR

3.3 Measurement

3.3.1 Protocol Flowchart

As a proof-of-concept measurement, we have slightly modified the control message exchange protocol described in Section 1.4.4.2. To simplify the exchange of CRTS, CCTS and CCTH, we will replace them with simple ACK and NACK messages transmitted by PR.³ The protocol flowchart for the transmission of a single data

³Do note that this simplification is done to simplify the implementation and it does not compromise the viability and performance of the proposed spectrum sharing protocol.

Figure 3.4. Frame structure

frame from PT is shown in Fig. 3.5. During the measurement, this flowchart is continuously repeated until the required number of frames for the calculation of PER has been transmitted. In the flowchart, we have used indicator functions PE_p , PE_s and SA to denote the packet error and spectrum access opportunity as follows,

Primary system

$$PE_p = \begin{cases} 0 & \text{no packet error} \\ 1 & \text{packet error exists} \end{cases},$$

Secondary system

$$PE_s = \begin{cases} 0 & \text{no packet error} \\ 1 & \text{packet error exists} \end{cases},$$

$$SA = \begin{cases} 0 & \text{no spectrum access} \\ 1 & \text{obtain spectrum access} \end{cases}.$$

Figure 3.5. Measurement flowchart

PE_p and PE_s for every iteration will be recorded and the average is taken at the end of the measurement to obtain the PER. Similarly, SA is recorded after every iteration and the average is taken at the end to obtain the spectrum access probability for the secondary system.

The flowchart is first initiated by PT transmitting a data frame to PR. This transmission is also overheard by ST and SR. If decoding at PR is successful, it transmits an ACK message and this will initiate the next iteration of the flowchart. This event will be recorded as $PE_p = 0$. When ST receives this ACK message, it will understand that PT does not require any assistance for the transmission and thus it does not have an opportunity to access the spectrum, i.e. $SA = 0$.

However, if decoding at PR fails, it will transmit a NACK message. Upon reception of the NACK message, PT automatically switches to the two-phase cooperative relaying mode.⁴ At the same time, ST checks whether it has successfully decoded the primary packet. If yes, ST generates the composite signal and transmits it in Phase II. Thus, ST has gained a spectrum access opportunity by relaying the primary signal, i.e. $SA = 1$.⁵ PR will then perform a MRC with the packets received from PT and ST, and SR will perform interference cancellation to retrieve the secondary signal. Depending on the decoding results, PR and SR will update PE_p and PE_s respectively.

On the other hand, if ST fails to decode the primary packet, it will simply remain silent and there will be no transmission in Phase II. This event will be recorded as $PE_p = 1$ and $SA = 0$. The next iteration of the flowchart will then be initiated.

3.3.2 A Benchmark: PT-PR Retransmission Protocol

In order to have a meaningful benchmark for comparison, we consider a conventional automatic repeat request (ARQ) protocol for the primary system. If PT receives a NACK from PR after Phase I, it retransmits the same data in Phase II and PR performs MRC after receiving the data from the two phases. This protocol represents the performance of a conventional system with ARQ (1 retransmission)

⁴This NACK acts as both the CRTS and CCTS.

⁵Thus CCTH is not required as ST will always assist PT upon the reception of NACK as long as it has successfully decoded the primary signal in Phase I.

Figure 3.6. PT-PR retransmission protocol flowchart

in the absence of secondary system. The protocol flowchart for PT-PR retransmission protocol shown in Fig. 3.6.

3.3.3 Measurement Set-Up

The measurements were taken in a computer laboratory within our Positioning & Wireless Technology Center (PWTC) which is located at the 4th floor of Research Techno Plaza in the campus of Nanyang Technological University, Singapore. The floorplan of PWTC as well as the computer laboratory are shown in Fig. 3.7. The photo of measurement environment is shown in Fig. 3.8. PWTC is a typical

Figure 3.7. Floor plan of measurement environment.

modern office with width 27.3m, length 27.3m and height 2.8m. The laboratory has a number of workstations and benches for staff and students. The total workforce of the centre is around 120. The working hours in the centre is from 8.30 to 5pm, but some people prefer to stay late in the evening till 20:00 - 20:30. On a typical working day, an average of 70 people move in and out of the computer laboratory. The peak movement of people is observed from 10:00 to 14:00 as most of the students prefer to do their project work during this period. The environment in the laboratory is fairly dynamic due to consistent human traffic in the lab. All the antennas for PT, PR, ST and SR were installed at a height of 1.4m from floor level. The path between PT and PR is obstructed due to non line of sight (NLOS) condition whereas ST is in line of sight (LOS) with both PT and PR. SR is located in LOS with PT and ST.

3.3.3.1 Path loss between the nodes

The instantaneous path loss in the present set-up can be obtained by subtracting the received power from the transmit power. If transmit power is P_p dBm and received power is P_r dBm, the path loss, P_L (in dB) would be $P_L = P_p - P_r$. This path loss gives the estimate of total loss in dB between transmitter and receiver.

The average path loss can be calculated by keeping P_p constant and measuring P_r over the measurement period (8:00-22:00 in this case). Since, P_r changes over time (due to human traffic or other factors) the path loss will change over the measurement period. Hence an average is taken over the measurement period to calculate the average path loss.

The average path loss values between PT-PR, PT-ST, PT-SR, ST-SR and ST-PR for the current set-up is 62.5dB, 53.5dB, 52dB, 47.5dB and 55dB respectively.

3.4 Measurement Results

3.4.1 Qualitative results

Fig. 3.9 and Fig. 3.10 shows the primary system and secondary system images that are sent to PR and SR respectively. Fig. 3.11 and Fig. 3.12 shows the received images at PR and SR after the two phase transmission. In order to observe a

Figure 3.8. Measurement environment

Figure 3.9. Primary system transmit image

Figure 3.10. Secondary system transmit image

qualitative performance with our proposed spectrum sharing scheme, no channel coding was used for the transmission of images. As observed, the image received at PR after the proposed two phase transmission and MRC is better than that obtained with direct transmission (From PT). Moreover, SR is also able to decode the image of ST albeit with some degradation. Therefore, the proposed protocol is beneficial for both the primary as well as secondary system since it helps to improve the QoS of primary system and it also allows secondary spectrum access.

3.4.2 Quantitative results

Fig. 3.13 - 3.18 shows the quantitative performance achieved with the proposed scheme. Note that channel coding was used in order to emulate actual PER performance.

3.4.2.1 Packet error rate measurements

Fig. 3.13 shows the PER for the primary system with and without the secondary system. Conventional primary system (CPS) in Fig. 3.13 denotes the PER for the primary system without ARQ and secondary system. In the measurements, we let $P_p = P_s$ and the measurements are repeated with different transmit powers to obtain the data points for each of the average received SNR in Fig. 3.13 -

Figure 3.11. Image received at PR

3.15. Each PER plot is calculated by averaging over 10000 data frames. The average received SNR at SR and PR is calculated by taking the average of the SNR received from PT. As shown in Fig. 3.13, the performance of the proposed cognitive protocol is better than the PT-PR retransmission protocol. Thus we can achieve spectrum access for the secondary system without compromising the performance of primary system. This would be useful in scenarios in which the channel between PT and PR is bad and the primary system desires to improve its performance without any additional increase in transmit power. It can request for relay assistance which will allow the secondary system to gain spectrum access. To elaborate this point further, it will be insightful to observe the performance of CPS and PT-PR retransmission protocol in the low SNR region. It can be observed from Fig. 3.13 that at low SNR ($\text{SNR} \leq 7.5\text{dB}$), PER is almost 1 for the case of CPS

Figure 3.12. Image at SR after interference cancellation

and PT-PR retransmission protocol, thus no meaningful communication is possible between PT and PR and the link is completely broken. In this scenario, assistance from ST can be received and a $PER \approx 0.4$ can be achieved by the proposed scheme. This will in turn provide spectrum access opportunities to the secondary system.

Fig. 3.14 shows the spectrum access probability (SAP) for the secondary system for different values of SNR at PR. It is obvious from this figure that at low SNR values the spectrum access probability is fairly high. Thus when the channel between PT and PR is bad and primary PER is high, there are more chances for secondary system to gain spectrum access. Another observation is that the SAP is fairly constant for $SNR \leq 7.5\text{dB}$. This is because PER for PT-PR is close to 1 for $SNR \leq 7.5\text{dB}$, thus ST has approximately similar chances of accessing the spectrum. When the SNR is between 7.5dB and 11dB, SAP reduces to 0.35. As the SNR increases further, PT-PR link improves and the secondary system will have a decreasing chance of accessing the spectrum.

The PER for the secondary system is shown in Fig. 3.15. A meaningful PER of around 0.1 to 0.4 is achievable at SR. Please note that the PER performance at SR is independent of PT-PR link and depends solely on PT-SR and ST-SR links.

The PER performance of the primary system and secondary system with the proposed scheme for different values of α is given in [25]. The general trend observed was that as the value of α increases PER for the primary system decreases

Figure 3.13. Packet error rate for the primary system.

with a corresponding increase in PER for the secondary system. This is obvious as the greater the value of α , the higher is the fraction of power allocated by ST to the primary signal in Phase II.

3.4.2.2 Hourly measurements

Fig. 3.16 and Fig. 3.17 show the PER for primary system during different times of the day for $P_p = -2\text{dBm}$ and $P_p = -8\text{dBm}$ respectively. From Fig. 3.16 it can be observed that PER is highest during 12:00 to 14:00, as the official lunch

Figure 3.14. Spectrum access probability for the secondary system.

break falls during this period. This is when most people in the research centre access the washroom by passing through exit A and B of the computer laboratory. For Fig. 3.17, the human traffic has no significant impact on the PER of CPS and PT-PR retransmission protocol, and their performance is limited by the received SNR at PR. A trend similar to PER can be observed for the spectrum access probability plot for $P_p = -2\text{dBm}$ and $P_p = -8\text{dBm}$ shown in Fig. 3.18. The SAP for the secondary system for $P_p = -2\text{dBm}$ peaks around 10:00-14:00 then gradually decreases as the day progresses. From the measurement results, it can be seen that the proposed CSS protocol is most effective in crowded situations when the human traffic is high.

Figure 3.15. Packet error rate for the secondary system.

3.5 Discussion and Key Lessons Learned

After working on the experimental set-up of the above project for the past four years, a number of factors that should be taken into account while designing and implementing such a system have become apparent.

Hardware and Software

Foremost important is the selection of hardware and software on which one intends to implement a CSS system. We choose to implement our CSS protocol on NI PXIe devices which inherently supports NI LabVIEW 2009 graphical programming language. While selecting the programming language and platform, the following

Figure 3.16. Packet error rates for primary system during the different times of the day, $P_p = -2\text{dBm}$.

factors should be taken into account.

- Synchronization between the various nodes: This is very important if a two-phase cooperative relaying protocol is to be implemented. We achieved the synchronization by synchronizing the respective CPU clocks of all the nodes. Additional synchronization was achieved by using the wait function in LabVIEW 2009.
- Protocol Support: The hardware should be able to support the design

Figure 3.17. Packet error rates for primary system during the different times of the day, $P_p = -8\text{dBm}$.

parameters of CSS protocol. For e.g. it should support the selected sampling rate, modulation scheme, carrier frequency and transmitted power.

- **Response Time:** This is another important factor while implementing a CSS on the hardware. The hardware should have a minimal response time so that it can react spontaneously to any change in the environment.
- **Software Reconfigurability:** Hardware should be software reconfigurable, this allows flexibility in the design. Furthermore, additional features can be in-

Figure 3.18. Spectrum access probability for the secondary system during the different times of the day.

incorporated in the design in the future without any modification in hardware.

- **Robustness:** There should not be any significant drift in performance of hardware with respect to change in room temperature and environmental conditions.

Physical layer modem

Physical layer modem plays a crucial role in designing a cognitive spectrum sharing protocol. We chose to follow the OFDM standards in IEEE 802.11a as this is one of the well established and universally accepted standard for a point-to-point wireless communication. Overlaying the proposed scheme on this standard helps to calibrate the performance of the proposed scheme in comparison with a conventional well established standard. In the proposed implementation, all the nodes follow the same physical layer modem standard.

Placement of nodes

The node placement plays an important part in implementing a cognitive spectrum sharing scheme. As proven in [16], the placement of ST with respect to PT and PR is an important factor in deciding whether the proposed scheme will assist the secondary system to gain spectrum access. The results may vary depending upon the location of nodes with respect to each other.

3.6 Conclusion

A RF testbed for a cognitive spectrum sharing protocol based on cooperative relaying was designed and implemented on a NI PXIe hardware. The testbed has been successfully used to demonstrate the viability of the proposed spectrum sharing protocol to achieve spectrum access for the secondary system without causing any degradation in the performance of the primary system. The performance of both systems were quantified by measuring their PER and also spectrum access probability for the secondary system.

Analytical Evaluation of Impact of Nonlinear HPA on an OFDM Communication System

4.1 Introduction

In Chapters 1-3, we discussed the CSS protocols and showed through theoretical derivations as well as measurement results how CSS can be utilized by the secondary systems to gain spectrum access without compromising the performance of primary systems. Specifically in Chapter 3, we used OFDM physical layer to demonstrate the performance of a CSS protocol in a realistic environment.

Over the past decade much research has been devoted to the field of multicarrier transmission exemplified by OFDM. Due to its efficient utilization of available spectrum bandwidth as well as its inherent capability to alleviate the effects of frequency selective fading, OFDM has gained an edge over single carrier systems, thus being the natural choice of the current and upcoming broadband wireless communication standards such IEEE 802.11 WiFi, IEEE 802.16 WiMax, Advanced LTE, etc. As a consequence, most of the research and development on CSS, spectrum sensing and cooperative relay communication use OFDM as the physical layer technology to demonstrate the viability of their proposed algorithms [5],[6],[92]- [94], [103], [107]-[109].

However, as stated by authors in [110]-[116] the RF receivers front end nonlinearity (caused mainly due to nonlinear HPA) and dynamic range has significant impact on the performance of an OFDM system and might give misleading measurement results if not properly compensated or rectified. Furthermore, OFDM has high peak to average power ratio (PAPR) [117] thus making it much more sensitive to nonlinear distortion as compared to single carrier systems.

Another factor that affects the performance of an OFDM system is the memory effect present in a HPA [118]. A HPA typically exhibits memory effects, especially when driven with wideband signal like OFDM, and its effects on the nonlinear distortions of OFDM waveforms are expected to be fundamentally different from what is currently known by considering memoryless HPA [119]-[121].

In this chapter, by taking the above issues into consideration, we propose an analytically methodology to evaluate the impact of nonlinear HPA with memory on the performance of an OFDM system. Since OFDM forms the backbone of our implementation of CSS protocols, and HPA forms an indispensable part of the measurement, the proposed analysis will help us to estimate the back-off required in a HPA while obtaining measurement results for our testbed in Chapter 3. As shown later in this chapter, a improper selection of back-off has considerable impact on the SER performance of an OFDM system. The theoretical analysis developed in this chapter will also be an efficient and convenient tool to gauge the performance of an OFDM signal impaired by a nonlinear HPA with memory, without resorting to extensive simulations.

The behavioral model of HPA considered for theoretical analysis is a memory polynomial model, which is a truncated form of the Volterra series. In the theoretical framework developed, this chapter shows that nonlinear HPA considerably degrades the performance of OFDM in terms of symbol error rate (SER), and the distortion itself can be canonically characterized by a complex attenuation component and nonlinear noise component. Closed-form expressions for SER in additive white Gaussian noise (AWGN) channel is derived and the SER for fading channel is approximated by the adaptive Gauss-Kronrod (G-K) quadrature method. Simulation results are shown for a realistic HPA, based on the Wiener-Hammerstein model, and compared with the analytical results to validate the proposed analysis.

4.1.1 Impact of a nonlinear HPA on a two-tone signal

Before analytically evaluating the impact of nonlinear HPA on a multicarrier OFDM signal, let us first consider the impact of nonlinear HPA on a two tone input signal $x(t)$.

$$x(t) = \cos(\omega_1 t) + \cos(\omega_2 t) \quad (4.1)$$

The HPA is modeled as a odd order nonlinear polynomial model with a maximum order of nonlinearity restricted to three. When $x(t)$ is input to such a HPA its output will be

$$\begin{aligned} y(t) &= \alpha_1 x(t) + \alpha_3 \{x(t)\}^3 \\ &= \alpha_1 \{\cos(\omega_1 t) + \cos(\omega_2 t)\} + \alpha_3 \{\cos(\omega_1 t) + \cos(\omega_2 t)\}^3 \\ &= \alpha_1 \{\cos(\omega_1 t) + \cos(\omega_2 t)\} + \alpha_3 \left[\frac{9}{4} \{\cos(\omega_1 t) + \cos(\omega_2 t)\} + \frac{1}{4} \{\cos(3\omega_1 t) + \cos(3\omega_2 t)\} \right. \\ &\quad \left. + \frac{3}{4} \{(\cos(2\omega_1 + \omega_2)t) + (\cos(2\omega_2 + \omega_1)t)\} \right. \\ &\quad \left. + \frac{3}{4} \{(\cos(2\omega_1 - \omega_2)t) + (\cos(2\omega_2 - \omega_1)t)\} \right] \end{aligned} \quad (4.2)$$

where α_1 and α_3 are the 1st order and 3rd order coefficients of a nonlinear power amplifier. Thus the output from HPA along with the desired frequency components (fundamental frequency components) will also contain third order inter modulation components(IM3) given by

$$\text{IM3} = \alpha_3 \frac{3}{4} \{(\cos(2\omega_1 - \omega_2)t) + (\cos(2\omega_2 - \omega_1)t)\} \quad (4.3)$$

IM3 can be divided into two bands¹,

$$\begin{aligned} \text{IM3, Upper side band (USB)} &= \alpha_3 \frac{3}{4} (\cos(2\omega_1 - \omega_2)t) \\ \text{IM3, Lower side band (LSB)} &= \alpha_3 \frac{3}{4} (\cos(2\omega_2 - \omega_1)t) \end{aligned} \quad (4.4)$$

This intermodulation components interferes with the desired signal thus distorting the signal. The input and output response of a nonlinear HPA to a two-tone input

¹Here we assume $\omega_1 > \omega_2$

Figure 4.1. Two tone input signal. F-1 and F1 represents the two dominant fundamental frequency components of two-tone input signal whereas df represent the frequency spacing between them.

signal is also shown in Fig. 4.1 and Fig. 4.2.

4.2 System Models for Analysis

4.2.1 OFDM signal model

OFDM is a form of frequency division multiplexing (FDM) [122]-[124] where data are transmitted in several narrowband streams (a.k.a subcarrier) at different carrier frequencies as shown in Fig. 4.3. However, unlike conventional FDM systems, where the subcarrier signals are separated by guard bands in frequency domain, OFDM systems allow overlapping of adjacent subcarriers while maintaining the orthogonality between them. As a result, OFDM are more spectrally efficient.

Figure 4.2. Response of a nonlinear HPA to two-tone input signal.

Moreover, in single-carrier systems ISI occurs and can only be compensated by using complex equalizers at the receiver. In a OFDM system, only one tap equalizer is required to overcome ISI.

Complex baseband samples of the m th OFDM symbol with N subcarriers and appended with guard interval of length N_g samples are expressed as

$$x^{(m)}[n] = \begin{cases} \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} a_k^{(m)} e^{j2\pi kn/N} & \text{for } -N_g \leq n \leq N-1 \\ 0 & \text{otherwise} \end{cases} \quad (4.5)$$

where n is the discrete time index and $a_k^{(m)}$ is the complex data symbol in the k th subcarrier of the m th OFDM symbol. Without loss of generality, we presume that $a_k^{(m)}$ are independent and identically distributed with zero mean and variance P .

By the Central Limit Theorem [125], when N is large $x^{(m)}[n]$ can be assumed

Figure 4.3. OFDM

to be a complex Gaussian process with zero mean and variance $P_{av} = P$.

4.2.2 Model for nonlinear HPA

Owing to the dynamic (frequency dependent) nature of the nonlinear distortions of HPA when driven by multicarrier signals, the conventional AM/AM and AM/PM static nonlinear functions are insufficient to characterize a HPA. Thus more elaborate models like Volterra series [126], [127] are required to model the HPA. The main problem associated with the Volterra series is the computational complexity involved in the measurement of the Volterra kernels [129]. Therefore, as a compromise between a full Volterra series and a memoryless nonlinear model, the Memory Polynomial Model (MPM) has been proposed in [129]. In [118], MPM was shown to be a good model for a HPA. In addition, frequency dependent characteristics (memory effects) present in a nonlinear HPA is more analytically tractable in a MPM as compared to other models [31]. Hence, in this paper we have chosen the MPM as a model for HPA for our theoretical analysis. When an input $x[n]$ is applied to a MPM with sparse delay taps, the output $y[n]$ can be expressed as [118]

$$y[n] = \sum_{d=1}^D \sum_{q=0}^Q \alpha_{2d-1,q} x[n - B_q] |x[n - B_q]|^{2(d-1)} \quad (4.6)$$

where $\alpha_{2d-1,q}$ is the complex coefficient for the $(2d - 1)$ th order nonlinearity of the q th delay tap, B_q is the number of delay samples for the q th delay tap, and Q is the total number of delay taps. If $Q = 0$ and $B_0 = 0$ then (4.6) reduces to a

memoryless nonlinear system [23]

$$y[n] = \sum_{d=1}^D \alpha_{2d-1} x[n] |x[n]|^{2(d-1)}. \quad (4.7)$$

4.3 Characterization of the Received Signal

The multi-path channel is modeled as a wide sense stationary uncorrelated scattering (WSSUS) Rayleigh fading channel and has a discrete impulse response,

$$h^{(m)}[n] = \sum_{l=1}^L h_l^{(m)} \delta[n - \tau_l] \quad (4.8)$$

where τ_l is the normalized discrete delay, L is the number of multipaths and $h_l^{(m)} \sim \mathcal{CN}(0, \sigma_l^2)$ is the channel response of the l th multipath. Assuming that N_g is larger than the sum of the maximum delay of the channel and memory length of the HPA i.e. $N_g \geq B_Q + \tau_L$, the received complex data symbol at k th subcarrier obtained after removal of the guard interval and Discrete Fourier Transform (DFT) can be written as

$$\hat{a}_k^{(m)} = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} \sum_{l=1}^L h_l^{(m)} y^{(m)}[(n - \tau_l)_N] e^{-j2\pi kn/N} + n_o^{(m)}[k] \quad (4.9)$$

where $(k)_N$ denotes the residue of k modulo N and $n_o^{(m)}(k) \sim \mathcal{CN}(0, N_0)$ is the AWGN in the k th subcarrier. Substituting (4.6) in (4.9), we obtain

$$\begin{aligned} \hat{a}_k^{(m)} = & \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} \sum_{l=1}^L h_l^{(m)} \sum_{d=1}^D \sum_{q=0}^Q \alpha_{2d-1,q} x^{(m)}[n - B_q - \tau_l] |x^{(m)}[n - B_q - \tau_l]|^{2(d-1)} e^{-j2\pi kn/N} \\ & + n_o^{(m)}[k]. \end{aligned} \quad (4.10)$$

For the ease of analysis that follows, we adopt a MPM with third-order nonlinearity i.e. $2D - 1 = 3$. From (4.10), we obtain

$$\hat{a}_k^{(m)} = \frac{1}{\sqrt{N}} \sum_{l=1}^L \sum_{q=0}^Q \hat{b}_{q,l}^{(m)}[k] + n_o^{(m)}[k] \quad (4.11)$$

where
$$\hat{b}_{q,l}^{(m)}[k] = \sum_{n=0}^{N-1} \sum_{d=1}^2 h_l^{(m)} \alpha_{2d-1,q} x^{(m)}[n - B_q - \tau_l] |x^{(m)}[n - B_q - \tau_l]|^{2(d-1)} e^{-j2\pi kn/N}.$$

Using the circular time shift property of DFT, we have

$$\begin{aligned} \hat{b}_{q,l}^{(m)}[k] &= h_l^{(m)} \alpha_{1,q} \sum_{n=0}^{N-1} x^{(m)}[n] e^{-j2\pi kn/N} e^{-j2\pi k(B_q + \tau_l)/N} \\ &\quad + h_l^{(m)} \alpha_{3,q} \sum_{n=0}^{N-1} x^{(m)}[n] |x^{(m)}[n]|^2 e^{-j2\pi kn/N} e^{-j2\pi k(B_q + \tau_l)/N}. \end{aligned} \quad (4.12)$$

After some manipulations (refer to Appendix B.1), (4.12) can be rewritten as

$$\hat{b}_{q,l}^{(m)}[k] = a_k^{(m)} \mu_{q,l}^{(m)}[k] + \eta_{q,l}^{(m)}[k] \quad (4.13)$$

where

$$\mu_{q,l}^{(m)}[k] = h_l^{(m)} \alpha_{1,q} e^{-j2\pi k(B_q + \tau_l)/N} + \frac{h_l^{(m)} \alpha_{3,q} e^{-j2\pi k(B_q + \tau_l)/N}}{N} \sum_{n=0}^{N-1} \gamma_n, \quad (4.14)$$

$$\eta_{q,l}^{(m)}[k] = \frac{h_l^{(m)} \alpha_{3,q} e^{-j2\pi k(B_q + \tau_l)/N}}{N} \left(\sum_{p=0, p \neq k}^{N-1} a_p^{(m)} \sum_{n=0}^{N-1} \gamma_n e^{j2\pi n(-k+p)/N} \right) \quad (4.15)$$

and $\gamma_n = |x^{(m)}[n]|^2$. Substituting (4.13) into (4.11), we obtain

$$\begin{aligned} \hat{a}_k^{(m)} &= \frac{1}{\sqrt{N}} \sum_{l=1}^L \sum_{q=0}^Q \hat{b}_{q,l}^{(m)}[k] + n_o^{(m)}[k] \\ &= \frac{1}{\sqrt{N}} a_k^{(m)} \sum_{l=1}^L \sum_{q=0}^Q \mu_{q,l}^{(m)}[k] + \frac{1}{\sqrt{N}} \sum_{l=1}^L \sum_{q=0}^Q \eta_{q,l}^{(m)}[k] + n_o^{(m)}[k] \\ &= \frac{1}{\sqrt{N}} a_k^{(m)} \sum_{l=1}^L \mu_l^{(m)}[k] + \frac{1}{\sqrt{N}} \sum_{l=1}^L \eta_l^{(m)}[k] + n_o^{(m)}[k] \end{aligned} \quad (4.16)$$

where $\mu_l^{(m)}[k] = \sum_{q=0}^Q \mu_{q,l}^{(m)}[k]$ and $\eta_l^{(m)}[k] = \sum_{q=0}^Q \eta_{q,l}^{(m)}[k]$.

From (4.16) it can be stated that the received symbol conditioned on $h_l^{(m)}$ after the DFT operation consists of an attenuation component $\sum_{l=1}^L \mu_l^{(m)}[k]$ multiplied to $a_k^{(m)}$, nonlinear noise component $\sum_{l=1}^L \eta_l^{(m)}[k]$ and AWGN. It can also be shown that the nonlinear noise component can be modeled as a zero mean complex Gaussian process uncorrelated with $a_k^{(m)}$ for a sufficiently large value of N (refer to Appendix B.1.1). The variance of the nonlinear noise conditioned on $h_l^{(m)}$ is given by (refer to Appendix B.1.2)

$$\sigma_{NL}^2[k] = \psi[k] \bar{\sigma}_{NL}^2[k] \quad (4.17)$$

where $\psi[k] = |\sum_{l=1}^L h_l^{(m)} e^{-j2\pi k \tau_l / N}|^2 = |H^{(m)}[k]|^2$ is the discrete channel frequency response and $\bar{\sigma}_{NL}^2[k] = \frac{|\sum_{q=0}^Q \alpha_{3,q} e^{-j2\pi B q k / N}|^2 P^3 (3N^2 - 5N + 2)}{N^4}$.

4.4 Average Symbol Error Rate in Rayleigh Fading Channel

From (4.16), the instantaneous SNR for each subcarrier conditioned on $h_l^{(m)}$ can be expressed as

$$SNR|_{h_l^{(m)}}[k] = \frac{P \cdot \frac{E}{a_k^{(m)}} \left[\left| \sum_{l=1}^L \mu_l^{(m)}[k] \right|^2 \right]}{\sigma_{NL}^2[k] + N_o} = \frac{P \psi[k] \bar{\mu}[k]}{\psi[k] \bar{\sigma}_{NL}^2[k] + N_o} \quad (4.18)$$

where $\bar{\mu}[k]$ is given in Appendix B.1.3. Using (4.18), we can express the SER for each subcarrier conditioned on $h_l^{(m)}$ for a M -QAM system as [130]

$$f(k, \psi[k]) \approx 2 \left(1 - (1/\sqrt{M}) \right) \operatorname{erfc} \left(\sqrt{\frac{3SNR|_{h_l^{(m)}}[k]}{2(M-1)}} \right). \quad (4.19)$$

The closed-form expression of SER for each subcarrier in a non-fading AWGN channel ($\psi[k] = 1$) [31] is thus given by (4.19). In practice, this closed-form result can be used as a convenient tool to evaluate the degradation in performance at the transmitter due to the nonlinear distortions, in the absence of the effects from the fading channel.

As stated in Section 4.3, the multi-path channel is modeled as a WSSUS, thus the discrete time channel taps are uncorrelated with each other and the probability density function of $\psi[k]$, $p_\psi(\psi)$, is the same for all k . Furthermore, $p_\psi(\psi) = (1/\theta)e^{-\psi/\theta}$ and $\theta = E[\psi[k]]$. Therefore the SER averaged across the fading channel can be expressed as

$$\begin{aligned} P_s[k] &= E_\psi[f(k, \psi)] = \int_0^\infty f(k, \psi) p_\psi(\psi) d\psi \\ &= \int_0^\infty f(k, \psi) (1/\theta) e^{-\psi/\theta} d\psi \end{aligned} \quad (4.20)$$

The average SER across all the sub-carriers can thus be calculated as

$$\bar{P}_s = \frac{1}{N} \sum_{k=0}^{N-1} P_s[k]. \quad (4.21)$$

The integral in (4.20) is the analytical expression of SER for each subcarrier in a frequency selective Rayleigh fading channel. We use the adaptive Gauss-Kronrod (G-K) quadrature method to numerically solve the integral in (4.20). Interested readers may also refer to [132] for a heuristic approach to approximate the integral in (4.20) with hypergeometric functions. The result in (4.20) allows us to evaluate the end-to-end SER performance of the OFDM system, inclusive of the effects from the fading channel.

4.5 Simulation Results

For validation of the theoretical analysis, we simulated two different HPA models.

4.5.1 Results with HPA modeled as a MPM with delay taps

In this section we show the results for four sets of extracted coefficients from a HPA modeled as a MPM [129], [118] as given in Table 4.1. The MPM for a HPA can be represented as a FIR filter with unit delay taps or sparse delay taps [118] where

Table 4.1. Coefficients for MPM

Model	Coefficients	Delay
NPM (Memoryless Model)	$\alpha_{1,0} = 0.9798 - 0.2887i, \quad \alpha_{3,0} = -0.2901 + 0.4350i,$	0
MPM1 (Unit Delay)	$\alpha_{1,0} = 1.4045 + 0.8149i, \quad \alpha_{3,0} = -0.2484 + 0.3705i,$	0
	$\alpha_{1,1} = -1.0554 - 2.3300i, \quad \alpha_{3,1} = -0.0517 + 0.0896i$	1
	$\alpha_{1,2} = 0.6497 + 1.2759i, \quad \alpha_{3,2} = 0.0289 - 0.0404i$	2
MPM2 (Unit Delay)	$\alpha_{1,0} = 1.0513 + 0.0904i, \quad \alpha_{3,0} = -0.0542 - 0.2900i,$	0
	$\alpha_{1,1} = -0.0680 - 0.0023i, \quad \alpha_{3,1} = 0.2234 + 0.2317i$	1
	$\alpha_{1,2} = 0.0289 - 0.0054i, \quad \alpha_{3,2} = -0.0621 - 0.0932i$	2
MPM3 (Sparse Delay)	$\alpha_{1,0} = 0.98 - 0.30i, \quad \alpha_{3,0} = -0.30 + 0.42i,$	$B_0 = 0$
	$\alpha_{1,1} = 0.06 + 0.03i, \quad \alpha_{3,1} = -0.02 + 0.05i$	$B_1 = 10$
	$\alpha_{1,2} = 0.02i + 0.08i, \quad \alpha_{3,2} = -0.01 - 0.08i$	$B_2 = 100$
	$\alpha_{1,3} = -0.01 + 0.02i, \quad \alpha_{3,3} = 0.02 - 0.01i$	$B_3 = 50$

the weights at each tap can be characterized by memoryless nonlinear polynomials. The input back-off (IBO) for a HPA is defined as

$$\text{IBO} = \frac{A^2}{P_{av}} \quad (4.22)$$

where A is the input amplitude for the maximum amplifier output power. Thus by adjusting the back-off we can control the operating point of the HPA. The theoretical SER for this case was calculated using (4.19). NPM in Table 4.1 is nonlinear polynomial model which represents a memoryless nonlinear HPA. Fig. 4.4 shows the comparison between the analytical and simulated values for the above model in a AWGN channel. As observed from Fig. 4.4 the nonlinear noise floor reduces as we increase IBO. From it is obvious that P_{av} is inversely related to IBO, thus as we increase the IBO, we are operating the HPA at a higher back-off resulting in a lower error floor and SER. MPM1 and MPM2 in Table 4.1 are two different memory polynomial models with unit delay taps which represents two different HPA with memory. Fig. 4.5 and Fig. 4.6 shows the comparison

Figure 4.4. Theoretical (solid lines) and simulation (marker points) results for symbol error rate for NPM, 16 QAM, $N = 512$ for different values of IBO (a) = 15dB (b) = 10dB (c) = 7.5dB (d) = 5dB (e) = 2dB in an AWGN channel.

between theoretical and simulated values of average SER for each subcarrier due to nonlinear noise only, i.e. $N_o = 0$. It is apparent from Fig. 4.5 and Fig. 4.6 that nonlinear distortion with memory affects each subcarrier differently, depending upon the coefficients of the memory polynomial models.

The fourth set of coefficients given in Table 4.1 represents a memory polynomial model with a three sparse delay taps, MPM3. This set of coefficients are introduced to highlight the sparse delay nature of memory effects present in a nonlinear HPA.

Figure 4.5. Theoretical and simulation results for average SER for each subcarrier in MPM1 due to nonlinear noise only, 16 QAM, $N = 256$, $N_g = 64$, IBO = 3dB.

Fig. 4.7 shows the comparison between theoretical and simulated values of average SER for each subcarrier due to nonlinear noise only, i.e. $N_o = 0$. Fig. 4.7 also underscores the fact that the nonlinear distortion due to memory effects is dependent on the subcarrier position, which is reflected by the different SER for each subcarrier. Fig. 4.8 shows the effect of nonlinear noise on the average SER of MPM3 for various values of IBO in an AWGN channel. Good agreement between the simulation and theoretical results in the above figures validates our theoretical

Figure 4.6. Theoretical and simulation results for average SER for each subcarrier in MPM2 due to nonlinear noise only, 16 QAM, $N = 256$, $N_g = 64$, IBO = 3dB.

analysis for a OFDM system operating with a nonlinear HPA.

4.5.2 Results with HPA modeled as a Wiener-Hammerstein (W-H) model

To replicate a more realistic scenario for a HPA we simulated HPA which is modeled by the Wiener-Hammerstein (W-H) model. A W-H model is a linear time-invariant (LTI) system followed by a memoryless nonlinearity, which in turn is followed by

Figure 4.7. Theoretical and simulation results for average SER for each subcarrier in MPM3 due to nonlinear noise only, 16 QAM, $N = 512$, $N_g = 128$, IBO = 8dB.

another LTI system. The LTI blocks before and after the memoryless nonlinearity are respectively chosen as

$$H(z) = \frac{1 + 0.2z^{-2}}{1 - 0.1z^{-1}}, \quad G(z) = \frac{1 - 0.1z^{-2}}{1 - 0.2z^{-1}}. \quad (4.23)$$

Figure 4.8. Theoretical (solid lines) and simulation (marker points) results for average SER in AWGN channel for MPM3, 16 QAM, $N = 512$, $N_g = 128$ with different values of IBO (a) = 15dB (b) = 10dB (c) = 7.5dB (d) = 5dB (e) = 2dB.

For the memoryless nonlinearity we consider a TWTA [131] model which is given by

$$A[\rho] = \frac{\nu_A \rho}{1 + \psi_A \rho^2}, \quad \phi[\rho] = \frac{\nu_\phi \rho^2}{1 + \psi_\phi \rho^2} \quad (4.24)$$

where $A[\rho(t)]$ and $\phi[\rho(t)]$ denotes the normalized AM/AM and AM/PM characteristics respectively. ν_A, ψ_A, ν_ϕ and ψ_ϕ are selected to be 1, 0.25, 0.26 and 0.25

Table 4.2. Exponential power delay profile

Path (l)	Average Power (σ_l^2)[dB]	Delay (τ'_l)
1	$\sigma_1^2 = 0$,	$\tau'_1 = 0$
2	$\sigma_2^2 = -4.3$	$\tau'_2 = 1$
3	$\sigma_3^2 = -8.6$	$\tau'_3 = 2$
4	$\sigma_4^2 = -21.7$	$\tau'_4 = 5$
5	$\sigma_5^2 = -43.4$	$\tau'_5 = 10$

respectively [133].

For comparison with MPM, the coefficients for our third-order MPM were extracted from the W-H model by the least-squares method proposed in [118]². The frequency selective Rayleigh fading channel was simulated using a 5-tap model with an exponential power delay profile [134] given in Table 4.2. The theoretical SER for the OFDM system in an AWGN and frequency selective fading channel was calculated using (4.19) and (4.21) respectively.

Fig. 4.9 and Fig. 4.10 show the SER of an OFDM system with a HPA with memory for different IBO in an AWGN and frequency selective Rayleigh fading channel respectively. Simulation results are shown for the W-H model as well as the MPM model. A slight discrepancy observed between the simulation results of the W-H model and MPM model is due to the inability of the third order MPM to capture all the nonlinear distortion in the W-H model. However, MPM can still be considered a reasonably good model to represent a nonlinear HPA with memory given its analytical tractability [129], [118].

Good agreement between the theoretical estimates and the simulation results for MPM validates the analytical results obtained in this paper. This analytical results will be an useful tool to gauge the performance of practical OFDM systems

²Please refer to Appendix B.2 for a procedure to extract MPM coefficients from an actual HPA

Figure 4.9. Theoretical (solid lines) and simulation (marker points) results for average SER in AWGN channel for a HPA with memory, 16 QAM, $N = 512$, $N_g = 128$ with different values of IBO, (a) = 6dB (b) = 10dB (c) = 13dB (d) = 15dB.

impaired by nonlinear HPA with memory thus avoiding a need to perform extensive simulations.

The corresponding simulation results based solely on the TWTA model in (4.24) for a memoryless HPA are also plotted in Fig. 4.9 and Fig. 4.10. It is obvious from Fig. 4.9 and Fig. 4.10 that memory in HPA does indeed degrade the performance further. In particular, memory in HPA increases the noise floor. However, it can also be observed from Fig. 4.10 that as long as the IBO is large enough e.g. 15dB

Figure 4.10. Theoretical (solid lines) and simulation (marker points) results for average SER in frequency selective Rayleigh fading channel for a HPA with memory, 16 QAM, $N = 512$, $N_g = 128$ with different values of IBO, (a) = 6dB (b) = 10dB (c) = 13dB (d) = 15dB.

in Fig. 4.10, SER degradation due to memory is negligible.

Another important observation is that the SER performance is very sensitive to the IBO - a slight difference of a few dB will result in a large variation in SER performance, and for intermediate values of IBO (e.g. 13dB in Fig. 4.10), the degradation caused by memory is more pronounced. Thus, it is important that special attention be paid to the selection of IBO in the design of practical systems.

4.6 Conclusion

In this chapter, an analytical methodology to assess the SER of OFDM signals distorted by nonlinear HPA is proposed. The HPA is modeled by a memory polynomial model, which is a truncated form of Volterra series. It was proven that the distortion due to a HPA with memory is composed of a complex attenuation component, and a nonlinear noise component which can be modeled as a complex Gaussian process uncorrelated with the input.

As a consequence of the above analysis, it has been shown that the nonlinearity in HPA contributes significantly to the distortion of an OFDM signal and thus it should be taken into consideration while designing any practical OFDM systems. It has also been observed that the IBO for a nonlinear HPA should be selected meticulously as it has considerable impact on the SER performance of an OFDM system.

Conclusions and Future Work

Cognitive radio has often been touted as a novel and promising technology that will resolve the spectrum scarcity problem faced by emerging wireless systems and services. Although cognitive radio technology is well established in theory and simulation, its performance in real wireless environment is still relatively unknown. Through measurement results in a real world environment, the work in this thesis has proved, to a certain extent, that cognitive radio can be a viable technology to efficiently utilize the scarce radio spectrum for future wireless networks.

The research work done in the thesis can be categorized into four parts. In the first part fundamentals of wireless communication were discussed. Diversity techniques to combat fading in a wireless channel were investigated. Cooperative diversity in which different users or nodes in a wireless network cooperate to form a virtual antenna array was discussed as an alternative to transmit diversity. The problems with the current spectrum allocation policies for wireless systems were highlighted and utilization of cognitive radios to alleviate this problems were also discussed. By using cooperative spectrum sensing and CSS protocols, we demonstrated that cognitive and cooperative techniques are complementary to each other, and future wireless systems should utilize this synergy to achieve a high performance, high efficiency wireless network. A brief overview of the existing experimental deployments of cognitive radio testbeds was also presented.

In the second part, an OSSS based on CSS was proposed. It was shown how an OSSS utilizes the cooperative and cognitive techniques to effectively cancel out the interference from the primary system to secondary system and vice versa. The

obtained analytical and simulation results showed that unlike the conventional CSS protocols, in OSSS there is no compromise between the performance of primary and secondary systems, and the secondary system is always able to achieve spectrum access as long as it is willing to increase its transmit power such that SNR_T is met. The performance of primary and secondary systems were quantified in terms of average SNR.

In the third part, a RF testbed based on CSS was designed and implemented on a software reconfigurable NI PXIe hardware. Quantitative and qualitative results were shown to demonstrate the viability of the testbed to gain spectrum access for the secondary system while maintaining the performance of the primary system. The performance of both systems were quantified by measuring their PER under two different conditions.

1. By varying the transmit power and measuring the PER for different received SNR. This demonstrated the effect of transmit power on the PER values.
2. Keeping the transmit power constant and taking PER values during different times of the day. This demonstrated the effects of human traffic on the measured PER values.

The spectrum access probability for the secondary system was also measured.

In the fourth part, an analytical methodology to assess the the impact of distortion due to nonlinear HPA on the design and implementation of an OFDM communication system is proposed. The excellent agreement between the simulated results and the analytically obtained expressions authenticated the theoretical analysis. It was also proven that the distortion due to a HPA is composed of a complex attenuation component, and a nonlinear noise component which can be modeled as a complex Gaussian process uncorrelated with the input. Furthermore, it was also observed that the IBO for a nonlinear HPA should be selected meticulously as it has considerable impact on the end-to-end SER performance of an OFDM system.

5.1 Future work

Finally we discuss some issues which need further investigations. These can be interesting directions for future work.

- For the measurement results, only one secondary (cognitive) system is considered. In practice, there can be multiple secondary system fighting for the under-utilized resources of the primary system. One possible scenario can be two secondary system operating with one primary system. The primary system might optimize its performance by selecting the secondary system which has better channel conditions (e.g. the one which is closer to PT and PR) and thus the selected secondary system will obtain the spectrum access.
- It will be interesting to observe the performance enhancement (if any) when the transmit power of ST is increased by keeping the transmit power of PT constant. Another point which can be exploited for future research is varying the power allocation factor (α). Some work on this direction has already been done in [25]. Further investigation might be needed to explicitly establish the effect of α on the performance of primary and secondary system.
- The performance of the CSS protocol on RF testbed has been obtained by taking the measurement results in the indoor office environment. For future research an outdoor environment should be considered. In outdoor environment, shadowing and path loss might become dominant factors. It would be interesting to obtain the performance assessment in outdoor environments.
- The impact of location of PT, PR, ST and SR for the OSSS should be further investigated. Only single transmitting and single receiving antennas are considered for PT,PR, ST and SR in this thesis. Multiple transmitting/receiving antennas can provide the array gain and diversity gain and hence improve the performance. In future work, multiple antennas should be investigated.

Appendix A

Derivation for Average SNR of primary system with OSSS

From (2.12) and (2.13), we obtain

$$\begin{aligned}\overline{\text{SNR}}_p &= \frac{P_p}{\sigma^2} \text{E} \left[\frac{\gamma_1 + 2g^2\gamma_2\gamma_3 - 2\sqrt{2}\text{Re}(h_2h_3h_1)g}{(2g^2\gamma_3 + 3)} \right] \\ &= \frac{P_p}{\sigma^2} (\delta_1 + \delta_2 - \delta_3)\end{aligned}\quad (\text{A.1})$$

where $\delta_1 = \text{E} \left[\frac{\gamma_1}{2g^2\gamma_3 + 3} \right]$, $\delta_2 = \text{E} \left[\frac{2g^2\gamma_2\gamma_3}{2g^2\gamma_3 + 3} \right]$ and $\delta_3 = \text{E} \left[\frac{2\sqrt{2}g\text{Re}(h_3^*h_1h_2)}{2g^2\gamma_3 + 3} \right]$. δ_1 can be evaluated as

$$\delta_1 = \int_0^\infty \int_0^\infty \int_0^\infty \left(\frac{\gamma_1}{2g^2\gamma_3 + 3} \right) p_{\gamma_1}(\gamma_1) p_{\gamma_2}(\gamma_2) p_{\gamma_3}(\gamma_3) d\gamma_1 d\gamma_2 d\gamma_3 \quad (\text{A.2})$$

where $p_{\gamma_1}(\gamma_1)$, $p_{\gamma_2}(\gamma_2)$ and $p_{\gamma_3}(\gamma_3)$ are the probability density function (pdf) of γ_1 , γ_2 and γ_3 respectively. Additionally, $\gamma_i \sim \varepsilon(d_i^\nu)$, $i = 1, 2, 3$. Thus from A.2,

$$\begin{aligned}\delta_1 &= \int_0^\infty \int_0^\infty \frac{\gamma_1}{2g^2\gamma_3 + 3} p_{\gamma_2}(\gamma_2) p_{\gamma_3}(\gamma_3) d\gamma_2 d\gamma_3 \int_0^\infty \gamma_1 p_{\gamma_1}(\gamma_1) d\gamma_1 \\ &= \frac{1}{d_1^\nu} \int_0^\infty \int_0^\infty \frac{\gamma_1}{2g^2\gamma_3 + 3} p_{\gamma_2}(\gamma_2) p_{\gamma_3}(\gamma_3) d\gamma_2 d\gamma_3 \\ &= \frac{1}{d_1^\nu} \int_0^\infty p_{\gamma_3}(\gamma_3) \int_0^\infty p_{\gamma_2}(\gamma_2) \frac{1}{P_p\gamma_2 + \sigma^2\gamma_3 + 3} d\gamma_2 d\gamma_3.\end{aligned}\quad (\text{A.3})$$

Assuming $\frac{\sigma^2}{P_p} \approx 0$, then A.3 can be rewritten as

$$\begin{aligned}
\delta_1 &\approx \frac{1}{d_1^\nu} \int_0^\infty p_{\gamma_3}(\gamma_3) \int_0^\infty p_{\gamma_2}(\gamma_2) \frac{1}{\frac{P_s}{P_p \gamma_2} \gamma_3 + 3} d\gamma_2 d\gamma_3 \\
&= d_3^\nu \frac{P_p (-d_2^\nu P_s + 3d_3^\nu P_p - d_2^\nu P_s \ln(3) + d_2^\nu P_s \ln(d_2^\nu) + d_2^\nu P_s \ln(P_s) - d_2^\nu P_s \ln(d_3^\nu) - d_2^\nu P_s \ln(P_p))}{d_1^\nu (-d_2^\nu P_s + 3d_3^\nu P_p)^2} \\
&= d_3^\nu \frac{P_p \left\{ -d_2^\nu P_s + 3d_3^\nu P_p - d_2^\nu P_s \left[\ln \left(\frac{3d_3^\nu P_p}{d_2^\nu P_s} \right) \right] \right\}}{d_1^\nu (-d_2^\nu P_s + 3d_3^\nu P_p)^2}. \tag{A.4}
\end{aligned}$$

Similarly we can obtain

$$\begin{aligned}
\delta_2 &= \int_0^\infty \int_0^\infty \left(\frac{2g^2 \gamma_2 \gamma_3}{2g^2 \gamma_3 + 3} \right) p_{\gamma_2}(\gamma_2) p_{\gamma_3}(\gamma_3) d\gamma_2 d\gamma_3 \\
&\approx \int_0^\infty p_{\gamma_3}(\gamma_3) \int_0^\infty p_{\gamma_2}(\gamma_2) \frac{\frac{P_s}{P_p \gamma_2} \gamma_2 \gamma_3}{\frac{P_s}{P_p \gamma_2} \gamma_3 + 3} d\gamma_2 d\gamma_3 \\
&= \frac{P_s \left\{ -P_s^2 (d_2^\nu)^2 + 9d_3^{\nu 2} P_p^2 - 6d_3^\nu P_p P_s d_2^\nu \left[\ln \left(\frac{3d_3^\nu P_p}{d_2^\nu P_s} \right) \right] \right\}}{(-P_s d_2^\nu + 3d_3^\nu P_p)^3}, \tag{A.5}
\end{aligned}$$

and $\delta_3 = 0$. Thus substituting A.4 and A.5 in A.1 we obtain

$$\begin{aligned}
\overline{\text{SNR}}_p &= \frac{d_3^\nu P_p^2 \left\{ 3d_3^\nu P_p - d_2^\nu P_s - d_2^\nu P_s \left[\ln \left(\frac{3d_3^\nu P_p}{d_2^\nu P_s} \right) \right] \right\}}{d_1^\nu (3d_3^\nu P_p - d_2^\nu P_s)^2 \sigma^2} \\
&+ \frac{P_p P_s \left\{ 9d_3^{\nu 2} P_p^2 - P_s^2 (d_2^\nu)^2 - 6d_3^\nu P_p P_s d_2^\nu \left[\ln \left(\frac{3d_3^\nu P_p}{d_2^\nu P_s} \right) \right] \right\}}{(3d_3^\nu P_p - P_s d_2^\nu)^3 \sigma^2}. \tag{A.6}
\end{aligned}$$

Derivation of analytical results

B.1 Canonical Decomposition of Received Signal

Substituting (4.5) into (4.12), we can express $\hat{b}_{q,l}^{(m)}[k]$ as

$$\hat{b}_{q,l}^{(m)}[k] = h_l^{(m)} \alpha_{1,q} e^{-j2\pi k(B_q + \tau_l)/N} a_k^{(m)} + \beta_{q,l}^{(m)}[k] \quad (\text{B.1})$$

$$\text{where} \quad \beta_{q,l}^{(m)}[k] = \frac{h_l^{(m)} \alpha_{3,q} e^{-j2\pi k(B_q + \tau_l)/N}}{N} \sum_{n=0}^{N-1} \gamma_n e^{-j2\pi kn/N} \sum_{p=0}^{N-1} a_p^{(m)} e^{j2\pi pn/N} \quad (\text{B.2})$$

and $\gamma_n = |x^{(m)}[n]|^2$. After rearranging the terms by consolidating $a_k^{(m)}$,

$$\begin{aligned} \hat{b}_{q,l}^{(m)}[k] &= a_k^{(m)} \left(h_l^{(m)} \alpha_{1,q} e^{-j2\pi k(B_q + \tau_l)/N} + \frac{\alpha_{3,q} e^{-j2\pi k(B_q + \tau_l)/N}}{N} \sum_{n=0}^{N-1} \gamma_n \right) \\ &\quad + \frac{h_l^{(m)} \alpha_{3,q} e^{-j2\pi k(B_q + \tau_l)/N}}{N} \sum_{p=0, p \neq k}^{N-1} a_p^{(m)} \sum_{n=0}^{N-1} \gamma_n e^{j2\pi n(-k+p)/N} \\ &= a_k^{(m)} \mu_{q,l}^{(m)}[k] + \eta_{q,l}^{(m)}[k] \end{aligned}$$

where $\mu_{q,l}^{(m)}[k]$ and $\eta_{q,l}^{(m)}[k]$ are given by (4.14) and (4.15).

B.1.1 Nonlinear noise component is uncorrelated with the desired data symbol

From (4.16)

$$\begin{aligned} \sum_{l=1}^L \eta_l^{(m)}[k] &= \sum_{l=1}^L h_l^{(m)} \sum_{q=0}^Q \frac{\alpha_{3,q} e^{-j2\pi k(B_q + \tau_l)/N}}{N} \left(\sum_{p=0, p \neq k}^{N-1} a_p^{(m)} \sum_{n=0}^{N-1} \gamma_n e^{j2\pi n(-k+p)/N} \right) \\ &= \sum_{p=0, p \neq k}^{N-1} R_p^{(m)}[k] \end{aligned} \quad (\text{B.3})$$

where $R_p^{(m)}[k] = a_p^{(m)} \frac{\Gamma[k]}{N} \sum_{n=0}^{N-1} \gamma_n e^{j2\pi n(-k+p)/N}$ and $\Gamma[k] = \sum_{l=1}^L h_l^{(m)} \sum_{q=0}^Q \alpha_{3,q} e^{-j2\pi(B_q + \tau_l)k/N}$. Thus,

$$\begin{aligned} E_{a_k^{(m)}} \left[a_k^{*(m)} \sum_{p=0, p \neq k}^{N-1} R_p^{(m)}[k] \right] &= \Gamma[k] E_{a_k^{(m)}} \left[a_k^{*(m)} \sum_{p=0, p \neq k}^{N-1} a_p^{(m)} \sum_{n=0}^{N-1} \gamma_n e^{j2\pi n(-k+p)/N} \right] \\ &= \Gamma[k] P^2 \frac{(N-1)}{N^3} \\ &\approx 0 \quad \text{as } N \rightarrow \infty. \end{aligned} \quad (\text{B.4})$$

From (B.4), it can be deduced that the correlation between the nonlinear noise component and the data symbol approaches 0 as $N \rightarrow \infty$.

B.1.2 Modeling of nonlinear noise component as a zero mean Gaussian process

From (B.3)

$$\begin{aligned} E_{a_k^{(m)}} \left[\sum_{l=1}^L \eta_l^{(m)}[k] \right] &= E_{a_k^{(m)}} \left[\sum_{p=0, p \neq k}^{N-1} a_p^{(m)} \sum_{l=1}^L h_l^{(m)} \sum_{q=0}^Q \frac{\alpha_{3,q} e^{-j2\pi k(B_q + \tau_l)/N}}{N} \left(\sum_{n=0}^{N-1} \gamma_n e^{j2\pi n(-k+p)/N} \right) \right] \\ &= \Gamma[k] E_{a_k^{(m)}} \left[\sum_{p=0, p \neq k}^{N-1} a_p^{(m)} \sum_{n=0}^{N-1} \gamma_n e^{j2\pi n(-k+p)/N} \right] \\ &= \Gamma[k] \frac{1}{N^2} E_{a_k^{(m)}} \left[\sum_{p=0, p \neq k}^{N-1} \sum_{n=0}^{N-1} \sum_{u=0}^{N-1} \sum_{v=0}^{N-1} a_p^{(m)} a_u^{(m)} a_v^{*(m)} e^{j2\pi n(u-v-k+p)/N} \right] \end{aligned}$$

$$= 0. \quad (\text{B.5})$$

Thus from (B.5) we have proven that the nonlinear noise component has zero mean. It can also be seen from (B.3) that the nonlinear noise is the sum of $N - 1$ identically distributed random variables. Although $R_p^{(m)}[k], p = 0, 1, \dots, N - 1$ are not uncorrelated, we can approximate the nonlinear noise to be Gaussian by using the Central Limit Theorem. The good agreement between the theoretical and simulation results in Section 5.5 shows that this approximation is acceptable. Thus $\sum_{l=1}^L \eta_l^{(m)}[k] \sim \mathcal{CN}(0, \sigma_{NL}^2[k])$, where $\sigma_{NL}^2[k]$ is the variance of nonlinear noise which is given by

$$\begin{aligned} \sigma_{NL}^2[k] &= E_{a_k^{(m)}} \left[\left| \sum_{l=1}^L \eta_l^{(m)}[k] \right|^2 \right] = E_{a_k^{(m)}} \left[\sum_{p=0, p \neq k}^{N-1} R_p^{(m)}[k] \sum_{p'=0, p' \neq k}^{N-1} R_{p'}^{*(m)}[k] \right] \\ &= (N-1)\sigma_{pp}^2[k] + (N-1)(N-2)C_{ps}[k] \end{aligned} \quad (\text{B.6})$$

where $\sigma_{pp}^2[k]$ and $C_{ps}[k]$ are the variance and covariance given by [31]

$$\begin{aligned} \sigma_{pp}^2[k] &= E_{a_k^{(m)}} [R_p^{(m)}[k] \cdot R_p^{*(m)}[k]] \\ &= \frac{|\Gamma[k]|^2}{N^2} E_{a_k^{(m)}} \left[a_p^{(m)} \sum_{n=0}^{N-1} \gamma_n e^{j2\pi(-k+p)n/N} a_p^{*(m)} \sum_{m=0}^{N-1} \gamma_m e^{-j2\pi(-k+p)m/N} \right] \\ &= \frac{|\Gamma|^2}{N^6} E_{a_k^{(m)}} \left[\sum_{n=0}^{N-1} \sum_{m=0}^{N-1} \sum_{u=0}^{N-1} \sum_{v=0}^{N-1} \sum_{u'=0}^{N-1} \sum_{v'=0}^{N-1} a_p a_p^* a_u a_u^* a_v a_v^* a_{u'} a_{v'}^* e^{j2\pi[p(n-m)+k(m-n)+n(u-v)+m(u'-v')]/N} \right] \\ &= \frac{|\Gamma[k]|^2 (N+2) P^3}{N^4}, \end{aligned} \quad (\text{B.7})$$

$$\begin{aligned} C_{ps}[k] &= E_{a_k^{(m)}} [R_p^{(m)}[k] \cdot R_s^{*(m)}[k]] \\ &= \frac{|\Gamma[k]|^2}{N^2} E_{a_k^{(m)}} \left[a_p^{(m)} \sum_{n=0}^{N-1} \gamma_n e^{j2\pi(-k+p)n/N} a_s^{*(m)} \sum_{m=0}^{N-1} \gamma_m e^{-j2\pi(-k+s)m/N} \right] \\ &= \frac{|\Gamma|^2}{N^6} E_{a_k^{(m)}} \left[\sum_{n=0}^{N-1} \sum_{m=0}^{N-1} \sum_{u=0}^{N-1} \sum_{v=0}^{N-1} \sum_{u'=0}^{N-1} \sum_{v'=0}^{N-1} a_p a_s^* a_u a_u^* a_v a_v^* a_{u'} a_{v'}^* e^{j2\pi[pn-sm+k(m-n)+n(u-v)+m(u'-v')]/N} \right] \end{aligned}$$

$$= \frac{2|\Gamma[k]|^2 P^3}{N^4}. \quad (\text{B.8})$$

Substituting (B.7) and (B.8) in (B.6) we obtain

$$\begin{aligned} \sigma_{NL}^2(k) &= \frac{|\Gamma|^2 P^3 (3N^2 - 5N + 2)}{N^4} \\ &= \frac{\left| \sum_{l=1}^L h_l \sum_{q=0}^Q \alpha_{3,q} e^{-j2\pi(B_q + \tau_l)k/N} \right|^2 P^3 (3N^2 - 5N + 2)}{N^4} \\ &= \frac{\left| \sum_{l=1}^L h_l e^{-j2\pi k \tau_l / N} \right|^2 \left| \sum_{q=0}^Q \alpha_{3,q} e^{-j2\pi B_q k / N} \right|^2 P^3 (3N^2 - 5N + 2)}{N^4} \\ &= \psi \bar{\sigma}_{NL}^2(k) \end{aligned} \quad (\text{B.9})$$

where $\psi = \left| \sum_{l=1}^L h_l e^{-j2\pi k \tau_l / N} \right|^2$ and $\bar{\sigma}_{NL}^2(k) = \frac{\left| \sum_{q=0}^Q \alpha_{3,q} e^{-j2\pi B_q k / N} \right|^2 P^3 (3N^2 - 5N + 2)}{N^4}$.

The above derivation for variance of nonlinear noise component is also applicable for non-fading channel condition i.e. $\psi = 1$, thus for non-fading channel (B.9) reduces to

$$\sigma_{NL}^2(k) = \bar{\sigma}_{NL}^2(k) \quad (\text{B.10})$$

B.1.3 Derivation for complex attenuation component

From (4.16),

$$\begin{aligned} E_{a_k, \forall k} \left[\left| \sum_{l=1}^L \mu_l(k) \right|^2 \right] &= E_{a_k, \forall k} \left[\left| \sum_{l=1}^L h_l e^{-j2\pi k \tau_l / N} \right|^2 \left| \sum_{q=0}^Q \alpha_{1,q} e^{-j2\pi k B_q / N} + \frac{\alpha_{3,q} e^{-j2\pi k B_q / N}}{N} \sum_{n=0}^{N-1} \gamma_n \right|^2 \right] \\ &= E_{a_k, \forall k} \left[\left| \sum_{l=1}^L h_l e^{-j2\pi k \tau_l / N} \right|^2 \left| \sum_{q=0}^Q c_q(k) + d_q(k) \sum_{n=0}^{N-1} \gamma_n \right|^2 \right] \\ &= \psi E_{a_k, \forall k} \left[\left| \sum_{q=0}^Q c_q(k) + d_q(k) \sum_{n=0}^{N-1} \gamma_n \right|^2 \right] \\ &= \psi \bar{\mu}(k) \end{aligned} \quad (\text{B.11})$$

where $\psi = \left| \sum_{l=1}^L h_l e^{-j2\pi k \tau_l / N} \right|^2$, $c_q(k) = \alpha_{1,q} e^{-j2\pi k B_q / N}$, $d_q(k) = \frac{\alpha_{3,q} e^{-j2\pi k B_q / N}}{N}$ and $\bar{\mu}(k) = E_{a_k, \forall k} \left[\left| \sum_{q=0}^Q c_q(k) + d_q(k) \sum_{n=0}^{N-1} \gamma_n \right|^2 \right]$

$\bar{\mu}(k)$ can be calculated as

$$\begin{aligned}
\bar{\mu}(k) &= E_{a_k, \forall k} \left[\left| \sum_{q=0}^Q c_q(k) + d_q(k) \sum_{n=0}^{N-1} \gamma_n \right|^2 \right] \\
&= \left| \sum_{q=0}^Q c_q(k) \right|^2 + \left| \sum_{q=0}^Q d_q(k) \right|^2 E \left[\left(\sum_{n=0}^{N-1} \gamma_n \right)^2 \right] \\
&\quad + \left(\sum_{q=0}^Q c_q(k) \right) \left(\sum_{q=0}^Q d_q(k) \right)^* E \left[\sum_{n=0}^{N-1} \gamma_n \right] \\
&\quad + \left(\sum_{q=0}^Q c_q(k) \right)^* \left(\sum_{q=0}^Q d_q(k) \right) E \left[\sum_{n=0}^{N-1} \gamma_n \right] \\
&= \left| \sum_{q=0}^Q c_q(k) \right|^2 + \left| \sum_{q=0}^Q d_q(k) \right|^2 P^2 + \left(\sum_{q=0}^Q c_q(k) \right) \left(\sum_{q=0}^Q d_q(k) \right)^* P \\
&\quad + \left(\sum_{q=0}^Q c_q(k) \right)^* \left(\sum_{q=0}^Q d_q(k) \right) P.
\end{aligned} \tag{B.12}$$

The above derivation for variance of nonlinear noise component is also applicable for non-fading channel condition i.e. $\psi = 1$, thus for non-fading channel (B.12) reduces to

$$E_{a_k, \forall k} \left[\left| \sum_{l=1}^L \mu_l(k) \right|^2 \right] = \bar{\mu}(k). \tag{B.13}$$

B.2 A Method to Extract Coefficients for a HPA

In this section we describe a method to extract coefficients from an actual HPA. For this purpose we have design a OFDM test-bed based IEEE 802.11a standard. In order to extract the coefficients for a HPA modeled as memory polynomial model, the memory polynomial function is represented by a matrix equation. From the measured input and output data in the time domain we can define a MPM as,

$$y(n) = \sum_{d=1}^D \sum_{q=0}^Q \alpha_{2d-1,q} x(n-q) |x(n-q)|^{2(d-1)} \tag{B.14}$$

Substituting $z_{2d-1,q} = x(n-q)|x(n-q)|^{2(d-1)}$, (B.14) can be rewritten as

$$y(n) = \sum_{d=1}^D \sum_{q=0}^Q z_{2d-1,q} \alpha_{2d-1,q} \quad (\text{B.15})$$

For $D = 2$, i.e restricting the nonlinearity to 3rd order we obtain the above equation in matrix form as

$$\begin{bmatrix} y_0 \\ y_1 \\ \cdot \\ \cdot \\ y_{N-1} \end{bmatrix} = \begin{bmatrix} z_{10}(0) & z_{30}(1) & z_{11}(0) & z_{31}(0) & z_{1Q}(0) & z_{3Q}(0) \\ z_{10}(1) & z_{30}(1) & z_{11}(1) & z_{31}(1) & z_{1Q}(1) & z_{3Q}(1) \\ \cdot & \cdot & \cdot & \cdot & \cdot & \cdot \\ \cdot & \cdot & \cdot & \cdot & \cdot & \cdot \\ z_{10}(N-1) & z_{30}(N-1) & z_{11}(N-1) & z_{31}(N-1) & z_{1Q}(N-1) & z_{3Q}(N-1) \end{bmatrix} \times \begin{bmatrix} \alpha_{10} \\ \alpha_{30} \\ \alpha_{11} \\ \alpha_{31} \\ \alpha_{1Q} \\ \alpha_{3Q} \end{bmatrix} \quad (\text{B.16})$$

Thus the coefficients can easily be obtained by solving (B.16) using the least square estimation technique. Fig. B.1 shows the block diagram of the complete hardware setup for extraction of coefficients and practical validation of our theoretical analysis.

Fig. B.2 shows the implementation error of the above system in comparison with the theoretical values for an AWGN channel¹. Even for a SER of 10^{-6} the implementation error is less than 2dB. So by utilizing the above system, we can extract the coefficients of a nonlinear HPA with memory by using a extraction procedure shown in Fig. B.1 as well as validate our theoretical analysis by comparing the measured SER at the receiver with the theoretically obtained values.

¹Wired connection for the testbed

Figure B.1. Block diagram of the hardware setup to extract the coefficient for a nonlinear HPA

Figure B.2. Implementation error:- Theoretical vs Practical(Test-bed measurements),16QAM

Author's Publications

Journal Papers

1. V. A. Bohara and S. H. Ting, "Analytical performance of orthogonal frequency division multiplexing systems impaired by a non-linear high-power amplifier with memory, *IET Communications*, vol. 3, no. 10, pp. 1659-1666, Oct. 2009.
2. V. A. Bohara, S. H. Ting, Y. Han and A. Pandharipande "An orthogonal spectrum sharing scheme for wireless sensor networks." *EURASIP Journal on Wireless Communications and Networking* 2011 2011:10.
3. V. A. Bohara and S. H. Ting, "Measurement results for cognitive spectrum sharing based on cooperative relaying," *IEEE Transactions on Wireless Communication*, vol. 10, no. 7, pp. 2052-2057, July 2011.

Conference Papers

1. V. A. Bohara, S. H. Ting, Y. Han and A. Pandharipande, "Interference free overlay cognitive radio network based on cooperative space time coding," in *Proceedings of 5th International Conference on cognitive radio oriented wireless networks and communications, CrownCom 2010*, Cannes, France, June 2010.

2. V. A. Bohara and S. H. Ting, "Theoretical analysis of OFDM signals in nonlinear polynomial models," in *Proceedings of 6th ICICS*, Singapore, Dec. 2007.
3. V. A. Bohara and S. H. Ting, "Analysis of OFDM signals in nonlinear high power amplifier with memory," in *Proceedings of International Conference on Communications, ICC 2008*, Beijing, Peoples Republic of China, May 2008.
4. V. A. Bohara and S. H. Ting, "Preliminary measurement results for cognitive spectrum sharing based on cooperative relaying," in *Proceedings of International Conference on Wireless Communications & Signal Processing, WCSP 2010*, Suzhou, China, Oct. 2010.
5. V. A. Bohara, S. H. Ting and Y. Han, "Experimental results for cooperative spectrum sharing," accepted to *Proceedings of IEEE Globecom*, Houston, Texas, USA, Dec. 2011.
6. V. A. Bohara, A. Ramakrishanaiah, M. M. Haq and R. Pattarkine, "Underwater acoustic communication system using orthogonal frequency division multiplexing," in *Proceedings of International conference on Computers, Controls and Communication- INCON-CCC-2004*, Sai Ram Engineering College-Chennai, pp. 474-479, 20-23 Aug. 2004.

Others

1. V. A. Bohara and S. H. Ting, "Design and implementation of overlay cognitive radio network on NI PXIe platform," in *N.I. ASEAN Virtual Instrumentation Applications Contest 2010*, Sept. 2010. (Awarded the best paper in academic segment)
Available: <http://digital.ni.com/worldwide/singapore.nsf/web/all/9E11D52A0EE58F2C862577C0002A28F2>
2. V. A. Bohara, Z. Hongzhi, S. H. Ting, Y. L. Guan and C. L. Law, "Design and implementation of mobile WiMAX (IEEE 802.16e) system based on

2x1 MISO-OFDM configuration,” in *N.I. ASEAN Virtual Instrumentation Applications Contest 2007*, Sept. 2007. (Awarded the best paper in academic segment)

Available: <http://digital.ni.com/worldwide/singapore.nsf/web/all/35076F22AA2D964F862573770040AB18>

Bibliography

- [1] IDA Singapore, "Singapore Spectrum Allocation Chart" Available: <http://www.ida.gov.sg/Policies>
- [2] M.H. Islam, C. L. Koh, S. W. Oh, Q. Xianming, Y. Y. Lai, C. Wang, Y. C. Liang, B. E. Toh, F. Chin, G. L. Tan, W. Toh, "Spectrum survey in Singapore: Occupancy measurements and analyse", in *in Proceedings of 3rd International Conference on Cognitive Radio Oriented Wireless Networks and Communications*, May 2008.
- [3] J. Mitola, "Cognitive Radio: An Integrated Agent Architecture for Software Defined Radio," Ph.D. dissertation, KTH, Stockholm, Sweden, Dec. 2000.
- [4] S. Haykin, "Cognitive radio: Brain-empowered wireless communications," *IEEE Journal on Selected Areas in Communications* vol. 23, no. 2, pp. 201-220, Feb. 2005.
- [5] K. Chowdhury and T. Melodia, "Platforms and testbeds for experimental evaluation of cognitive ad hoc networks," *IEEE Communications Magazine*, vol. 48, no. 9, pp. 96-104, Sept. 2010.
- [6] D. Taubenheim, W. Chiou, N. Correal, P. Gorday, S. Kyperountas, S. Machan, P. Minh, S. Qicai, E. Callaway and R. Rachwalski, "Implementing an experimental cognitive radio system for DySPAN", in *Proceedings of GLOBECOM 07*, pp. 4040 - 4044, 26-30 Nov. 2007.
- [7] P. Amini, E. Azarnasab, P. Amini S. Akoum and B. F. Boroujeny, "An experimental cognitive radio for first responders," in *3rd IEEE Symposium on New Frontiers in Dynamic Spectrum Access Networks, DySPAN 2008*, Chicago, 2007, pp. 1-6.
- [8] L. Huang, Q. Deng, X. Qiao, D. Gu, W. Wang, H. Yang, "SDR implementation of cognitive radio: Sensing and disjoint spectrum access," in *Fourth*

International Conference on Communications and Networking in China, Xian, pp. 1-5, Aug. 2009.

- [9] D. Cabric, S. M. Mishra, and R. W. Brodersen, "Implementation issues in spectrum sensing for cognitive radio," in *Proceedings of Asilomar Conference on Signals, Systems and Computers*, Pacific Grove, CA, 2004, pp. 772-776.
- [10] D. Cabric, A. Tkachenko, and R. W. Brodersen, "Experimental study of spectrum sensing based on energy detection and network cooperation," in *Proceedings ACM 1st Int. Workshop on Technology and Policy for Accessing Spectrum (TAPAS)*, Aug. 2006.
- [11] T. R. Newman, S. M. S. Hasan, D. DePoy, T. Bose and J. H. Reed, "Designing and deploying a building-wide cognitive radio network testbed," *IEEE Communications Magazine*, vol. 48, no. 9, pp. 106-112, Sept. 2010.
- [12] A. Goldsmith, S. A. Jafar, I. Maric, S. Srinivasa, "Breaking spectrum gridlock with cognitive radios: an information theoretic perspective," in *Proceedings of IEEE*, vol. 97, pp. 894-914, May 2009.
- [13] O. Simeone, I. Stanojev, S. Savazzi, Y. Bar-Ness, U. Spagnolini, and R. Pickholtz, "Spectrum leasing to cooperating secondary ad hoc networks," *IEEE Journal on Selected Areas of Communications*, vol. 26, pp. 203-213, Jan. 2008.
- [14] O. Simeone, Y. Bar-Ness, U. Spagnolini, "Stable throughput of cognitive radios with and without relaying capability," *IEEE Transactions on Communications*, vol. 55, pp. 2351-2360, Dec. 2007.
- [15] Q. Zhang, J. Jia, and J. Zhang, "Cooperative relay to improve diversity in cognitive radio networks", *IEEE Communications Magazine*, Vol. 47, Issue 2, Feb. 2009, pp. 111-117.
- [16] Y. Han, A. Pandharipande and S. H. Ting, "Cooperative decode-and-forward relaying for secondary spectrum access," *IEEE Transactions on Wireless Communications*, vol. 8, no. 10, pp. 4945-4950, Oct. 2009.
- [17] Y. Han, S. H. Ting, A. Pandharipande, "Cooperative spectrum sharing with relay selection," in *Proceedings of IEEE International Conference on Communication Systems, ICCS 2010*, Singapore, Nov. 2010.
- [18] Y. Han, A. Pandharipande, S. H. Ting, "Cooperative spectrum sharing via controlled amplify-and-forward relaying," in *Proceedings of IEEE International Symposium on Personal, Indoor and Mobile Radio Communications*,

International Workshop on Wireless Distributed Networks, PIMRC 2008, Cannes, France, Sept. 2008.

- [19] Y. Han, S. H. Ting, A. Pandharipande, "Cooperative spectrum sharing with distributed secondary user selection," in *Proceedings of IEEE International Conference on Communications*, Cape Town, South Africa, May 2010.
- [20] Y. Han, S. H. Ting, A. Pandharipande, "Cooperative spectrum sharing protocol with secondary user selection," *IEEE Transactions on Wireless Communications*, vol. 9, no. 9, pp. 2914-2923, Sept. 2010.
- [21] N. Devroye, P. Mitran, and V. Tarokh, "Achievable rates in cognitive radio channels," *IEEE Transactions on Information Theory*, vol. 52, pp. 1813-1827, May 2006.
- [22] I. Maric, A. Goldsmith, G. Kramer, and S. Shamai (Shitz), "On the capacity of interference channels with a partially cognitive transmitter," in *proceedings of 2007 IEEE International Symposium On Information Theory*, Nice, France, Jun. 2007.
- [23] V. A. Bohara, S. H. Ting, Y. Han and A. Pandharipande, "Interference free overlay cognitive radio network based on cooperative space time coding," in *Proceedings of 5th International Conference on cognitive radio oriented wireless networks and communications, CrownCom 2010*, Cannes, France, Jun. 2010.
- [24] V. A. Bohara, S. H. Ting, Y. Han and A. Pandharipande "An orthogonal spectrum sharing scheme for wireless sensor networks." *EURASIP Journal on Wireless Communications and Networking* 2011 2011:10.
- [25] V. A. Bohara, S. H. Ting, "Preliminary measurement results for cognitive spectrum sharing based on cooperative relaying," in *Proceedings of International Conference on Wireless Communications & Signal Processing, WCSP 2010*, Suzhou , China, Oct. 2010.
- [26] V. A. Bohara, S. H. Ting, "Design and implementation of overlay cognitive radio network on NI PXIe platform," in *N.I. ASEAN Virtual Instrumentation Applications Contest 2010*, Sept. 2010. (Awarded the best paper in academic segment) Available: <http://digital.ni.com/worldwide/singapore.nsf/web/all/9E11D52A0EE58F2C862577C0002A28F2>
- [27] V. A. Bohara, S. H. Ting and Y. Han, "Experimental results for cooperative spectrum sharing," accepted to *Proceedings of IEEE Globecom*, Houston, Texas, USA, Dec. 2011.

- [28] V. A. Bohara, S. H. Ting, "Measurement results for cognitive spectrum sharing based on cooperative relaying," *IEEE Transactions on Wireless Communication*, vol. 10, no. 7, pp. 2052-2057, July 2011.
- [29] Y. Han, S. H. Ting, V. A. Bohara, Q. Li, and A. Pandharipande, "Co-operative spectrum sharing protocols: Theory and measurement results," submitted to *IEEE Wireless Communications Magazine*, Mar. 2011.
- [30] V. A. Bohara and S. H. Ting, "Theoretical analysis of OFDM signals in nonlinear polynomial models," in *Proceedings of 6th ICICS*, Dec. 2007.
- [31] V. A. Bohara and S. H. Ting, "Analysis of OFDM signals in nonlinear high power amplifier with memory," in *Proc. of International Conference on Communications, ICC 2008*, Beijing, Peoples Republic of China, May 2008.
- [32] V. A. Bohara, S. H. Ting, "Analytical performance of orthogonal frequency division multiplexing systems impaired by a non-linear high-power amplifier with memory," *IET Communications*, vol. 3, no. 10, pp. 1659-1666, Oct. 2009.
- [33] V. A. Bohara, Z. Hongzhi, S. H. Ting, Y. L. Guan and C. L. Law, "Design and implementation of mobile WiMAX (IEEE 802.16e) system based on 2x1 MISO-OFDM configuration," in *N.I. ASEAN Virtual Instrumentation Applications Contest 2007*, Sept. 2007. (Awarded the best paper in academic segment) Available: <http://digital.ni.com/worldwide/singapore.nsf/web/all/35076F22AA2D964F862573770040AB18>
- [34] V. A. Bohara, A. Ramakrishanaiah, M. M. Haq and R. Pattarkine, "Under-water acoustic communication system using orthogonal frequency division multiplexing," in *Proceedings of International conference on Computers, Controls and Communication- INCON-CCC-2004, Sai Ram Engineering College-Chennai*, pp. 474-479, 20-23 Aug. 2004.
- [35] D. Tse and P. Viswanath, *Fundamentals of Wireless Communications*. Cambridge University Press, 2005.
- [36] H. Poor and G. Wornell, *Wireless Communications: Signal Processing Perspectives*. Prentice Hall, N.J., 1998.
- [37] T. S. Rappaport, *Wireless Communications: Principles and Practice*. Prentice Hall, 2002.
- [38] Tarokh, V., Jafarkhani, H., and Calderbank, A. R.: "Space-Time block coding for wireless communications: performance results", *IEEE Journal on Selected Areas in Communications*, vol. 17(3), pp. 451-460, Mar. 1999.

- [39] S. M. Alamouti, "A simple transmit diversity technique for wireless communications," *IEEE Journal on Selected Areas in Communications*, vol. 16, no. 8, pp. 1451-1458, Oct. 1998.
- [40] J. N. Laneman, D. N. C. Tse and G. W. Wornell, "Cooperative diversity in wireless networks: efficient protocols and outage behavior," *IEEE Transactions on Information Theory*, vol. 50, pp. 3062-3080, Dec. 2004.
- [41] J. N. Laneman and G. W. Wornell, "Distributed space-time coded protocols for exploiting cooperative diversity in wireless networks," *IEEE Transactions on Information Theory*, vol. 49, no. 10, pp. 2415-2425, Oct. 2003.
- [42] P. Mitran, H. Ochiai, and V. Tarokh, "Space-time diversity enhancements using space-time collaboration," *IEEE Transactions on Information Theory*, vol. 51, no. 6, pp. 2041-2057, Jun. 2005.
- [43] A. Nosratinia, T. E. Hunter, and A. Hedayat, "Cooperative communication in wireless networks," *IEEE Communications Magazine*, vol. 42, no. 10, pp. 74-80, Oct. 2004.
- [44] V. Stankovic, A. Hst-Madsen, and Z. Xiong, "Cooperative diversity for wireless ad hoc networks," *IEEE Signal Processing Magazine*, vol. 23, no. 5, pp. 37-49, Sep. 2006.
- [45] F. H. P. Fitzek and M. D. Katz, Eds., *Cognitive Wireless Networks Concepts, Methodologies and Visions Inspiring the Age of Enlightenment of Wireless Communications*. Netherlands: Springer, 2007.
- [46] L. Berlemann and S. Mangold, *Cognitive Radio and Dynamic Spectrum Access*. New York, United States: Wiley, Jan. 2009.
- [47] I. F. Akyildiz, W.-Y. Lee, M. C. Vuran, and S. Mohanty, "Next generation/dynamic spectrum access/cognitive radio wireless networks: A survey," *Computer Networks*, 2006.
- [48] I. F. Akyildiz, W.-Y. Lee, and K. Chowdhury, "CRAHNS: Cognitive Radio Ad Hoc Networks," *Ad Hoc Network Journal*, vol. 7, no. 5, Jul. 2009.
- [49] N. Devroye and V. Tarokh, "Fundamental limits of cognitive radio networks," in *Cognitive Wireless Networks: Concepts, Methodologies and Vision*, F. Fitzek and M. Katz, Eds. Springer, 2007.
- [50] N. Devroye, "Information theoretical limits on cognitive radio networks," in *Cognitive Radio Communications and Networks: Principles and Practice*, A. Wyglinski, M. Nekovee, and Y.T. Hou, Eds. Elsevier, Dec. 2009.

- [51] W. Wu, S. Vishwanath, and A. Arapostathis, "On the capacity of the interference channel with degraded message sets," *IEEE Transactions on Information Theory*, vol. 53, pp. 4391-4399, Nov. 2007.
- [52] H. Charmchi and M. Nasiri-Kenari, "Achievable rates for two interfering broadcast channels with a cognitive transmitter," in *Proceedings of IEEE International Symposium Information Theory*, Jul. 2008.
- [53] A. Somekh-Baruch, S. Shamai(Shitz), and S. Verdú, "Cognitive interference channels with state information," in *Proceedings of IEEE International Symposium Information Theory*, Jul. 2008.
- [54] S. Sridharan and S. Vishwanath, "On the capacity of a class of MIMO cognitive radios," *IEEE Journal on Selected Topics on Signal Processing*, vol. 2, pp. 103-117, Feb. 2008.
- [55] S. I. Gelfand and M. S. Pinsker, "Coding for channel with random parameters," *Problemy Peredachi Informatsii*, vol. 9, no. 1, pp. 19-31, 1980.
- [56] M. H. M. Costa, "Writing on dirty paper", *IEEE Transactions on Information Theory*, vol. IT-29, pp. 439-441, May 1983.
- [57] T. Yucek and H. Arslan, "A Survey of Spectrum Sensing Algorithms for Cognitive Radio Applications," *IEEE Communications Surveys and Tutorials*, Apr. 2007.
- [58] R. Tandra and A. Sahai, "SNR walls for signal detection", *IEEE Journal on Selected Topics on Signal Processing*, Feb. 2008.
- [59] Q. Zhao and B. M. Sadler, "A survey of dynamic spectrum access," *IEEE Signal Processing Magazine*, pp. 79-89, May 2007.
- [60] M. Gastpar, "On capacity under receive and spatial spectrum-sharing constraints," *IEEE Transactions on Information Theory*, vol. 53, pp. 471-487, Feb. 2007.
- [61] A. Ghasemi and E. S. Sousa, "Capacity of fading channels under spectrum-sharing constraints," in *Proceedings of IEEE International Conference on Communication*, Jun. 2006.
- [62] S. M. Mishra, A. Sahai, and R. W. Brodersen, "Cooperative sensing among cognitive radios", in *Proceedings of IEEE International Conference on Communication*, vol. 4, pp. 1658-1663, Jun. 2006.
- [63] A. Ghasemi and E. S. Sousa, "Collaborative spectrum sensing for opportunistic access in fading environments." in *Proceedings of the IEEE DySPAN 2005*, pp. 131-136, Nov. 2005.

- [64] J. Hillenbrand, T. A. Weiss, and F. Jondral, "Calculation of detection and false alarm probabilities in spectrum pooling systems", *IEEE Communications Letter*, vol. 9, no. 4, pp. 349-351, Apr. 2005.
- [65] E. Peh and Y.-C. Liang, "Optimization for cooperative sensing in cognitive radio networks," in *Proceedings of IEEE Wireless Communication and Networking Conference*, Hong Kong, Mar. 2007, pp. 27-32.
- [66] G. Ganesan and Y. (G.) Li, "Agility improvement through cooperative diversity in cognitive radio," in *Proceedings of IEEE Global Telecomm Conference (Globecom)*, vol. 5, St. Louis, Missouri, USA, Nov./Dec. 2005, pp. 2505-2509.
- [67] G. Ganesan and Y. (G.) Li, "Cooperative spectrum sensing in cognitive radio: Part I: two user networks," *IEEE Transactions on Wireless Communications*, vol. 6, pp. 2204-2213, Jun. 2007.
- [68] G. Ganesan and Y. (G.) Li, "Cooperative spectrum sensing in cognitive radio: Part II: multiuser networks," *IEEE Transactions on Wireless Communications*, vol. 6, pp. 2214-2222, Jun. 2007.
- [69] D. Cabric, A. Tkachenko, and R. Brodersen, "Spectrum sensing measurements of pilot, energy, and collaborative detection," in *Proceedings of IEEE Military Communication Conference*, Washington, D.C., USA, Oct. 2006, pp. 1-7.
- [70] M. Hyty, A. Hekkala, M. Katz, and A. Mmmel, "Spectrum awareness: techniques and challenges for active spectrum sensing," in *Cognitive Wireless Networks : concepts, methodologies and visions inspiring the age of enlightenment of wireless communications* edited by F. Fitzek and M. Katz, Springer 2007.
- [71] T. Cover, "Broadcast channels," *IEEE Transactions on Information Theory* vol. IT-18, pp. 2-14, 1972.
- [72] A. Jovicic and P. Viswanath, "Cognitive radio: An information-theoretic perspective," *IEEE Transactions on Information Theory*, vol. 55, pp. 3945-3958, Sept. 2009.
- [73] N. Devroye, P. Mitran, and V. Tarokh, "Limits on communications in a cognitive radio channel," *IEEE Communications Magazine*, vol. 44, pp. 44-49, Jun. 2006.
- [74] D. Cabric, D. Taubenheim, G. Cafaro, and R. Farrell, "Cognitive radio platforms and testbeds," in *Cognitive Radio Communications and Networks*:

Principles and Practice, A.Wyglinski, M. Nekovee, and Y.T. Hou, Eds. Elsevier, Dec. 2009.

- [75] GNU Radio Project; <http://www.gnuradio.org>
- [76] C. R. A. Gonzalez et al., "Open-Source SCA-Based Core Framework and Rapid Development Tools Enable Software- Defined Radio Education and Research," *IEEE Communications Magazine*, vol. 47, no. 10, Oct. 2009, pp. 48-55.
- [77] Ettus Research LLC; <http://www.ettus.com/>
- [78] M. Ott, I. Seskar, R. Siraccusa, and M. Singh, "ORBIT testbed software architecture: Supporting experiments as a service," in *Proceedings of TRIDENTCOM05*, 2005.
- [79] K.C. Huang, X. Jing, D. Raychaudhuri, "MAC Protocol Adaptation in Cognitive Radio Networks: An Experimental Study", in *Computer Communications and Networks, ICCCN 2009*.
- [80] E. Eide, L. Stoller, T. Stack, J. Freire and J. Lepreau, "Integrated Scientific Workflow Management for the Emulab Network Testbed." In *Proceedings of USENIX technical conference, Boston, MA*, Jun. 2006.
- [81] Artem Tkachenko, Danijela Cabric, R.W. Brodersen, "Cognitive Radio Experiments using Reconfigurable BEE2" in *Asilomar Conference on Signals, Systems, and Computers*, Nov. 2006.
- [82] S. Mishra, D. Cabric, C. Chang, D. Willkom, B. Schewick, A. Wolisz, and R. Brodersen, "A real time cognitive radio testbed for physical and network level experiments," in *Proceedings of IEEE DySPAN2005*, Oct 2005.
- [83] R. Frank R, 'Understanding Smart Sensors'. Artech House, 2000.
- [84] I. F. Akyildiz, W. Su, Y. Sankarasubramaniam, E. Cayirci, "Wireless sensor networks: a survey", *Computer Networks: The International Journal of Computer and Telecommunications Networking*, vol. 38 no. 4, pp. 393-422, 15 Mar. 2002.
- [85] F. L. Lewis, "Wireless sensor networks" In *Smart Environments: Technology, Protocols, and Applications*. D. J. Cook and S. K. Das, Eds. Wiley, 2004.
- [86] A. Kansal and M. Srivastava, "Energy-Harvesting-Aware Power Management", in *Wireless sensor networks*. N. Bulusu and S. Jha, Eds. Artech House, 2005.

- [87] G. Zhou, J. A. Stankovic, S. Son, "Crowded Spectrum in Wireless Sensor Networks," in *Proceedings of IEEE EmNets*, 2006.
- [88] B. Mercier, V. Fodor, et al. "Sensor networks for cognitive radio: Theory and system design" *ICT Mobile Summit*, June 2008.
- [89] R. Thobaben and E. G. Larsson, "Sensor-network-aided cognitive radio: On the optimal receiver for estimate-and-forward protocols applied to the relay channel, in *Proceedings of the Forty-First Asilomar Conference on Signals, Systems and Computers (ACSSC)*, pp. 777-781, Nov. 2007.
- [90] Q. Zhao, L. Tong and A. Swami, "Decentralized cognitive MAC for opportunistic spectrum access in ad hoc networks: A POMDP framework," *IEEE Journal on Selected Areas in Communications*, pp. 224-232, Apr. 2007.
- [91] P. Liu, Z. Tao, S. Narayanan, T. Korakis, and S. S. Panwar, "CoopMAC: A cooperative MAC for wireless LANs," *IEEE Journal on Selected Areas in Communications*, vol. 25, no. 2, pp. 340-354, 2007.
- [92] A. Bletsas, A. Khisti, D. P. Reed, and A. Lippman: "A simple cooperative diversity method based on network path selection," *IEEE Journal on Selected Areas in Communications*, vol. 24, no. 3, pp. 659-672, Mar. 2006.
- [93] S. Valentin, H. S. Lichte, D. Warneke, T. Biermann, R. Funke, and H. Karl, "Mobile cooperative WLANs MAC and transceiver design, prototyping, and field measurements," in *Proceedings of IEEE Vehicular Technology Conference (VTC2008-Fall)*, Sep. 2008.
- [94] S. Valentin, H. S. Lichte, H. Karl, S. Simoens, G. Vivier, J. Vidal, and A. Agustin, "Implementing cooperative wireless networks," in *Cognitive Wireless Networks*, F. H. P. Fitzek and M.D. Katz, Ed. Netherlands: Springer, 2007, pp. 155-178.
- [95] V. Genc, S. Murphy, Y. Yang and J. Murphy, "IEEE 802.16J relay-based wireless access networks: an overview," *IEEE Communications Magazine*, vol. 15, no. 5, pp. 56-63, Oct. 2008.
- [96] Q. Li, X. L. Lin, J. Zhang and W. Roh, "Advancement of MIMO Technology in WiMAX: From IEEE 802.16d/e/j to 802.16m," *IEEE Communications Magazine*, vol. 47, no. 6, Jun. 2009, pp. 100-107.
- [97] Y. Li, "Pilot-symbol-aided channel estimation for OFDM in wireless systems," *IEEE Transactions on Vehicular Technology* vol. 49, no. 4, pp. 1207-1215, Jul. 2000.

- [98] L. Tong, B. M. Sadler, and M. Dong “Pilot-assisted wireless transmissions: General model, design criteria, and signal processing,” *IEEE Signal Processing Magazine*, vol. 21, pp. 12-25, Nov. 2004.
- [99] C. S. Patel and G. L. Stuber, “Channel estimation for amplify and forward relay based cooperation diversity systems,” *IEEE Transactions on Wireless Communications* vol. 6, no. 6, pp. 2346-2356, Jun. 2007.
- [100] M. Dottling, B. Raaf and J. Michel, “Efficient channel quality feedback schemes for adaptive modulation and coding of packet data,” in *Proceedings of IEEE Vehicular Technology Conference*, Los Angeles, Sept. 2004.
- [101] T. Korakis, Z. Tao, S. Makda, B. Gitelman, S. Panwar, “To Serve is to Receive: Implications of Cooperation in a Real Environment”, in *Proceedings of Networking 2007*, Atlanta, Georgia, USA, May 2007.
- [102] P. Liu, Z. Tao, S. Narayanan, T. Korakis, and S. S. Panwar, “CoopMAC: A cooperative MAC for wireless LANs,” *IEEE Journal on Selected Areas in Communications*, vol. 25, no. 2, pp. 340-354, 2007.
- [103] A. Bletsas and A. Lippman, “Implementing cooperative diversity antenna arrays with commodity hardware,” *IEEE Communications Magazine*, vol. 44, no. 12, pp. 33-40, 2006.
- [104] C. R. A. Gonzalez et. al., “ Open-source SCA-based core framework and rapid development tools enable software-defined radio education and research”, *IEEE Communications Magazine*, vol. 47, no. 10, pp. 48-55, 2009.
- [105] National Instruments; <http://www.ni.com/pxi/>
- [106] IEEE 802.11a-1999, “Wireless LAN medium access control (MAC) and Physical layer (PHY) Specifications: High Speed Physical Layer in the 5 GHz band,” 1999.
- [107] A. Sharma et al., “Implementation of a cooperative MAC protocol using a software defined radio platform,” in *Proceedings of IEEE LANMAN*, Cluj-Napoca, Romania, Sept. 2008.
- [108] P. Murphy, A. Sabharwal, and B. Aazhang, “Building a cooperative communications system,” Computing Research Repository, <http://arxiv.org/abs/0707.2998>.
- [109] S. Chaudhari, V. Koivunen, and H. V. Poor, “Autocorrelation-based decentralized sequential detection of OFDM signals in cognitive radios,” *IEEE Transactions on Signal Processing*, vol. 57, pp. 2690-2700, July 2009.

- [110] J. K. Cavers, "Amplifier linearization using a digital predistorter with fast adaptation and low memory requirements," *IEEE Transactions on Vehicular Technology*, vol. 39, pp. 374-382, Nov. 1990.
- [111] C. Liang, J. Jong, W. E. Stark and J. R. East, "Nonlinear amplifier effects in communications systems," *IEEE Transactions on Microwave Theory and Techniques*, vol. 47, no. 8, pp. 1461-1466, Aug. 1999.
- [112] G. T. Zhou and J. S. Kenney, "Predicting spectral regrowth of nonlinear power amplifiers," *IEEE Transactions on Communications*, vol. 50, no. 5, pp. 718-722, May 2002.
- [113] E. Costa, M. Midrio, and S. Pupolin, "Impact of amplifier nonlinearities on OFDM transmission system performance," *IEEE Communications Letters*, vol. 3, pp. 37-39, Feb. 1999.
- [114] A. Chini, W. Yiyan, M. El-Tanany, and S. Mahmoud, "Hardware nonlinearities in digital TV broadcasting using OFDM modulation," *IEEE Transactions on Broadcasting*, vol. 44, pp. 1220, Mar. 1998.
- [115] P. Marshall, "Cognitive Radio as a Mechanism to Manage Front-End Linearity and Dynamic Range", *IEEE Communications Magazine*, vol. 47, No. 3, pp 81-87, Mar. 2009.
- [116] B. Come, R. Nessand, S. Donnay, L. van der Perre, W. Eberle, P. Wambacq, M. Engels, and I. Bolsens, "Impact of front-end nonidealities on bit error rate performance of WLAN-OFDM transceivers," in *Proceedings of IEEE Radio and Wireless Conference 2000*, pp. 91-94.
- [117] T. Jiang, and Y. Wu, "An overview: peak-to-average power ratio reduction techniques for OFDM signals", *IEEE Transactions on Broadcasting*, vol. 54, no. 2, pp. 257-268, Jun. 2008.
- [118] H. Ku and J. S. Kenney, "Behavioral modeling of nonlinear RF power amplifiers considering memory effects," *IEEE Transactions on Microwave Theory and Techniques*, vol. 51, no. 12, pp. 2495-2504, Dec. 2003.
- [119] D. Dardari, V. Tralli and A. Vaccari, "A theoretical characterization of nonlinear distortion effects in OFDM systems," *IEEE Transactions in Communications*, vol. 48, no. 10, pp. 1755-1763, Oct. 2000.
- [120] D. Dardari, V. Tralli, and A. Vaccari, "Analytical evaluation of total degradation for OFDM systems with TWTA or SSPA," CSITE Tech. Rep.003-99.
- [121] P. Banelli and S. Cacopardi, "Theoretical analysis and performance of OFDM signals in nonlinear AWGN channels," *IEEE Transactions in Communications*, vol. 48, no. 3, pp. 430-440, Mar. 2000.

- [122] S. Chen and A. Wyglinski, "Digital communication fundamentals for cognitive radio," in *Cognitive Radio Communications and Networks: Principles and Practice*, A. Wyglinski, M. Nekovee, and Y.T. Hou, Eds. Elsevier, Dec. 2009.
- [123] L. J. Climini, Jr., "Analysis and simulation of digital mobile channel using orthogonal frequency division multiplexing," *IEEE Transactions in Communications*, vol. 33, no. 7, pp. 665-675, Jul. 1985.
- [124] U. Reimers, "Digital video broadcasting," *IEEE Communications Magazine*, vol. 36, no. 10, pp. 104-110, Jun. 1998.
- [125] Athanasios Papoulis, *Probability, Random Variables and Stochastic Processes*. 3rd Edition, McGraw-Hill, 1991.
- [126] M. Schetzen, *The Volterra and Wiener Theories of Nonlinear Systems*. Wiley, 1980.
- [127] J. Li and J. Ilow, "A least-squares Volterra predistorter for compensation of non-linear effects with memory in OFDM transmitters," in *Proceedings of the 3rd Annual Communication Networks and Services Research Conference (CNSR05)* 2005.
- [128] M. Isaksson, D. Wisell and D. Rnnow, "Wide-Band dynamic modeling of power amplifiers using radial-basis function neural networks," *IEEE Transactions on Microwave Theory and Techniques*, vol. 53, no. 11, pp. 3422-3428, Nov. 2005.
- [129] L. Ding, G. T. Zhou, D. R. Morgan, Z. Ma, J. S. Kenney, J. Kim and C. R. Giardina, "A robust digital baseband predistorter constructed using memory polynomials," *IEEE Transactions on Communications*, vol. 52, no. 1, pp. 159-165, Jan. 2004.
- [130] S. Haykin, *Communication Systems*. 4th Edition, Wiley, 2000.
- [131] A. Saleh, "Frequency-independent and frequency- dependent nonlinear models of TWT amplifiers," *IEEE Transactions on Communications*, vol. COM-29, no. 11, pp. 1715-1720, Nov. 1981.
- [132] L. Rugini and P. Banelli, "Joint impact of frequency synchronization errors and intermodulation distortion on the Performance of multicarrier DS-CDMA systems," *EURASIP Journal on Applied Signal Processing*, 2005, pp. 730-742.

- [133] E. Bogenfeld, R. Valentin, K. Metzger and W. Sauer-Greff, "Influence of nonlinear HPA on trellis-coded OFDM for terrestrial broadcasting of digital HDTV," in *Proceedings of IEEE International Conference on Communications*, 1993, pp. 1433-1438.
- [134] A.M.Saleh and , R.A. Valenzuela, "A statistical model for indoor multipath propagation," in *IEEE Journal on Selected Areas in Communications*, vol. 5, pp. 128-137, Feb. 1987.