ENHANCED COGNITIVE RADIO WITH ENERGY HARVESTING AND NON-ORTHOGONAL MULTIPLE ACCESS

Sultangali Arzykulov

A thesis submitted in partial fulfillment of the requirement of Nazarbayev University for the degree of Doctor of Philosophy

April 2019
Declaration

I declare that the research contained in this thesis, unless otherwise formally indicated within the text, is the original work of the author. The thesis has not been previously submitted to this or any other university for a degree, and does not incorporate any material already submitted for a degree.

Signed:

Dated:
Abstract

The increasing demand for wireless applications is making radio spectrum scarce. Meanwhile, studies show that the assigned spectrum is not thoroughly utilized. The cognitive radio (CR) technology is proposed as a feasible key technology to solve issues related to the spectrum scarcity. CR can improve the spectrum utilization by reusing the unused spectrum occupied by licensed users. Introduction of CR networks produces two kinds of interference: interference from the CR network (secondary network) to the primary network (PN) and the interference among secondary users. All unwanted interference should be adequately managed in order not to jeopardize the performance of the PN and at the same time improve the performance of CR systems. Interference alignment (IA) is a promising technique that can efficiently manage interference. One of the aims of this thesis is to mitigate the interference by deploying multiple antennas at both transmitter and receiver sides in order to improve the performance of CR networks.

The rapid growth of data-hungry wireless applications is forcing us to perform energy harvesting (EH) from external power sources for the next-generation of wireless communication systems. Especially, CR networks, where receiver nodes need advanced hardware to process a large amount of data, require higher energy consumption. Thus, another goal of the current thesis is to investigate simultaneous wireless information and power transfer (SWIPT) in CR networks in the presence of intra- and inter-network interference over various channel state conditions.

Firstly, a cooperative CR network is investigated over the general $\alpha-\mu$ fading channel. The contribution of this study is mainly described by the exact closed-form expression for the outage probability (OP) of secondary users, which clearly shows how the outage saturation paradigm appears when the interference level at primary receiver is applied.

Secondly, the proposed cooperative CR is extended by applying multiple-input multiple out (MIMO) antennas and an IA technique to deal with intra- and inter-network interference. The negative effect of interference at both primary and secondary receivers is mitigated by using precoding and interference suppression beamforming matrices. The management of interference at primary receivers allows secondary transmitters to increase the transmit power level. Moreover, the instantaneous capacity performance is assessed for the same CR system by applying two EH methods, i.e., time-switching (TS) and power-splitting (PS). Then, the optimal values of TS and PS portions are determined for different channel state information (CSI) scenarios. In addition, the effect of imperfect CSI on bit error rate and capacity performance is provided.

Finally, we jointly study a cooperative CR and non-orthogonal multiple access (NOMA), where we derive closed-form expressions for the OP of NOMA secondary destination users for detect-and-forward and amplify-and-forward relaying techniques. Furthermore, power allocation factors for different distances of secondary NOMA users are found to satisfy OP fairness for all users. In addition, the proposed CR-NOMA network model is further studied with enabled SWIPT technology.
Acknowledgments

First and above all, I thank the Almighty for giving me the strength and patience to work through all these years.

My greatest appreciations go to my supervisor - Professor Theodoros A. Tsiftsis. Thank him for introducing me into the research realm of wireless communications, the patient guidance, encouragement and advice he has provided throughout my Ph.D. study. He has taught me the methodology to carry out the research and to present the research works as clearly as possible. It was a great privilege and honor to work and study under his guidance.

My heartfelt thanks also go to my supervisor - Associate Professor Behrouz Maham, co-supervisor - Assistant Professor Refik C. Kizilirmak and external supervisor - Assistant Professor Mohamed M. Abdallah who cared about my work so much and responded to my questions and queries so promptly. I feel extremely lucky to have them as supervisors.

I want to express my deep gratitude to Assistant Professor Galymzhan Nauryzbayev for all his help from the beginning of my studies, for generously giving his time to offer me valuable comments toward improving my work and for being a good friend.

I would also like to take this opportunity to thank Professor Mohamed-Slim Louini and Associate Professor Mohammad S. Hashmi - my viva examiners, for their very helpful comments and suggestions.

I am also very grateful to Professor Mohammad S. Obaidat, who has given me valuable support and who had a word of encouragement during my Ph.D. studies.

Some special words of gratitude go to Director of Ph.D. Studies Associate Professor Luis R. Rojas-Solórzano for his suggestions throughout my research and dealing with the bureaucracies of doing a Ph.D.

I would like to thank all my Ph.D. colleagues, with whom I have shared moments of deep anxiety but also of big excitement. Their presence was very important in a process that is often felt as tremendously solitaire. I would also like to thank all my friends for accepting nothing less than excellence from me.

I am extremely grateful to my parents for their love, prayers, caring and sacrifices for educating and preparing me for my future. I am very much thankful to my wife and my sons for their love, understanding and continuing support to complete this research work. Also, I express my thanks to my sisters and brothers for their support and valuable prayers. To you, I owe the deepest and most sincere thanks.

Finally, my thanks go to all the people who have supported me to complete the research work directly or indirectly.

Sultangali Arzykulov
Preface

Mr. Sultangali Arzykulov received the B.Sc. (Hons.) degree in radio engineering, electronics and telecommunications from Kazakh National Technical University named after K. I. Satpayev, Almaty, Kazakhstan, in June 2010. He received M.Sc. degree in communication engineering from the University of Manchester in September 2013. From 2010 to 2011, he worked as a network engineer in JSC “Kazakhtelecom”, and then as a teaching assistant at Nazarbayev University from 2014 to 2018. Since September 2015, he has been pursuing his Ph.D. degree in Science, Engineering and Technology with a research focus on cognitive radio, multi-user MIMO systems, interference mitigation, wireless energy harvesting, non-orthogonal multiple access and millimeter wave communications.

Main Supervisors

The topic of the current thesis was proposed by Prof. Theodoros A. Tsiftsis, who was the main supervisor from September 2015 to March 2018. He left Nazarbayev University in March 2018 and his position as the main supervisor was taken by Dr. Behrouz Maham who was also in the initial team of the Ph.D. supervisory committee. This change was made for typical reasons due to the Ph.D. Program regulations of Nazarbayev University.
Contents

Declaration i
Abstract ii
Acknowledgments iii
Preface iv
List of Figures ix
List of Tables xii
List of Abbreviations xiii
List of Symbols xv

1 Introduction 1
1.1 Modern Wireless Systems .............................. 1
1.2 Motivation ............................................. 1
1.3 Key Contributions ..................................... 3
1.4 List of Publications .................................... 4
1.5 Thesis Organization .................................... 6

2 Literature Review 8
2.1 Radio Spectrum ........................................ 8
2.2 Cognitive Radio Technology ............................ 9
  2.2.1 Cognitive Radio Paradigms ......................... 10
  2.2.2 Multiple Input Multiple Output (MIMO) ............. 12
  2.2.3 Interference in Cognitive Radio Networks .......... 13
2.3 Interference Alignment .................................. 14
2.4 Wireless Energy Harvesting ............................ 15
2.5 Non-Orthogonal Multiple Access (NOMA) ............ 17
  2.5.1 Cognitive Radio NOMA ............................ 19
# 3 Cognitive Relaying Networks

3.1 Introduction ................................................. 21

3.1.1 Cooperative Communication .......................... 21

3.1.2 Cognitive Relaying Networks ......................... 23

3.2 Contribution of the Chapter ............................... 23

3.3 System Model .............................................. 24

3.4 Outage Probability Analysis .............................. 26

3.4.1 Outage Analysis for the Rayleigh Distribution .... 27

3.4.2 The $\alpha-\mu$ Distribution .......................... 28

3.4.2.1 Outage Analysis for the $\alpha-\mu$ Distribution 28

3.4.3 Symbol Error Rate Analysis .......................... 30

3.5 Numerical Results and Discussion ...................... 31

3.6 Chapter Summary .......................................... 33

# 4 Wireless Powered CRN with Interference Alignment

4.1 Introduction .............................................. 34

4.1.1 Wireless Powered CRNs with IA ...................... 34

4.1.2 Practical Implementation ............................. 36

4.1.3 Motivation ............................................. 36

4.2 Contribution of the Chapter ............................... 37

4.3 System Model .............................................. 39

4.3.1 Imperfect CSI ........................................ 41

4.4 Time-Switching Relaying ................................. 42

4.4.1 Capacity for TSR ..................................... 44

4.5 Power-Splitting Relaying ................................. 45

4.5.1 Capacity for PSR ..................................... 47

4.6 Outage Probability Analysis ............................. 48

4.6.1 Outage Probability of PSR ......................... 48

4.6.2 Outage Probability of TSR ......................... 49

4.6.3 Bit-Error Rate ........................................ 50

4.7 Simulation Results and Discussion ...................... 50

4.8 Chapter Summary .......................................... 58

# 5 Cognitive Radio Non-Orthogonal Multiple Access Networks

5.1 Introduction .............................................. 59

5.1.1 Cooperative NOMA .................................. 59

5.2 Contribution of the Chapter ............................. 60

5.3 Organization of the Chapter ............................. 60

5.4 Outage analysis of dual-hop CR-NOMA: Case 1 .... 61
Appendices for Chapter 4

B.1 Proof of Proposition 3 113
B.2 Proof of Proposition 4 113

Appendices for Chapter 5

C.1 Proof of Proposition 5 115
C.2 Proof of Proposition 6 116
C.3 Proof of Proposition 7 117
C.4 Proof of Proposition 8 120
C.5 Proof of Proposition 9 121
C.6 Proof of Proposition 10 123
C.7 Proof of Proposition 11 125
C.8 Proof of Proposition 12 126
C.9 Proof of Proposition 13 127

Bibliography 135
# List of Figures

2.1 The radio spectrum allocations in the USA [1]. .......................... 9  
2.2 Cognitive radio network. .................................................... 10  
2.3 The Cognitive radio cycle. .................................................. 11  
2.4 Cognitive radio paradigms. .................................................. 12  
2.5 The MIMO network with $N_T$ transmit and $N_R$ receive antennas. 13  
2.6 A transceiver with energy harvesting module. .......................... 16  
2.7 An example of a power-domain NOMA for two users. .................. 18  
2.8 An illustration of CR-NOMA [2]. ........................................... 20  

3.1 The basic cooperative relaying schemes: amplify-and-forward and decode-and-forward. ...................................................... 22  
3.2 A system model for cooperative underlay CR network consisting of a secondary source, relay and destination. ............................. 25  
3.3 The OP versus SNR performance for various ITC over the Weibull distribution. .......................................................... 31  
3.4 The OP versus SNR performance for different ITC over Rayleigh and Negative Exponential distributions. ............................. 32  
3.5 The SER versus SNR performance over the Weibull distribution when $I = 20$ dB. ................................................................. 32  

4.1 Time-switching and power-splitting energy harvesting modes. ...... 35  
4.2 The IA- and EH-based CRN with $L$ PUs and one SN sharing the spectrum simultaneously. ......................................................... 38  
4.3 The time frame structure of the TSR. ...................................... 42  
4.4 Time frame structure of the PSR. .......................................... 46  
4.5 Average capacity versus the EH TS and PS coefficients for $D$ in TSR and PSR protocols at 20 dB, respectively. .......................... 51  
4.6 Average capacity versus SNR of $PR_j$ and $D$ operating in the TSR and PSR protocols under different CSI scenarios: a) Perfect CSI versus CSI mismatch ($\kappa = 0$, $\psi = 0.001$). b) CSI mismatch: ($\kappa = 1.5$, $\psi = 15$) versus ($\kappa = 1$, $\psi = 10$). ......................................................... 52
4.7 BER versus SNR for the PSR protocol under different CSI scenarios: a) BER performance for perfect CSI and SNR-independent CSI mismatch cases ((0, 0.001) and (0, 0.005)). b) BER performance for SNR-dependent CSI mismatch cases ((0.75, 10), (1, 10) and (1.5, 15)).

4.8 BER versus CSI mismatch parameters $\kappa$ and $\psi$ of $D$ at 20 dB in the PSR mode.

4.9 The OP of the PN and the SN with different values of data rate threshold at 25 dB with the perfect CSI case: a) The OP of the system in the PSR protocol. b) The OP of the system in the TSR protocol.

4.10 Average harvested power versus distance for the TSR and PSR protocols at 20 dB: a) The TSR protocol. b) The PSR protocol.

5.1 A system model of the downlink dual-hop underlay CR-NOMA network with transmit power constraint at the secondary source.

5.2 The OP of $x_1$ and $x_2$ versus SNR performance when $d_1 = \{1.5d, 2d, 3d, 5d\}$, $\alpha_1 = 0.8$ and $\alpha_2 = 0.2$.

5.3 The OP of $x_1$ and $x_2$ versus the SNR threshold at 20 dB when $d_1 = \{1.5d, 2d, 3d, 5d\}$, $\alpha_1 = 0.8$ and $\alpha_2 = 0.2$.

5.4 The OP of $x_1$ and $x_2$ versus PA factors at 20 dB when $d_1 = \{1.5d, 2d, 3d, 5d\}$.

5.5 The OP of $x_1$ and $x_2$ versus the transmit SNR performance with OF-based PA factors when $d_1 = \{2d, 3d, 5d\}$.

5.6 The OP versus SNR threshold with OF-based PA factors at 20 dB, when $d_1 = \{2d, 3d, 5d\}$.

5.7 The OP versus the transmit SNR performance for NOMA and OMA users with $\alpha_1 = 0.8$, $d_1 = 2d$ and $u = 4.7$ dB.

5.8 The OP of $D_1$ and $D_2$ versus the transmit SNR when $d_1 = \{2d, 4d\}$, $\alpha_1 = 0.8$, $\alpha_2 = 0.2$ and $\tau = 0$.

5.9 The OP of $D_1$ and $D_2$ versus the transmit SNR when $d_1 = 3d$, $\alpha_1 = 0.8$, $\alpha_2 = 0.2$ and $\tau = 0$.

5.10 The OP of $D_1$ and $D_2$ versus the PA factors at 20 dB when $d_1 = \{1.5d, 2d, 4d\}$ and $\tau = 0$.

5.11 The OP of $D_1$ and $D_2$ versus the transmit SNR with OF-based PA factors when $d_1 = \{1.5d, 2d, 4d\}$ and $\tau = 0$.

5.12 The OP of $D_1$ and $D_2$ versus the predefined SNR threshold with OF-based PA factors at 20 dB when $d_1 = \{1.5d, 2d, 4d\}$ and $\tau = 0$.

5.13 A dual-hop underlay dual-hop CR-NOMA network.

5.14 The OP versus the transmit SNR for $x_1$ and $x_2$ when $d_2 = \{2m, 4m, 6m\}$, $a_1 = b_1 = 0.2$ and $a_2 = b_2 = 0.8$.

5.15 The OP versus $a_1$ for $x_1$ and $x_2$ at 30 dB when $d_{SR} = \{d, 2d\}$.
5.16 The optimized OP versus the transmit SNR for $x_1$ and $x_2$ when $d_2 = \{2d, 4d, 6d\}$.

5.17 A downlink underlay dual-hop CR-NOMA network.

5.18 The OP of messages $x_1$ and $x_2$ versus the transmit SNR when $d_2 = \{1.5d, 2d, 3d\}$.

5.19 The OP of messages $x_1$ and $x_2$ versus the SNR threshold at 20 dB when $d_1 = \{1.5d, 2d, 3d\}$.

5.20 The OP of messages $x_1$ and $x_2$ versus PA factors at 20 dB when $d_1 = \{1.5d, 2d, 3d\}$.

5.21 The OP of messages $x_1$ and $x_2$ versus the transmit SNR with OF-based PA factors when $d_1 = \{1.5d, 2d, 3d\}$.

5.22 The OP of messages $x_1$ and $x_2$ versus the SNR threshold with OF-based PA factors at 20 dB when $d_1 = \{1.5d, 2d, 3d\}$.

5.23 The OP of messages $x_1$ and $x_2$ versus the transmit SNR for NOMA and OMA scenarios with $\alpha_1 = 0.8$, $d_1 = 2d$, $\nu = 0$ and $\sigma_{h_i}^2 = 0$.

5.24 The OP of messages $x_1$ and $x_2$ versus the transmit SNR with imperfect CSI and PN’s interference when $\alpha_1 = 0.8$ and $d_1 = 2d$.

5.25 The average throughput versus the transmit SNR for NOMA messages with $\alpha_1 = 0.8$ and $d_1 = 2d$.

6.1 A downlink dual-hop underlay CR-NOMA network.

6.2 The time frame structure of the PSR protocol.

6.3 The OP of $x_1$ and $x_2$ versus the transmit SNR threshold at 25 dB when $d_1 = 3d$, $\alpha_1 = 0.8$ and $\alpha_2 = 0.2$.

6.4 The OP of $x_1$ and $x_2$ versus PA factors at 20 dB when $d_1 = \{1.5d, 3d, 5d\}$.

6.5 The OP of $x_1$ and $x_2$ versus the transmit SNR performance with OF-based PA factors when $d_1 = \{1.5d, 3d, 4d\}$.

6.6 Harvested power versus the $S-R$ distance performance with various $\theta$ values.
**List of Tables**

5.1 OF-based PA factors for different $d_1$ for the DF mode. . . . . . . . . . . . . . . . 69  
5.2 The observed OF-based PA factors for the AF mode. . . . . . . . . . . . . . . . 73  
5.3 The observed OF-based PA factors ($K$ users). . . . . . . . . . . . . . . . 89  
6.1 OF-based PA factors for different $d_1$ with EH. . . . . . . . . . . . . . . . . 103
# List of Abbreviations

<table>
<thead>
<tr>
<th>Abbreviation</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>1G</td>
<td>First Generation</td>
</tr>
<tr>
<td>3GPP</td>
<td>Third Generation Partnership Project</td>
</tr>
<tr>
<td>4G</td>
<td>Fourth Generation</td>
</tr>
<tr>
<td>5G</td>
<td>Fifth Generation</td>
</tr>
<tr>
<td>AF</td>
<td>Amplify-and-Forward</td>
</tr>
<tr>
<td>AWGN</td>
<td>Additive White Gaussian Noise</td>
</tr>
<tr>
<td>BF</td>
<td>Beamforming</td>
</tr>
<tr>
<td>BPSK</td>
<td>Binary Phase Shift Keying</td>
</tr>
<tr>
<td>CDF</td>
<td>Cumulative Distribution Function</td>
</tr>
<tr>
<td>CDMA</td>
<td>Code Division Multiple Access</td>
</tr>
<tr>
<td>CF</td>
<td>Compress-and-Forward</td>
</tr>
<tr>
<td>CR</td>
<td>Cognitive Radio</td>
</tr>
<tr>
<td>CRN</td>
<td>Cognitive Relaying Network</td>
</tr>
<tr>
<td>CSI</td>
<td>Channel State Information</td>
</tr>
<tr>
<td>DF</td>
<td>Decode-and-Forward</td>
</tr>
<tr>
<td>DoF</td>
<td>Degrees of Freedom</td>
</tr>
<tr>
<td>DSA</td>
<td>Dynamic Spectrum Access</td>
</tr>
<tr>
<td>EH</td>
<td>Energy Harvesting</td>
</tr>
<tr>
<td>FCC</td>
<td>Federal Communications Commission</td>
</tr>
<tr>
<td>FDMA</td>
<td>Frequency-Division Multiple Access</td>
</tr>
<tr>
<td>i.i.d.</td>
<td>Independent and Identically Distributed</td>
</tr>
<tr>
<td>i.n.i.d.</td>
<td>Independent But Not Necessarily Identically Distributed</td>
</tr>
<tr>
<td>IA</td>
<td>Interference Alignment</td>
</tr>
<tr>
<td>ID</td>
<td>Information Decoding</td>
</tr>
<tr>
<td>IoT</td>
<td>Internet of Things</td>
</tr>
<tr>
<td>IT</td>
<td>Information Transfer</td>
</tr>
<tr>
<td>ITC</td>
<td>Interference Temperature Constraint</td>
</tr>
<tr>
<td>LDS</td>
<td>Low Density Spreading</td>
</tr>
<tr>
<td>LPMA</td>
<td>Lattice Partition Multiple Access</td>
</tr>
<tr>
<td>LTE</td>
<td>Long Term Evolution</td>
</tr>
<tr>
<td>MA</td>
<td>Multiple Access</td>
</tr>
<tr>
<td>MGF</td>
<td>Moment Generating Function</td>
</tr>
<tr>
<td>Acronym</td>
<td>Description</td>
</tr>
<tr>
<td>---------</td>
<td>-------------</td>
</tr>
<tr>
<td>MIMO</td>
<td>Multiple-Input Multiple-Output</td>
</tr>
<tr>
<td>MPA</td>
<td>Message Passing Algorithm</td>
</tr>
<tr>
<td>MUST</td>
<td>Multiple User Superposition Transmission</td>
</tr>
<tr>
<td>NOMA</td>
<td>Non-Orthogonal Multiple Access</td>
</tr>
<tr>
<td>Ofcom</td>
<td>Office of Communications</td>
</tr>
<tr>
<td>OFDMA</td>
<td>Orthogonal Frequency Division Multiple Access</td>
</tr>
<tr>
<td>OMA</td>
<td>Orthogonal Multiple Access</td>
</tr>
<tr>
<td>OF</td>
<td>Outage Fairness</td>
</tr>
<tr>
<td>OP</td>
<td>Outage Probability</td>
</tr>
<tr>
<td>PA</td>
<td>Power Allocation</td>
</tr>
<tr>
<td>PDF</td>
<td>Probability Density Function</td>
</tr>
<tr>
<td>PDMA</td>
<td>Pattern Division Multiple Access</td>
</tr>
<tr>
<td>PN</td>
<td>Primary Network</td>
</tr>
<tr>
<td>PS</td>
<td>Power-Splitting</td>
</tr>
<tr>
<td>PSR</td>
<td>Power-Splitting Relaying</td>
</tr>
<tr>
<td>PU</td>
<td>Primary User</td>
</tr>
<tr>
<td>QoS</td>
<td>Quality of Service</td>
</tr>
<tr>
<td>RF</td>
<td>Radio Frequency</td>
</tr>
<tr>
<td>RV</td>
<td>Random Variable</td>
</tr>
<tr>
<td>SCMA</td>
<td>Sparse Code Multiple Access</td>
</tr>
<tr>
<td>SER</td>
<td>Symbol Error Rate</td>
</tr>
<tr>
<td>SIC</td>
<td>Successive Interference Cancellation</td>
</tr>
<tr>
<td>SINR</td>
<td>Signal-to-Interference-plus-Noise Ratio</td>
</tr>
<tr>
<td>SN</td>
<td>Secondary Network</td>
</tr>
<tr>
<td>SNR</td>
<td>Signal-to-Noise Ratio</td>
</tr>
<tr>
<td>SR</td>
<td>Selective Relaying</td>
</tr>
<tr>
<td>SU</td>
<td>Secondary User</td>
</tr>
<tr>
<td>SWIPT</td>
<td>Simultaneous Wireless Information and Power Transfer</td>
</tr>
<tr>
<td>TDMA</td>
<td>Time-Division Multiple Access</td>
</tr>
<tr>
<td>TS</td>
<td>Time-Switching</td>
</tr>
<tr>
<td>TSR</td>
<td>Time-Switching Relaying</td>
</tr>
<tr>
<td>WBAN</td>
<td>Wireless Body Area Network</td>
</tr>
<tr>
<td>WEH</td>
<td>Wireless Energy Harvesting</td>
</tr>
<tr>
<td>WiFi</td>
<td>Wireless Fidelity</td>
</tr>
<tr>
<td>WiMax</td>
<td>Worldwide Interoperability for Microwave Access</td>
</tr>
<tr>
<td>WPT</td>
<td>Wireless Power Transfer</td>
</tr>
</tbody>
</table>
# List of Symbols

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>$(\cdot)^{[t]}$</td>
<td>Time slot index</td>
</tr>
<tr>
<td>$(\cdot)^\dagger$</td>
<td>Hermitian transpose of a matrix</td>
</tr>
<tr>
<td>$\mathcal{CN}(a, b)$</td>
<td>Complex normal distribution with $a$ mean and $b$ variance</td>
</tr>
<tr>
<td>$f_X(x)$</td>
<td>Probability density function of random variable $X$</td>
</tr>
<tr>
<td>$F_X(x)$</td>
<td>Cumulative distribution function of random variable $X$</td>
</tr>
<tr>
<td>$\Gamma(\cdot)$</td>
<td>Gamma function</td>
</tr>
<tr>
<td>$\Gamma(a, b)$</td>
<td>Upper incomplete gamma function</td>
</tr>
<tr>
<td>$\gamma_{\text{inc}}(a, b)$</td>
<td>Lower incomplete gamma function</td>
</tr>
<tr>
<td>$\mathbb{E}(\cdot)$</td>
<td>Statistical expectation operator</td>
</tr>
<tr>
<td>$h_{ji}$</td>
<td>Channel gain between nodes $j$ and $i$</td>
</tr>
<tr>
<td>$P_{\cdot}$</td>
<td>Transmit power at the secondary transmitter</td>
</tr>
<tr>
<td>$P_{\cdot,\text{max}}$</td>
<td>Maximum allowed transmit power at the secondary transmitter</td>
</tr>
<tr>
<td>$I$</td>
<td>Interference temperature constraint</td>
</tr>
<tr>
<td>$\sigma^2$</td>
<td>Noise power at the receive node</td>
</tr>
<tr>
<td>$\xi$</td>
<td>Predefined SNR threshold</td>
</tr>
<tr>
<td>$K_1(\cdot)$</td>
<td>Modified Bessel function of the second kind of order 1</td>
</tr>
<tr>
<td>$\mathcal{R}_{\text{th}}$</td>
<td>Predefined rate threshold</td>
</tr>
<tr>
<td>$\min(a, b)$</td>
<td>Function that finds minimum between $a$ and $b$</td>
</tr>
<tr>
<td>$\mathcal{M}_\gamma(\cdot)$</td>
<td>Moment generating function</td>
</tr>
<tr>
<td>$x$</td>
<td>Lower bold case denotes vector</td>
</tr>
<tr>
<td>$X$</td>
<td>Upper bold case denotes matrix</td>
</tr>
<tr>
<td>$\mathbf{H}$</td>
<td>MIMO channel</td>
</tr>
<tr>
<td>$I$</td>
<td>Identity matrix</td>
</tr>
<tr>
<td>$\text{tr}(X)$</td>
<td>Trace of a matrix</td>
</tr>
<tr>
<td>$</td>
<td>x</td>
</tr>
<tr>
<td>$|x|$</td>
<td>Euclidean norm of $x$</td>
</tr>
<tr>
<td>$\mathbb{V}(\cdot)$</td>
<td>Variance operator</td>
</tr>
<tr>
<td>$\mathbb{C}^{N\times M}$</td>
<td>Space of complex $N \times M$ matrices</td>
</tr>
<tr>
<td>$\text{span}(X)$</td>
<td>Vector space spanned by the column vectors</td>
</tr>
<tr>
<td>$\text{eig}(X)$</td>
<td>Eigenvectors of $X$</td>
</tr>
<tr>
<td>$d_{m,n}$</td>
<td>Distance between nodes $m$ and $n$</td>
</tr>
</tbody>
</table>
\[ \tau \quad \text{Path-loss coefficient} \]
\[ \mathbf{V} \quad \text{Precoding BF matrix} \]
\[ \mathbf{U} \quad \text{Interference suppression matrix} \]
\[ f \quad \text{Number of data streams} \]
\[ T_j \quad \text{Primary transmitter } j \]
\[ n \quad \text{AWGN noise} \]
\[ \rho \quad \text{Nominal transmit SNR} \]
\[ \alpha \quad \text{Time fraction of the TS} \]
\[ \theta \quad \text{Power fraction of the PS} \]
\[ Q(x) \quad \text{Gaussian-Q function} \]
\[ \alpha_{(s)} \quad \text{NOMA power allocation factor at the source} \]
\[ \beta_{(r)} \quad \text{NOMA power allocation factor at the relay} \]
\[ D \quad \text{Secondary Destination} \]
\[ G \quad \text{Amplification factor at the relay} \]
\[ PR \quad \text{Primary receiver} \]
\[ S \quad \text{Secondary Source} \]
\[ R \quad \text{Secondary Relay} \]
DEDICATED WITH EXTREME AFFECTION AND GRATITUDE TO

my parents Mr. Ussenbatyr ARZYKULOV and Mrs. Zhantore ARZYKULOVA
my brothers NURZHAN and RAKHYMZHAN
my sisters NURGUL, MARZHAN and YERKEZHAN
my wife DINARA and my sons
   DAMIR and ARSEN
Chapter 1

Introduction

1.1 Modern Wireless Systems

Wireless communication is the technology that has certainly changed the lives of human beings. Nowadays, many applications establish their broadcasting using wireless communication concept. The range of those applications varies from highly commercialized satellite and mobile communication networks to amateur radio communications, from regularly practiced Wi-Fi systems to rarely used deep-space networks, from fully infrastructure-based radio/television networks to non-infrastructure ad-hoc systems. The number of wireless applications will be increased dramatically in the future due to the high demand for the wireless medium.

1.2 Motivation

The radio frequency (RF) spectrum, an irreplaceable carrier underlying the wireless communication system, is conditionally assigned to a particular wireless application for independent use by regulatory authorities, such as the Federal Communications Commission (FCC) in the United States, the Office of Communications (Ofcom) in the UK and Ministry of Information and Communication of the Republic of Kazakhstan. Nowadays, the radio spectrum is becoming increasingly scarce due to the fact that more and more devices become wireless and occupy more RF spectrum. Meantime, studies show that some spectrum bands are left without being fully utilized [1]. The research works on the measurement of spectrum usage showed that up to 85% of the spectrum in some frequency bands below 3 GHz are not utilized [3]. The imbalance between the likely spectrum deficit and possible inadequate spectrum utilization inspired the innovative paradigm shift in access to the spectrum by permitting dynamic spectrum sharing and reuse, which is called as dynamic spectrum access (DSA) [4]. The most prominent candidate technology for the DSA is Cognitive Radio (CR), which is able to sense the environment and adapt its operating parameters dynamically. Hence, CR can coexist with existing systems (primary
Chapter 1. Introduction

systems) autonomously in a non-intrusive fashion [1]. Doing so, CR is a promising technology that can significantly improve the spectrum usage by reusing the spectrum that belongs to primary systems [5].

One of the major impairments in wireless communications is interference, which appears due to the broadcast and superposition nature of wireless medium. In the meaning of CR systems, interference problems are an extremely serious issue due to the following two main aspects. On the one hand, CR adheres to the fundamental premise of not creating any harmful interference to the primary system. On the other hand, the efficiency of CR can be limited by interference coming from either primary or other secondary CR nodes (secondary users). Hence, the problems associated with interference in CR networks deserve a thorough and extensive study, which is one of the main focuses of the thesis.

The eventual fate of mobile communication is expected to be completely different from what we experience today. Ultra top-notch videos and better wide-screen resolutions in mobile devices are forcing us either to look for better sustainable power sources or to perform energy harvesting (EH) from external power sources for the next-generation of wireless communication systems. Especially, CR networks, where receiver nodes need advanced hardware to process a large amount of information data, require higher energy consumption. Thus, another focus of the thesis is to investigate EH in CR networks in the presence of intra- and inter-network interference.

The fast increase in the demand for the mobile Internet has propelled 1000-fold data traffic boost by 2020 for the fifth generation (5G) mobile communication networks. In addition, 5G networks need to maintain massive connectivity of devices to meet the requirement for low-cost devices, low latency, and various service types due to the rapid development of the Internet of Things (IoT). Henceforth, such explosive data traffic delivers even more key challenges to efficiently handle the wireless spectrum. So far, non-orthogonal multiple access (NOMA) has been proposed as a potential candidate technology for 5G [6], which not only efficiently utilize the wireless spectrum as CR but also expected to increase the system throughput and support massive connectivity. Therefore, to further improve the performance of 5G networks, the application of NOMA to CR networks seems to be very crucial. Therefore, one more aim of the thesis is to examine a synergy between these two advanced communication techniques considering embedded EH techniques in the presence of interference.
1.3 Key Contributions

The main contributions of the current thesis are briefly summarized as follows:

• We model a cooperative underlay CR system where a primary receiver applies interference temperature model to secondary transmit nodes. The end-to-end closed-form expressions for the outage performance of secondary users are provided over the generalized $\alpha - \mu$ fading distribution, which includes various other distributions, i.e., Rayleigh, Gamma, Exponential, Nakagami-$m$, etc. The impact of interference temperature on the outage performance and symbol error rate of the secondary network is investigated.

• Further, the proposed cooperative underlay CR is studied by applying multiple antennas and interference managing techniques to deal with intra- and inter-network interference. We apply precoding and interference suppression beamforming matrices to mitigate the negative effect of interference at primary and secondary receivers. The management of interference at primary receivers allows secondary transmitters to increase the transmit power level. Closed-form expressions for the outage performance of primary and secondary networks are determined.

• The performance of CR relaying networks is investigated by applying EH at the secondary relay node. EH methods are studied under imperfect channel conditions. Moreover, the instantaneous capacity performance is assessed for both methods. In addition, some useful insights into how the channel state information quality impacts on capacity performance are also provided.

• The synergy of cooperative underlay CR and NOMA is studied in terms of the outage performance of NOMA users by applying relaying protocols over Rayleigh and Nakagami-$m$ fading channels. We find power allocation factors for NOMA secondary users to guarantee outage fairness among NOMA users. Moreover, considering the effect of the imperfect CSI on the outage performance, we derive a general closed-form solution for the outage probability when the number of secondary NOMA nodes is extended to $K$ users. Finally, EH-enabled CR-NOMA is investigated in terms of the outage and EH performance.
1.4 List of Publications

The work presented in this thesis has led to the following publications:

**Published and Accepted Papers**

**Journal Papers**


**Conference Papers**


Submitted Papers

Journal Papers


Conference Papers


Chapter 1. Introduction


1.5 Thesis Organization

The thesis consists of seven chapters. Chapter 1 is the introduction, which provides an overview of the research, motivations and contributions. The following chapters are organized as the next.

Chapter 2 presents a literature review on the relevant background for the research work presented in this thesis. We first provide an introduction to the concepts of the CR with its main three paradigms. Then, some review on using multiple antennas in CR networks are provided. Moreover, different types of interference involved in CR networks are analyzed, and an interference managing technique named interference alignment is proposed as a novel approach to deal with interference. In the end, we review works where the improvement in the performance of the CR networks by applying wireless energy harvesting and NOMA techniques is shown.

Chapter 3 starts with some introduction of relaying networks in the CR systems. Then, a system model of a dual-hop CR with interference temperature constraints at a primary receiver is presented. The end-to-end outage probability (OP) for the secondary network is derived for Rayleigh and general $\alpha - \mu$ fading channels. Moreover, evaluation of the symbol error rate is derived for the proposed system model. Furthermore, in the numerical part, the correctness of the derived analytical results are validated through extensive Monte Carlo simulations.

In Chapter 4, we investigate a dual-hop decode-and-forward wireless powered cognitive relaying network with interference alignment over Rayleigh fading channels. We assume a secondary relay node to be energy-constrained which needs to harvest energy from both information and interference signals. By applying beamforming matrices to primary and secondary networks, the performance metrics such as the OP, capacity and bit error rate are studied under perfect and imperfect channel state information scenarios for both power-splitting relaying and time-switching relaying protocols. Moreover, the Chapter studies the optimal network performance by calculating the optimal energy harvesting time-switching and power-splitting coefficients. Finally, closed-form expressions for the OP of primary and secondary users are derived which are validated through Monte Carlo simulations.

Chapter 5 focuses on a downlink dual-hop CR-NOMA model with an imposed interference constraint at the primary receiver. The chapter considers several different
Chapter 1. Introduction

scenarios for the proposed system model. First, it proposes a system model with two NOMA secondary destination users, where only one secondary transmit node is limited by maximum transmit power. We derive the OP for the considered system model applying decode-and-forward (DF) and amplify-and-forward (AF) relaying methods over Rayleigh and Nakagami-$m$ fading channels. Then, by considering the effect of imperfect CSI on the system performance, we derive a general closed-form solution for the OP when the number of secondary NOMA users is extended to $K$ users. In addition, the OP for the scenario where both source and relay are restricted by the maximum transmit power is studied. Furthermore, in order to obtain fairness between NOMA users, the power allocation factors are optimized depending on the distance between secondary users. Finally, the obtained OP performance for cooperative NOMA compared to that for conventional cooperative OMA to show the supremacy of the former.

Chapter 6 studies the OP of the dual-hop underlay CR network consisted of a secondary relay node with energy harvesting and two NOMA secondary destination users. Moreover, the optimal power allocation factors are found for different distances of NOMA users in order to satisfy OP fairness for both secondary destination users. Additionally, the proposed CR-NOMA is compared with conventional CR multiple access in terms of the OP.

Finally, Chapter 7 concludes the thesis and discusses related future works.
Chapter 2

Literature Review

2.1 Radio Spectrum

In recent years, wireless communication networks have obtained significant popularity due to its convenient untethered connectivity and mobile access. However, the radio spectrum in wireless communication networks is a naturally limited resource. Traditionally, spectrum regulators adopt the fixed spectrum access technique in order to support various wireless services and applications, where each spectrum band is assigned to one or several dedicated users. As a result of doing this, only the delegated, a.k.a. licensed, users get access to communicate over the allotted spectrum band, while other unlicensed users do not have rights to use that spectrum, even if the spectrum is not used by licensed users. The recent fast-growing demand on wireless communications systems has resulted in the fact that most of the available spectrum bands have fully been occupied, which leads to the spectrum scarcity issue. Moreover, the increase of the different wireless services, i.e., voice, web, multimedia, etc., has triggered the overcrowding of spectrum usage, which has consequently brought forth the low quality of service (QoS) for wireless network users [5]. Fig. 2.1 shows the radio spectrum allocation in the USA, where it illustrates that the radio spectrum has been entirely reserved by different wireless applications. From the figure, one can judge that the spectrum is too limited to support additional wireless applications. However, due to the high demand for wireless communications, the spectrum needs to be further utilized by newly emerging wireless networks. For example, over the last three decades, the need for extra bandwidth (radio spectrum) of cellular communication systems has increased exponentially while evolving from the voice-oriented first generation (1G) to the multimedia-rich fourth generation (4G). Despite the widely observed spectrum deficiency, spectrum measures reveal a surprising fact on the spectrum utilization. Thus, the recent studies have affirmed that some licensed spectrum bands, namely, amateur radio, television bands, etc., are mostly not been utilized leaving large spectral holes [7, 8] due to the fixed spectrum allocation and the independent usage of the spectrum band. Hence, more dynamic and flexible spectrum utilization methods are required to avoid the scarce problem in the wireless networks.
2.2 Cognitive Radio Technology

Recently, the FCC has offered CR networks as a feasible key technology to solve the issue related to spectrum scarcity. The concept of CR was first proposed in 1999 by Joseph Mitola III [9, 10]. Cognition implies that the radio learns from its environment and modifies its operating parameters based on the gained information. Hence, the CR concept lies in allowing permission to unlicensed users, named as secondary users (SUs), to broadcast in a licensed spectrum only if licensed users, named primary users (PUs), are not affected by the harmful interference coming from SUs [5]. In point of fact, there are specific policies for unlicensed users to opportunistically utilize the unoccupied licensed spectrum bands developed by the FCC [11]. Fig. 2.2 illustrates an example of the CR network, where PUs coexist with SUs. A secondary network base station coordinates SUs to be in the keep-out radius of the PUs. Thus, the SUs are urged not to use the licensed channel to keep their interference caused to PUs at a tolerable level [4].

CR quickly gained importance in research areas, as it provides more intelligent and optimized networks. The CR cycle is presented in Fig. 2.3, where it summarizes the main stages of the CR process. The cognitive cycle provides intellectual adaptation through learning and information exchange between various objects in the network. The CR cycle consists of six basic elements:

- **Environment**: It involves any object in the surrounding environment that might
change the network condition, e.g., physical channels, neighbor users, obstacles, different climate conditions, etc.

- **Sense:** Various CR objects are intelligent to sense and monitor the spectrum bands, physically distributed channel parameters, levels of interference, and position of SUs and PUs.

- **Plan:** CR first makes a plan and assesses this plan before decisions are made.

- **Decide:** Decision is knowledge- and learning-based, where the decision-making process optimizes the resources of the system.

- **Act:** Depending on the decisions made, the SUs will act on the environment-based, which can be within a series of operations, e.g., the media access, packets routing, resource allocations and transmission scheme modifications.

- **Learn:** Learning is considered to be a prime part in the CR cycle. During this process, CR devices monitor all information related to network conditions and the environment using an obtained knowledge and a learning tool. Doing so, the system can learn from current and prior operations, expect future behavior and use all the obtained knowledge to intelligently plan and make a decision.

### 2.2.1 Cognitive Radio Paradigms

There are main three CR paradigms such as interweave, underlay, and overlay [5, 12]. The interweave is the simplest paradigm, however, it was originally what motivated the initiation of CR. In the interweave CR (See Fig. 2.4a), the SUs can opportunistically
access free primary network (PN) channels by detecting spectrum holes (unoccupied frequency bands). Hence, the SUs do not have to co-exist with the PUs. Not similar to the interweave, in underlay CR as shown in Fig. 2.4b, the SUs can simultaneously start their broadcasting along with the PUs by conditioning that the SUs should not cause harmful interference to primary receiver nodes [13]. Therefore, the SUs will be able to broadcast across PN frequency bands with restricted maximum transmit power in order to not exceed the primary receivers’ noise floor. Hence, the underlay approach does not need to sense the spectrum to get access to transmission, which is the main benefit of this paradigm. Yet, the SUs are required to reduce their transmit power in order to not disturb the PN, which can force the SN to use short-distance communications. The overlay CR paradigm, similar to the underlay approach, can co-exist with the PUs. But, the SUs need to assist to the PUs and should have entire information about signals of PUs (i.e., the message and codebook). Thus, the SUs can co-exist with the PUs by not being restricted with the maximum transmit power level. As it is assumed that CR has information about the message and the codebook of the PUs, the SUs can use this knowledge to eliminate the interference from the PUs by applying methods such as dirty paper coding. In addition, the SUs assist to transfer the primary transmitter messages to the dedicated PN destinations, which helps to compensate for the interference caused to the PUs [5]. The main disadvantage of the overlay approach is that the SUs need to have full knowledge of the PU signal and to obtain that may be impractical since it is necessary to use the considerable coordination between multiple PUs and SUs.
2.2.2 Multiple Input Multiple Output (MIMO)

Wireless channel is a complex and time-varying medium, which causes challenges to propagated signals. The signal transmitted through the wireless channel reaches the receiving node through a number of different paths (multipath). The multipath is the result of scattering and reflection of radiated energy from objects, e.g., trees, hills, buildings, etc. Each path has a different signal amplitude, an angle of arrival and a time-varying path delay. As a result, multipath propagation can cause destructive interference and deteriorate the wireless channel link quality, which leads to low level received signal amplitude. In digital communication systems, such as the wireless Internet, this can lead to a decrease in the data transfer rate and an increase in the number of errors. As a solution for this issue, the pioneering works in [14] and [15] showed that deploying multiple antennas together with the transmission of several signals at the source and destination can eliminate the issue induced by the multipath propagation, and can even take advantage of this effect. Hence, by employing multiple-input multiple-output (MIMO) antennas, the spectral efficiency in wireless communication networks can be increased linearly with the number of antennas. Thus, MIMO is recognized as an important technical invention, which can serve as a promising solution for the ever-increasing demand for high data rates in future wireless communications.

MIMO system provides the main two performance gains such as the spatial diversity gain and the spatial multiplexing gain [16]. Broadcasting and/or receiving replicated data symbols over various antennas cause space diversity, which helps to reduce the possibility of a transmission failure due to the fact that some symbol replica may experience deep fades, whereas others may not. Moreover, it has been admittedly proven that the ergodic channel capacity of a spatial multiplexing MIMO system increases linearly with the number of independent parallel spatial channels supported by the MIMO system, which is known as the degrees of freedom (DoF) [15]. Given its potential, MIMO has been adopted in next-generation WiFi, WiMax, Long Term Evolution (LTE), 4G and other cellular network standards.
Multi-antennas can be utilized to achieve many desirable functions for wireless transmissions, such as folded capacity increase without bandwidth expansion [15, 17, 18], a dramatic enhancement of transmission reliability via space-time coding [19], and effective co-channel interference suppression for multiuser transmissions [20]. All these benefits of MIMO techniques can be also applied to CR networks to further improve the performance of PNs and SNs [21]. Generally speaking, multi-antennas can be used to allocate transmit dimensions in space and hence provide the secondary transmitter in a CR network more DoF in space in addition to time and frequency so as to balance between maximizing its own transmit rate and minimizing the interference powers at the primary receivers.

### 2.2.3 Interference in Cognitive Radio Networks

Interference connected to CR systems can be divided into mainly two types, namely, intra-network interference and inter-network interference. When users within one network (either the PN or SN) cause self-interference to each other, it is considered as intra-network interference. Two types of intra-network interference are considered in CR systems: interference within PNs and SNs. However, when the mutual interference between the primary and secondary networks exists, it is considered as inter-network interference. The management of inter-network interference is more difficult as CR transmitters must strictly control their transmitted signals to assure that the PN’s QoS is guaranteed. In addition to that, CR receivers need to efficiently manage the interference coming from the PN, thus, providing a valuable QoS for the receivers of the SN.

The interference caused by the SN to the PN is very crucial and needs to be at a tolerable level due to the CR principle. As a solution to this, a new interference metric has been proposed named interference temperature model [22]. Traditional interference controlling methods use mainly transmitter-centric approach, while the interference temperature model applies interference temperature limits and practices receiver-centric approach to handling interference at primary receivers. The interference temperature limit indicates
the maximum amount of tolerable interference at primary receivers. Therefore, if that interference limit is acceptable, SN users can get access to utilize the primary spectrum.

By introducing MIMO in CR networks, interference in the network becomes denser due to several replicas of the same transmitted signal, which makes the issue of interference in CR networks more challenging.

## 2.3 Interference Alignment

Traditionally, interference is handled by methods where each user is provided with a certain portion of wireless resources, i.e., time-division multiple access (TDMA) and frequency-division multiple access (FDMA) [23]. These traditional approaches use interference managing techniques, i.e., orthogonalization, which characterizes interference as a background noise and interference decoding [24], which mitigates unwanted interference.

For instance, in TDMA, each user utilizes the whole available frequency band in certain allocated time slots. In the case of FDMA, each user communicates over various portions of a frequency band depending on the users’ priority. These methods allow users to utilize part of wireless resources without interference which makes these techniques achieve lower capacity performance compared to that of interference networks. All of these traditional methods are rather suboptimal for general multi-source networks. Recently, a new interference managing technique named interference alignment (IA) has been proposed by Cadambe and Jafar in [25]. IA is a linear precoding technique that attempts to align interfering signals in time, frequency, or space. The key idea is that users coordinate their transmissions, using linear precoding, such that the interference signal lies in a reduced dimensional subspace at each receiver. In MIMO networks, IA uses the spatial dimension offered by multiple antennas for alignment. Hence, IA is considered to be an emerging technique in the interference management, the basic concept of which is to remove interference at a receiver’s end in order to achieve effective capacity at the high SNR [26, 27]. Moreover, IA technique can design the transmit policies such that the interference can be aligned at all receivers. For instance, for N user pairs, the IA technique achieves a sum throughput on the order of $\frac{N}{2}$ interference free links [23, 28]. In other words, each user in the network can effectively get half the system capacity.

Mainly three types of IA have been studied in the literature, which can be attained in frequency, time or space dimensions [29, 30]. The most popular among them is the space dimensional IA technique, which applicable with MIMO antennas [31]. The fundamental idea of MIMO IA lies on eliminating all interference in the detection through the proper allocation of both the signal and interference into orthogonal subspaces. In the IA network, each node should be equipped with a sufficient number of antennas in order to
perfectly remove all interference [32]. Hence, the required number of antennas at every node to make the issue of IA solvable is calculated by a feasibility condition [33]. IA has found its wide use in different types of wireless network topologies by being an excellent technique in eliminating interference [29]. The idea of IA has been successfully applied to different network models, i.e., MIMO interference channels [33], cellular downlink and uplink [34, 35], X-networks [25], two-way communication networks [36], blind alignment [37], index coding networks [38] and multicast / compound networks [39], etc. The survey of IA can be found in [40–43]. The fundamental aspects of IA were given in [40], while the practical challenges in applying IA and potential future directions were discussed in [23]. The current research works on IA was studied in [41] with an antenna selection and multiuser diversity techniques. Furthermore, [42] reviewed IA in the interference channel with the features of DoF and imperfect channel state information (CSI). Moreover, a more extensive survey of IA research works was presented in [43], where the authors discussed some research problems and open challenges. As it is discussed above, IA is applicable to various wireless networks, namely cellular networks, device-to-device networks, CR network, etc. From here, IA can be said to be a comprehensive technique which can be applied in many multi-user networks to manage interference.

In Chapter 3, interference temperature model to investigate CR system performance over various wireless fading channels is studied, while, in Chapter 4, we will evaluate the performance of interference channels and study how to efficiently manage intra- and inter-network interference by applying IA methods in order to guarantee the QoS for both PNs and SNs.

### 2.4 Wireless Energy Harvesting

The future mobile communication will be totally different from its current version. Nowadays demand for high-quality videos and better screen resolutions from mobile users forces researchers to seek out better sustainable power sources or to execute wireless EH in the upcoming wireless networks. Recently, to deal with this issue, a simultaneous wireless information and power transfer (SWIPT) technique has been developed [44], where the same RF signal is used to deliver both energy and intended information. In the upcoming mobile generations, the SWIPT technology can be of primary importance with its energy and information transmission that can be adopted in various modern communications systems [45, 46]. The main three benefits of the SWIPT technology can be described by follows. First of all, the lifetime of SWIPT adopted wireless nodes can be prolonged by harvesting energy from ambient signals. Second, the transmission efficiency of the SWIPT method is better compare with that of the conventional time-division multiplexing mechanism due to simultaneous power and information transmission in the prior
method. The last gain is the fact that, in SWIPT, the interference is controlled and even can be used for EH purposes [41]. Fig. 2.6 illustrates a block diagram of a transceiver with an energy harvesting module. The wireless power receiver part consists of a receiving antenna, a matching network, an RF-DC rectifier, a power management unit and an energy storage unit (i.e., a rechargeable battery [47]). Then, after the energy storage unit is successfully charged, the power is provided to the central processing unit or transceiver to process information.

There are some review articles on EH and SWIPT [44, 45, 48–56]. A well-organized overview of SWIPT technology has been provided by the authors in [48], where recent advances and future challenges of the technology were presented. In [45], a survey on current studies in RF-EH networks was presented, which covered circuit designs, communication protocols and emerging operation designs for various EH systems. Furthermore, the authors in [50] have discussed smart antenna technologies applied in SWIPT adopted MIMO networks with the basics of SWIPT and receiver architectures. The above-mentioned survey works have contributed important insights into the foundational features of EH and SWIPT technologies as well as provided some recommendations. A review based on SWIPT from both the technical and theoretical perspectives was presented by the authors in [51]. The work has started with surveying of various kinds of SWIPT scenarios with their challenging problems. Furthermore, by presenting performance improvement through derived results, the authors suggested using massive MIMO
technology in order to obtain better performance in the SWIPT technique. The recent advances of EH and wireless power transfer (WPT) in terms of signal processing, information/communication theory and wireless networking were summarized in [52]. Moreover, a comprehensive survey of EH in full-duplex CR systems in perspective of architectures and related case studies were presented in [53]. The authors in [54] have discussed existing solutions for implementing WPT such as a static charger scheduling, a wireless charger deployment and a mobile charger dispatch. In addition, the authors in [54] have reviewed open problems and practical challenges in adopting wireless charging. The survey work on exploiting interference for wireless EH systems was introduced in [55], where the authors presented a classification of wireless EH systems which utilize interference. The work introduced current advances on an antenna dimension, a receiver architecture, a network topology and an interference management. Fundamental aspects of the energy scheduling and properties of various power sources in EH systems were presented in [57], while, the authors in [56] discussed fundamental limits of EH systems and it was concluded that, in EH networks, each transmitted symbol is instantaneously constrained by the energy available in the battery. Finally, works in [44] and [58] have illustrated overview based on the policies of EH and WPT in wireless sensor networks. The author in [44] concluded that RF EH is mostly suitable for small-sized battery-powered sensors which can be used in wireless body area networks (WBANs), future smart buildings, etc.

In Chapter 4, we study the EH performance of CR networks with MIMO antennas in the presence of interference. Saying so, we demonstrate how interference can be used as an EH source instead of just being canceled.

2.5 Non-Orthogonal Multiple Access (NOMA)

NOMA is another spectrum scarcity resolving technology, which has recently received much attention in wireless communications as being a promising candidate for upcoming 5G networks [59]. Each user in NOMA can utilize the whole available resources such as time or frequency, which results in better spectrum efficiency [60, 61]. There are various multiple access (MA) techniques for 5G networks have been introduced by academia and industry, e.g., sparse code multiple access (SCMA) [62, 63], low density spreading (LDS) [64], pattern division multiple access (PDMA) [65, 66], lattice partition multiple access (LPMA) [67] and power-domain NOMA [68–71]. All above-mentioned techniques adopt the same fundamental idea, where more than one user use all available orthogonal resource block such as a frequency channel, a time slot, a spreading code or an orthogonal spatial DoF, etc. In traditional orthogonal multiple access (OMA) methods, e.g., orthogonal frequency division multiple access (OFDMA) and TDMA, each user is served in an individual orthogonal resource block. Those OMA techniques are inefficient
as, despite some users’ poor channel conditions, each user is solely allocated with one of the scarce bandwidth resources for fairness purposes. Certainly, inefficient use of resources for poorer users negatively impacts on the spectrum efficiency and throughput of the overall system. On the other hand, the NOMA technique guarantees that users with weak and strong channel conditions can get simultaneous access to the whole available resources, i.e., time, frequency or code. As a consequence, by guarantying the users’ fairness, the system throughput of NOMA can be significantly larger than that of OMA [72].

Academic and industrial research works have proved that the NOMA technique can obtain better spectral efficiency gain compared to conventional OMA techniques. Apart from this, it has demonstrated that NOMA can effectively support massive connectivity, which is a vital criterion for the IoT [73, 74].

Despite the fact that the application of NOMA in wireless networks is novel, related concepts have been considered in information theory for a long period. Saying so, fundamental components of NOMA, i.e., successive interference cancellation (SIC), superposition coding and the message passing algorithm (MPA), have already been introduced at the beginning of this century [75, 76]. However, removing orthogonality which is the principle idea of NOMA, had not been practiced in the prior generations of cellular networks. In this context, it is noted that the conception of NOMA is quite different from that of code division multiple access (CDMA). Indeed, CDMA is basically based on the idea that users are separated using the differences between their spreading codes, while NOMA serves several users with the same code. Therefore, in CDMA, the chip rate should be much higher than the supported data rate. For example, to obtain the data rate of 10 Gbps, the requirement for the chip rate can be several hundred Gbps, which is tough to execute with practical equipment.

NOMA is compatible with existing and future wireless systems. Therefore, NOMA can be integrated with traditional OMA techniques such as OFDMA and TDMA [77]. Due to this ability, downlink NOMA has been adopted in the 3GPP LTE which is called multiple user superposition transmission (MUST) [78]. Especially, an application of the NOMA principle assures that two users can be served at the same time within the same
OFDMA subcarrier. Thus, the LTE resource blocks remain unchanged.

The most popular type of NOMA is the power-domain NOMA [69], where transmitters broadcast a superimposed signal to all receivers by serving them simultaneously with different power levels depending on their QoS or channel conditions [68], i.e., users with better channel conditions are allocated with less transmit power, while users with worse channel conditions are assigned with higher transmit power to obtain user fairness (Fig. 2.7). Then, receiver nodes with better channel qualities apply the SIC to decode messages of users with weak channel qualities before decoding their own ones\textsuperscript{1} [71]. The performance of NOMA in a cellular downlink scenario with randomly deployed users was investigated in [69], where it was concluded that NOMA performs better than a conventional OMA in terms of the ergodic sum rate. However, it was also deduced that the OP of NOMA crucially depends on the appropriate choice of the power allocation (PA) factors. Authors in [68] studied the OP of NOMA with randomly distributed users as well, but with partial CSI. The analytical results there also showed the outperformance of NOMA. Moreover, the NOMA scheme was studied from a fairness standpoint in [79], where PA techniques were investigated to ensure fairness for downlink users.

2.5.1 Cognitive Radio NOMA

Recently, an advance of NOMA technique has also been applied into CR systems. The synergy of these promising technologies can significantly decrease the complexity of the power allocation design and surely support the QoS requirements of users [80–84]. For instance, the authors in [85] applied the NOMA policy to large-scale underlay CR networks to improve the connectivity of SNs. Not similar to conventional NOMA wireless networks, in CR-NOMA systems, the power of the superimposed signals need to be constrained in order to not exceed interference level at the PN. Related work has been studied in [86], where the secondary NOMA transmitters supported two functions: the first is to deliver information to SN’s receivers, the second is to act as a relay helping the PN’s receivers. Moreover, underlay CR-NOMA networks were studied in [85], where a secondary source transmits superimposed NOMA messages to the SUs. The results revealed that careful design of the target data rates and PA factors enhances the performance of NOMA users compared to OMA ones. The advantage of cooperative communications has also been applied in CR-NOMA networks [87], [88]. For instance, the authors in [87] studied the application of NOMA to achieve the superposition transmission for an unicast PU and multicast SUs, where the SN provides a cooperation between PUs in order to compensate the access to the primary spectrum, whereas, the same system model was studied in [84], where the direct connection along with the relay channel is considered.

\textsuperscript{1}For the sake of brevity, hereafter, the appearance of NOMA in the text assumes power-domain NOMA.
Figure 2.8. An illustration of CR-NOMA [2].

The authors in [88] proposed a novel cooperative multicast CR-NOMA scheme, where a two-stage cooperative strategy was devised to enhance fairness of SUs.
Chapter 3

Cognitive Relaying Networks

In this Chapter, a dual-hop detect-and-forward cognitive relaying system with interference temperature constraints over the general $\alpha - \mu$ fading channel is studied. The contribution of this Chapter is mainly described by the exact closed-form expression for the OP. Moreover, the symbol error rate calculation is derived for the proposed system model. Furthermore, the correctness of the derived analytical results is validated through Monte-Carlo simulations which show how the outage saturation paradigm appears when the interference level at the primary receiver is constrained. Finally, it is revealed that the system performs better when higher interference constraint is applied.

3.1 Introduction

3.1.1 Cooperative Communication

The large demand for mobile access and advances in wireless hardware technology have caused tremendous progress in wireless communication networks over the past decades. In previous years, many attempts to attain reliable and high data-rate communication over the wireless channels have been failed because of different wireless channel characteristics such as shadowing, multi-path fading, and path loss effects. Due to these effects, the wireless channel quality varies in space, time, and frequency. Therefore, techniques used in traditional wireline communications are not always suitable for wireless channels. One of the techniques to overcome such an issue is a transmit/receive diversity technique, where a diversity gain can be achieved in space, time or frequency domains. With a good knowledge of CSI, more resources such as power, time, and frequency can be allocated to users with stronger CSI and users with unreliable channels are allocated with fewer resources. Yet while CSI is not available, one can implement space-frequency or space-time resource allocation policies to improve the reliability of the network. Particularly, recent advances in MIMO antennas benefit in achieving spatial diversity gains in modern wireless networks [17, 89–92]. On the other hand, the cost and size of many wireless network applications, such as mobile phones, sensors, etc., are limited and installation of
multiple antennas on such devices can be impractical. In such a circumstance, a desirable and promising alternative technique can be when single antenna mobile users can cooperate with each other to build up a distributed antenna system to obtain spatial diversity gains. This method is called cooperative or relaying communications [93].

In cooperative communication systems, users help one another to relay their messages to a destination user [94–100]. On that basis, cooperative users can achieve spatial diversity gains by imitating a distributed antenna array of MIMO systems. Thus, due to wireless channel characteristics, different cooperative users experience various fading effects while relaying each others’ data to the destination user, which consequently improves transmission reliability.

Many relaying techniques have been studied in the literature, namely, AF [101, 102], DF [101, 103], coded cooperation [104], selective relaying (SR) [105], compress-and-forward (CF) [105] and so on. All relaying techniques work in a half-duplex mode, which means that, at each time slot, only one user serves as a source node, whereas other users act as relay nodes to forward messages from the source to the destination. Depending on the system requirement, any cooperative user can act as either a source or a relay. Fig. 3.1 illustrates AF and DF relaying schemes, which are the most basic and widely adopted relaying schemes. In both schemes, in the 1st time slot, the source transmits its data to both the destination and the relay. In the 2nd time slot, while the AF relaying scheme is implemented, the received signal at the relay is amplified and forwarded to the destination. Then, the destination combines the both received signal from the source and the relay to increase the performance of the signal detection. If the DF relaying scheme is adopted, within the 2nd time slot, the relay detects the received signal from the source and forwards a re-encoded message to the destination.
3.1.2 Cognitive Relaying Networks

CR is a promising technique that can utilize the wireless spectrum usage, the main idea of which is to allow SUs to broadcast in licensed frequency bands when no communication activity is required for PUs. The benefits of cooperative communications have also been applied in CR [106–110]. A relay network adopted to CR enhances the throughput and coverage area of the entire network [106]. In cognitive relaying networks (CRNs), the SN can obtain better throughput by implementing the main two approaches. The first one is while the PUs and the SUs cooperate with each other, whereas, in the second approach, the cooperation happens only among the SUs [107]. The authors in [108] compared the performance of licensed and unlicensed bands with that of the conventional relay network. The results showed that adopting cooperation in CR shows outstanding performance in wireless relay networks by reducing inter-cell interference. Another CRN studies [109] has developed the best relay selection technique in an underlay CRN, where obtained results showed that the CRN achieves the same full selection diversity order as traditional networks. Moreover, an increase in the number of relays resulted in a better OP compared to conventional relay networks. Furthermore, an underlay dual-hop DF CRN in Nakagami-\(m\) fading channels was studied in [110], where each relay demodulates the received signal and forwards it to the destination. The authors in [111] derived the OP of underlay CRN which operates in generalized-\(K\) and generalized-\(\gamma\) distributions.

In the wireless medium, a large number of distributions exist that can well represent the statistics of the wireless channel. For example, the long-term signal variation can be described by the Log-normal distribution, while the short-term signal variation can be presented by various other distributions, i.e., Rice, Rayleigh, Hoyt, Weibull, and Nakagami-\(m\). All studies in the literature of CRNs were mainly adopted Rayleigh, Rician and Nakagami-\(m\) fading channels, while neglecting other distributions such as Log-normal, Negative exponential, Weibull, etc. By motivated from the research above, the rest of this Chapter considers a study of CRNs over the general \(\alpha - \mu\) distribution that fully embraces all the aforementioned channel distributions [112].

3.2 Contribution of the Chapter

The main contributions of this Chapter are as follows:

- The outage performance study is provided for a dual-hop underlay CR system, where the primary receiver applies the interference temperature model to secondary transmit nodes.
• Considering imposed transmit power constraint at the SN, the OP is derived in the closed form for the secondary source and relay nodes over Rayleigh fading distribution.

• The closed-form OP expressions further derived for the generalized $\alpha-\mu$ fading distribution, which includes various other distributions, i.e., Rayleigh, Gamma, Exponential, Nakagami-$m$, etc.

• Finally, all derived analytical results are validated through Monte-Carlo simulations. Moreover, the numerical part shows how the level of interference constraint impacts on the end-to-end outage and symbol error rate performance of the secondary destination node.

The remainder of this Chapter is organized as follows. The proposed system model for the underlay CRN is described in Section 3.3. The OP analysis for different fading distributions is presented in section 3.4, while numerical results are illustrated in Section 3.5. Finally, the Chapter is summarized in Section 3.6.

3.3 System Model

The underlay CRN model is considered as shown in Fig. 3.2, where $PR$, $S$, $R$ and $D$ represent the primary receiver, the secondary source, the secondary relay and the secondary destination, accordingly. Moreover, the corresponding channel gains between nodes are denoted by $h_{SP}$, $h_{RP}$, $h_{SR}$ and $h_{RD}$. It is assumed that there is no direct link between $S$ and $D$. Thus, we assume short-distance communication where $R$ applies a DF relaying scheme to forward signal from $S$ to $D$. Regarding the principle of the underlay CR paradigm, the secondary nodes can broadcast only if they do not cause harmful interference to $PR$. Hence, source and relay transmit power constraints are accordingly written by

$$P_S \leq \min\left( \frac{I}{|h_{SP}|^2} , \bar{P}_S \right),$$  \hspace{1cm} (3.1)

$$P_R \leq \min\left( \frac{I}{|h_{RP}|^2} , \bar{P}_R \right),$$  \hspace{1cm} (3.2)

where $\bar{P}_S$ and $\bar{P}_R$ are the maximum source and relay transmit power, respectively, and $I$ is the interference temperature constraint (ITC), which is the maximum tolerable level of interference that $PR$ can accept for a reliable PN communication. The transmission of the secondary network is executed in two time-slot periods. In the 1st time slot, $S$ transmits the signal to $R$ with the average power $P_S$. Thus, the received signal at $R$ can be written
as

\[ y_R = \sqrt{P_S h_{SR}} x_S + n_R, \]  

(3.3)

where \( x_S \), with \( \mathbb{E}(|x_S|^2) = 1 \), denotes the intended data to \( D \) and \( n_R \) denotes the additive white Gaussian noise (AWGN) term at \( R \). Then, the received signal at \( R \) are fully decoded and re-encoded for the further transmission with the signal-to-noise ratio (SNR) of

\[ \gamma_R = \frac{P_S |h_{SR}|^2}{\sigma_R^2}, \]  

(3.4)

where \( \sigma_R^2 \) is the noise power at \( R \). Hence, in the 2\textsuperscript{nd} time slot, \( R \) transmits a re-encoded version of the signal to \( D \) with an average power of \( P_R \), which can be expressed by

\[ y_D = \sqrt{P_R |h_{RD}|^2} \tilde{x}_S + n_D, \]  

(3.5)

where \( \tilde{x}_S \) is the decoded message at \( R \) while \( n_D \) is the AWGN term at \( D \). Then, \( D \) receives and estimates the original signal by the following SNR

\[ \gamma_D = \frac{P_R |h_{RD}|^2}{\sigma_D^2}, \]  

(3.6)

where \( \sigma_D^2 \) is the noise power at \( D \). Therefore, the end-to-end instantaneous system capacity can be written as [109], [110]

\[ C = \frac{1}{2} \min \left[ \log_2 \left( 1 + \min \left( \frac{I}{|h_{SP}|^2}, \frac{|h_{SR}|^2}{\sigma_R^2} \right) \right), \right. \]

\[ \left. \log_2 \left( 1 + \min \left( \frac{I}{|h_{RP}|^2}, \frac{|h_{RD}|^2}{\sigma_D^2} \right) \right) \right]. \]  

(3.7)

\[ ^1 \text{Thereafter, time index is omitted for brevity reasons.} \]
The factor $\frac{1}{2}$ in (3.7) implies that the $S \rightarrow R \rightarrow D$ communication session requires two time slots.

### 3.4 Outage Probability Analysis

In wireless communication networks, the OP of a communication channel is the probability that an achievable rate is not supported because of variable channel capacity. Thus, an outage event can occur if the obtained SNR level is below than a predefined SNR threshold [113]. Now, by defining the probability of the instantaneous capacity in (3.7) is below than a predefined rate threshold $R_{th}$, the OP of $D$ can be written by followings

$$P_{out} = \Pr(C \leq R_{th}).$$  \hspace{1cm} (3.8)

Furthermore, the end-to-end OP of $D$ can be further rewritten as follows

$$P_{out}(\xi) = \Pr \left[ \min \left( \min \left( \frac{I}{|h_{SP}|^2}, \bar{P}_S \right), \frac{|h_{SR}|^2}{\sigma_R^2} \right), \min \left( \frac{I}{|h_{RP}|^2}, \bar{P}_R \right), \frac{|h_{RD}|^2}{\sigma_D^2} \right) < \xi \right],$$  \hspace{1cm} (3.9)

where $\xi = 2^{2R_{th}} - 1$ is the corresponding threshold for the SNR. Now, we divide (3.9) into two terms to simplify the demonstration of the cumulative distribution functions (CDFs) of the system model as

$$\mathcal{O}_S \triangleq \min \left( \frac{I}{|h_{SP}|^2}, \bar{P}_S \right), \frac{|h_{SR}|^2}{\sigma_R^2} \right),$$  \hspace{1cm} (3.10)

and

$$\mathcal{O}_R \triangleq \min \left( \frac{I}{|h_{RP}|^2}, \bar{P}_R \right), \frac{|h_{RD}|^2}{\sigma_D^2} \right).$$  \hspace{1cm} (3.11)

Then, by substituting (3.10) and (3.11) into (3.9), also knowing the independence of the random variables (RVs) $\mathcal{O}_S$ and $\mathcal{O}_R$, the OP of $D$ can be further rewritten as

$$P_{out}(\xi) = \Pr (\min (\mathcal{O}_S, \mathcal{O}_R) \leq \xi) = 1 - \Pr (\min (\mathcal{O}_S, \mathcal{O}_R) > \xi)$$

$$= 1 - \Pr (\mathcal{O}_S > \xi) \Pr (\mathcal{O}_R > \xi) = 1 - (1 - F_{\mathcal{O}_S}(\xi)) (1 - F_{\mathcal{O}_R}(\xi))$$

$$= F_{\mathcal{O}_S}(\xi) + F_{\mathcal{O}_R}(\xi) - F_{\mathcal{O}_S}(\xi) F_{\mathcal{O}_R}(\xi),$$  \hspace{1cm} (3.12)
where $F_{\mathcal{D}_S}$ and $F_{\mathcal{D}_R}$ represent the CDFs of the RVs $\mathcal{D}_S$ and $\mathcal{D}_R$, respectively. Now, we compute the CDF of $\mathcal{D}_S$ by followings

$$F_{\mathcal{D}_S}(\xi) = \Pr(\mathcal{D}_S < \xi) = \Lambda + \Psi,$$

(3.13)

where

$$\Lambda = \Pr \left\{ \frac{\bar{P}_S |h_{SR}|^2}{\sigma_R^2} < \xi, \bar{P}_S \leq \frac{I}{|h_{SP}|^2} \right\} = \Pr \left\{ |h_{SR}|^2 < \frac{\xi \sigma_R^2}{\bar{P}_S}, |h_{SP}|^2 \leq \frac{I}{\bar{P}_S} \right\},$$

(3.14)

$$\Psi = \Pr \left\{ \frac{|h_{SR}|^2 I}{|h_{SP}|^2 \sigma_R^2} < \xi, \bar{P}_S > \frac{I}{|h_{SP}|^2} \right\} = \Pr \left\{ |h_{SR}|^2 < \frac{\xi \sigma_R^2}{I}, |h_{SP}|^2 > \frac{I}{\bar{P}_S} \right\}.$$

(3.15)

### 3.4.1 Outage Analysis for the Rayleigh Distribution

Here, we first derive an outage analysis assuming that all channels in the proposed system model experience an independent but not necessarily identically distributed (i.n.i.d.) Rayleigh block fading, i.e., the channels remain constant during every transmission block but vary independently between various blocks.

The probability density function (PDF) of the Rayleigh distribution is written as [114]

$$f_X(x) = \frac{x}{\sigma^2} e^{-\frac{x^2}{2\sigma^2}}, \ x \geq 0,$$

(3.16)

where $\sigma$ denotes the scale parameter of the distribution. Hence, the CDF of the distribution can be expressed as

$$F_X(x) = 1 - e^{-\frac{x^2}{2\sigma^2}}, \ x \in [0, \infty).$$

(3.17)

**Definition** Assume that $X$ is a continuous RV with generic PDF $f_X(x)$ which is defined over the support $n_1 < x < n_2$ and $Y = g(X)$ be an invertible function of $X$ with inverse function $X = s(Y)$. Hence, by applying the “change of variable” technique [115], the PDF of $Y$ can be written by

$$f_Y(y) = f_X(s(y)) |s'(y)|,$$

(3.18)

which can be defined over the support $g(n_1) < y < g(n_2)$.

As it mentioned before each channel between nodes, i.e., $h_{(m)}$, $\forall m \in \{SP, RP, SR, RD\}$, is i.n.i.d. with zero mean and unit variance. Therefore, using (3.18), $|h_{(m)}|^2$, $\forall m \in$
Chapter 3. Cognitive Relaying Networks

\{SP, RP, SR, RD\}, follows exponential distribution [116], which PDF and CDF can be respectively derived as

\[
f_Y(y) = \begin{cases} 
\lambda e^{-\lambda y}, & y \geq 0, \\
0, & y < 0
\end{cases}
\] (3.19)

and

\[
F_Y(y) = \begin{cases} 
1 - e^{-\lambda y}, & y \geq 0, \\
0, & y < 0
\end{cases}
\] (3.20)

where \(\lambda = \frac{1}{2\sigma^2}\).

**Proposition 1** The closed-form OP expression of \(D\) for the Rayleigh distribution can be written as

\[
P_{\text{out}}(\xi) = 1 - \left( e^{-\frac{\lambda\xi^2_{\text{PS}}}{\sigma^2_{\text{PS}}} - \frac{\xi^2\sigma^2_{\text{PS}}}{I + \xi^2\sigma^2_{\text{PS}}}} - \frac{\lambda(1 + \xi^2\sigma^2_{\text{PS}})}{I + \xi^2\sigma^2_{\text{PS}}^2} \right) \left( e^{-\frac{\lambda\xi^2_{\text{PR}}}{\sigma^2_{\text{PR}}} - \frac{\xi^2\sigma^2_{\text{PR}}}{I + \xi^2\sigma^2_{\text{PR}}}} - \frac{\lambda(1 + \xi^2\sigma^2_{\text{PR}})}{I + \xi^2\sigma^2_{\text{PR}}^2} \right). \] (3.21)

**Proof:** See Appendix A.1 for details.

\[\blacksquare\]

### 3.4.2 The \(\alpha-\mu\) Distribution

The \(\alpha-\mu\) distribution is a general fading distribution that can introduce the Gamma, Rayleigh, Exponential, Weibull, one-sided Gaussian, and Nakagami-\(m\) [117]. Saying so, the \(\alpha-\mu\) distribution can be turned into the Weibull distribution by inserting \(\mu = 1\). Hence, from the Weibull distribution, setting \(\alpha = 1\) and \(\alpha = 2\), the negative exponential and the Rayleigh distributions can be obtained, accordingly. Furthermore, from the \(\alpha-\mu\) distribution, the Nakagami-\(m\) distribution results when \(\alpha = 2\). Then, from the Nakagami-\(m\) distribution by setting \(\mu = 1\) and \(\mu = \frac{1}{2}\), the Rayleigh and the one-sided Gaussian distributions can be obtained, respectively.

#### 3.4.2.1 Outage Analysis for the \(\alpha-\mu\) Distribution

The \(\alpha-\mu\) distribution is the most appropriate fading distribution that can describe the small-scale fading channels [112, 118]. The PDF of the \(\alpha-\mu\) fading signal with an envelope \(r\) can be written as [119, 120]

\[
p(r)_{\alpha-\mu} = \frac{\alpha^\mu r^{\alpha\mu-1}}{\Gamma(\mu)} e^{-\frac{\alpha^\mu r}{\mu}}, \] (3.22)
where $\alpha > 0$ is an arbitrary parameter, $\hat{r}$ stands for $\alpha$-root mean value which is equal to $\hat{r} = \sqrt[\alpha]{\mathbb{E}(r^\alpha)}$, $\mathbb{E}(\cdot)$ is the expected value, and $\Gamma(\cdot)$ is the Gamma function defined as $\Gamma(s) \triangleq \int_0^\infty t^{s-1}e^{-t}dt$ [121]. Also, $\mu \geq \frac{1}{2}$ is the inverse of normalized variance of $r^\alpha$ given by

$$
\mu = \mathbb{E}^2(r^\alpha)/\mathbb{E}(r^{2\alpha}) - \mathbb{E}^2(r^\alpha) .
$$

(3.23)

The PDF, $f_\gamma(\gamma)$, and the CDF, $F_\gamma(\gamma)$, of the SNR can be expressed as [113]

$$
f_\gamma(\gamma) = \frac{\alpha \mu_{\gamma}(\alpha/2)^{\alpha/2}}{2^{\alpha/2} \Gamma(\mu)} e^{-\mu_{\gamma}(\alpha/2)}
$$

(3.24) and

$$
F_\gamma(\gamma) = \frac{\gamma_{\text{inc}}(\mu, \mu_{\gamma}(\alpha/2))}{\Gamma(\mu)} ,
$$

(3.25)

respectively, where $\gamma_{\text{inc}}(s, x) = \int_x^\infty t^{s-1}e^{-t}dt$ is the lower incomplete gamma function [121].

**Proposition 2** The closed-form expression for the CDFs $F_{\odot S}$ and $F_{\odot R}$ of $D$ for the $\alpha - \mu$ distribution can be respectively written as

$$
F_{\odot S}(\xi) = \frac{\Gamma(\mu_{SR}, \mu_{SR}\left(\frac{\xi_{SR}}{\gamma_{PS}}\right)^{\frac{\alpha_{SR}}{2}})}{\Gamma(\mu_{SR})} \frac{\Gamma(\mu_{SP}, \mu_{SP}\left(\frac{I_{PS}}{\gamma_{PS}}\right)^{\frac{\alpha_{SP}}{2}})}{\Gamma(\mu_{SP})} 
$$

$$
+ \frac{\Gamma(\mu_{SP}, \mu_{SP}\left(\frac{I_{PS}}{\gamma_{PS}}\right)^{\frac{\alpha_{SP}}{2}})}{\Gamma(\mu_{SP})} \frac{\mu_{SP}}{\mu_{SR}} \frac{\gamma_{SP}^{\alpha_{SR}/2}}{\gamma_{SR}^{\alpha_{SR}/2}} 
$$

$$
\times \sum_{i=0}^{\mu_{SR}-1} \left[ \left( \frac{\mu_{SR}\left(\frac{\xi_{SR}}{\gamma_{PS}}\right)^{\frac{\alpha_{SR}}{2}}}{i!} \right)^{\mu_{SP}+i} \left( \frac{\mu_{SP}\left(\frac{I_{PS}}{\gamma_{PS}}\right)^{\frac{\alpha_{SP}}{2}}}{\gamma_{SR}^{\alpha_{SR}/2}} + \mu_{SR}\left(\frac{\xi_{SR}}{\gamma_{PS}}\right)^{\frac{\alpha_{SR}}{2}} \right)^{\mu_{SP}+i} \right] 
$$

$$
\times \Gamma\left(\mu_{SP} + i, \left(\frac{I_{PS}}{\gamma_{PS}}\right)^{\frac{\alpha_{SP}}{2}} + \mu_{SR}\left(\frac{\xi_{SR}}{\gamma_{PS}}\right)^{\frac{\alpha_{SR}}{2}} \right) \right]
$$

(3.26)

and
\[ F_R(\xi) = \frac{\Gamma(\mu_{RD}, \frac{\xi\sigma_D^2}{\bar{P}_R})}{\Gamma(\mu_{RD})} \frac{\Gamma(\mu_{RP}, \frac{I}{\bar{P}_R})^{\alpha_{RP}}}{\Gamma(\mu_{RP})} \]
\[ + \frac{\Gamma(\mu_{RP}, \frac{I}{\bar{P}_R})^{\alpha_{RP}}}{\Gamma(\mu_{RP})} \frac{\mu_{RP}^{\alpha_{RP}+i}}{\Gamma(\mu_{RP})} \sum_{i=0}^{\mu_{RD}-1} i! \left( \frac{\mu_{RP}}{\bar{P}_R} \frac{\xi\sigma_D^2}{\bar{P}_R} \right)^{\frac{\alpha_{RD}}{2}} \left( \frac{\mu_{RP}}{\bar{P}_R} + \mu_{RD} \frac{\xi\sigma_D^2}{\bar{P}_R} \right)^{\frac{\alpha_{RP}}{2}} \right]. \]

\[ (3.27) \]

**Proof:** See Appendix A.2 for details.  

Finally, the closed-form OP expression of \( D \) for the \( \alpha - \mu \) distribution can be derived by inserting (3.26) and (3.27) into (3.12).

The impact of the ITC on the OP can be explained by the next example. As it is indicated in the literature [109, 110], an increase of the transmit power of \( S \) and \( R \) saturates the OP of the destination user. That occurs because the transmit power, with respect to \( \min(I/|h|^2, \bar{P}) \), will be equal to \( (I/|h|^2) \) as \( P \to \infty \), i.e., a fixed \( I \) will control the maximum transmitted power from \( S \) and \( R \). Therefore, if the value of \( I \) increases it allows \( S \) and \( R \) to transmit at higher power levels, which further results in an improvement of the OP performance. The importance of \( I \) can be shown by another example when \( I \to 0 \). In that case, \( F_{\tilde{S}}(\xi) = F_{\tilde{R}}(\xi) \approx 1 \) which comes to \( P_{\text{out}}(\xi) \approx 1 \). Hence, \( PR \) cannot tolerate any interference from the SN transmitters when \( I \to 0 \), which means that a communication in the SN is unavailable.

### 3.4.3 Symbol Error Rate Analysis

The symbol error rate (SER) for binary phase shift keying (BPSK), in traditional wireless communications, can be written as [122]

\[ \text{SER} = \frac{1}{\pi} \int_0^{\pi/2} \mathcal{M}_\gamma \left( -\frac{g_{\text{psk}}}{\sin^2(\theta)} \right) d\theta, \]

where \( g_{\text{psk}} = \sin^2(\pi/M) \) denotes a constant value which equals to unity for BPSK while \( \mathcal{M}_\gamma(-g) \) indicates the moment generating function (MGF) of the obtained SNR which
can be written as

\[ \mathcal{M}_\gamma(-g) = \int_0^\infty f_\gamma(x)e^{-gx}dx. \] (3.29)

Thus, to evaluate the SER, the PDF of (3.12) should be substituted into (3.29). Hence, by using the derived MGF in (3.28), the SER can be calculated numerically.

### 3.5 Numerical Results and Discussion

This section discusses the OP and SER performance of the proposed system model for different cases of the $\alpha - \mu$ distribution.

Fig. 3.3 illustrates the results when the Weibull distribution ($\mu = 1$) is adopted. The comparison is made among various values of $\alpha$ while the ITC is taken as 10 dB and 20 dB. Moreover, $\xi$ is set to be equal to 5 dB. The simulated results validate the analytical ones by shown perfect match. The results show that the best performance of the system is attained when no ITC is implemented by $PR$. Furthermore, when $\alpha = 1.5$, the system obtains better performance compared with the case when lower values of $\alpha$ are set. For instance, when $I = 20$ dB, the system with parameters of $\alpha = 1.5$ and $\alpha = 0.75$ obtains the OP of 0.6 at 10 dB and 12 dB, accordingly. From these results, we can observe the impact of the arbitrary fading parameter on the OP. In addition, it is shown that the curve of the OP starts saturating when the value of the transmit SNR approaches to that of $I$. 

![Figure 3.3. The OP versus SNR performance for various ITC over the Weibull distribution.](image-url)
This can be explained by the fact that, due to the ITC, the secondary transmit nodes cannot broadcast at a higher power level than $\frac{I}{|h|^2}$.

Fig. 3.4 demonstrates the OP depicted for the Rayleigh distribution with $\alpha = 2$ and $\mu = 1$ as well as for the Negative Exponential distribution with $\alpha = 2$ and $\mu = 1/2$ when
various ITC values are implemented. From the figure, it is clearly seen that the Rayleigh distribution channel characteristics demonstrate better OP compared with the Negative Exponential one. For example, when $I = 20$ dB and the transmit SNR equals to 10 dB, the former obtains the OP of 0.2 while the latter achieves the OP of 0.7.

In Fig. 3.5, the SER given in (3.28) when the channel follows the Weibull distribution is depicted with $I = 20$ dB. In general, it is noted that networks with larger values of $\alpha$ achieve much better performance compared to that of lower values of $\alpha$. Indicatively, the network with $\alpha = 1.5$ attains the SER of 0.1 at about transmit SNR of 6.5 dB, while, when $\alpha = 1$ and $\alpha = 0.75$, the same SER can be obtained at about 9.5 dB and 12.5 dB, respectively. Finally, it is noticed from all SER results that the system obtains better SER performance in higher transmit SNRs. However, curves of the SER start saturating while transmit SNR tends to the value of $I$.

### 3.6 Chapter Summary

This Chapter introduced the $\alpha-\mu$ distribution in the dual-hop CRN where secondary transmit nodes were constrained from the maximum transmit power due to the underlay CR concept. The $\alpha-\mu$ is a general distribution that includes several others distributions, i.e., Rayleigh, Gamma (and its discrete versions Erlang and central Chi-squared), Exponential, Nakagami-$m$ (and its discrete version Chi), One-sided Gaussian, and Weibull. Based on the dual-hop CRN model proposed here, the analytical results for the OP were obtained in closed forms. Moreover, the impact of the ITC on the system outage and SER performance were investigated. From the derived results, it was observed that the ITC degrades the outage and SER performances which can be seen from the outage saturation. It was also revealed that the lower values of $I$ degrade the OP performance of the system. In addition, the correctness of the derived results was validated by the Monte-Carlo simulations, where the perfect match between simulation and analytical results were obtained.

For the contribution of this Chapter, we can indicate that the statistics of those distributions that are special cases of the $\alpha-\mu$ distribution can be obtained directly from the formulations developed here by simply reducing these formulations into the respective particular case. It is worth to mention that even the $\alpha-\mu$ distribution has one more parameter than Weibull or Nakagami-$m$ distributions, it does not pose any additional mathematical difficulty. Hence, the $\alpha-\mu$ distribution is suitable to adjust to field data by representing many other distributions in its special cases.
Chapter 4

Wireless Powered CRN with Interference Alignment

This Chapter studies a dual-hop DF wireless powered CRN with IA over Rayleigh fading channels. The energy-constrained secondary relay node harvests energy from both information and interference signals. Then, the secondary relay uses the harvested energy to forward the information signal from the secondary source to destination node. By applying beamforming matrices to the PNs and SNs, the performance metrics such as the OP, capacity and the bit error rate are studied under perfect and imperfect channel state information scenarios for both power-splitting relaying and time-switching relaying protocols. Moreover, the optimal network performance is achieved by calculating the optimal energy harvesting time-switching and power-splitting coefficients. Finally, closed-form OP expressions of primary and secondary users are derived. Monte-Carlo simulations are used to corroborate the analytical results, which show that the power-splitting relaying technique performs better than time-switching relaying in the proposed system model.

4.1 Introduction

4.1.1 Wireless Powered CRNs with IA

It was clearly explained in the previous chapters that SUs in CR systems can communicate in the licensed spectrum using three approaches, i.e., interweave, underlay and overlay [5]. These approaches impose some restrictions on their operation modes which, in turn, may degrade the performance of SUs. For instance, in the case of the underlay CR, the PUs impose the ITC to the SUs to guarantee the tolerable interference level at primary receivers (PRs). However, by employing the IA technique [25, 123–126], the interference caused by the SUs in wireless multi-user CR systems can be managed without imposing any ITC to the SUs. Furthermore, the advances of EH can be applied in CR systems [127] as CR nodes are mostly battery-powered devices which need to recharge or replace their batteries periodically.
Chapter 4. Wireless Powered CRN with Interference Alignment

The underlay CR with embedded IA technique was studied in [116, 128–130]. The authors in [128] employed IA in CR to gain frequency domain diversity at the PUs to increase the throughput of the SN. A cancellation algorithm for a practical IA was developed in [130], where the authors studied the interference management at the PUs and the DoF optimization of the SUs. In [129], the authors presented the DoF region in a MIMO CR with a cooperative relay and embedded IA algorithm.

There are main two EH approaches in CR systems, i.e, time-switching (TS) and power-splitting (PS) protocols [131]. In the TS method (see Fig. 4.1a), the receiver switches between EH and information decoding (ID) modes over time [127]. In the PS method (see Fig. 4.1b), at the receiver’s end, the received power is divided into two fractions: one is devoted for EH, whereas the rest is dedicated to ID purposes [127]. There are many research works related to SWIPT networks in the literature [50, 127, 132–138]. In [134], the authors proposed a joint study of IA and SWIPT in MIMO networks, where the performance improvement of the SWIPT network by dynamically selecting users as either EH or ID terminals was illustrated. The performance of the EH method in cooperative relaying networks, where a power-limited relay harvests energy to assist the source to deliver the signal to the destination, was investigated in [127, 133, 135, 139–141]. Lately, CR systems with the ability of EH have received significant attention [136, 137, 142]. The authors in [136] studied an optimal spectrum sensing policy for the EH-based CR systems.
which maximizes the expected total throughput under energy causality and collision constraints. In the proposed system model, secondary nodes harvest the energy from PUs’ transmissions within the TS mode. In [116], a novel SWIPT scheme to optimize wireless energy harvesting (WEH) and information transfer (IT) performance in symmetric IA networks were introduced. In all the aforementioned works, the authors assumed perfect CSI which is impractical due to the presence of channel estimation errors [143]. Thus, research works in [143] and [144] investigated the performance of the EH-based overlay CR under the imperfect CSI scenario. The authors in [143] studied the joint optimization of the PS ratio selection and power control, whereas, in [144], joint resource allocation and admission control schemes with several SUs were studied.

4.1.2 Practical Implementation

IA embedded systems may face some several challenging problems in a practical implementation, i.e., CSI acquisition, CSI overhead signaling and degradation of the system performance at low SNRs. These challenges were studied in [145–147]. The authors in [145] proposed an adaptive transmission scheme which considerably reduces the error probability and produces robustness to the CSI uncertainty. In addition, in [146], the authors presented a sequential antenna switching IA scheme that reduces the computational complexity and CSI-feedback overhead. Moreover, a low complexity algorithm for CSI feedback was developed in [147], which allows reducing feedback dimension while maintaining the IA feasibility. Furthermore, the authors in [146, 148] improved the IA network performance at low SNR values., while a MIMO-based orthogonal frequency division multiplexing interference network was initially studied practically in [149], where channel gains measured using a flexible software-defined radio were consistent with theoretical and simulation results. In [150], a transmit antenna selection was introduced in order to increase the sum-rate performance of the IA-based MIMO network, where authors verified the practical feasibility of the system through practical implementation. These studies have demonstrated a significant move to the IA implementation in practice. Hence, IA techniques can be efficiently coupled with CR and EH since both latter techniques have been already used in practical systems [151–153].

4.1.3 Motivation

Nowadays, numerous techniques are considered to be a candidate that can be executed in the 5G communication networks. The main conditions required for the 5G candidates are higher spectral and energy efficiency, lower end-to-end latency, more connected nodes, etc. [154]. Therefore, IA-embedded networks can improve the spectral efficiency by mitigating interference while the CRNs can utilize unused spectrum bands to connect more
nodes. In addition, 5G networks can be more energy efficient using RF SWIPT with different EH protocols. Hence, a joint study of the CR, IA and EH techniques could contribute a thorough insight into how the next generation communication networks should be designed. Despite a large number of research works on the CR, IA and SWIPT, all of them have been studied separately. In other words, in the literature, each of these techniques has been studied separately from the other two technologies. For example, recent IA advances were ignored in CR or SWIPT systems \[50, 80, 132\]. Moreover, the interference in IA networks was just removed instead of being employed for EH goals in \[116, 129\]. In addition, the SN’s performance was suffered from the limited transmit power due to the applied ITC from the PRs in underlay CR systems, while the interference at PRs might be managed through IA \[136, 155\]. Hence, the main aim of this Chapter is to investigate IA-based underlay CR systems with the EH-enabled relay over Rayleigh fading channels in the case of the perfect and the imperfect CSI.

4.2 Contribution of the Chapter

Main contributions are as follows:

- The performance of the PRs and the EH-based relay in the TS and PS modes under the interference presence is investigated. Moreover, precoding and interference suppression beamforming (BF) matrices are applied to mitigate the negative effect of interference at primary and secondary receivers. The management of interference allows avoiding ITC implementation at the PRs, which consequently gives a possibility to increase the power level at the secondary transmitters.

- The instantaneous capacity performance is assessed for the PUs and SUs in the TS and PS methods. In addition, the optimal values of TS and PS portions are determined for different CSI scenarios. From the obtained results, PS relaying (PSR) shows better performance than the TS relaying (TSR). Then, some useful insights into how the CSI quality impacts on BER and capacity performance for the TS and PS-based systems are provided.

- Closed-form OP expressions of the PUs and the SU are determined. Moreover, Monte-Carlo simulations for the OP are consistent with the analytical results showing that the performance of the PSR mode outperforms that of the TSR system for secondary destinations, while, for the PRs in the 1st time slot, the OP of the TSR outperforms that of the PSR.

The remainder of the Chapter is organized as the next. The system model with employed IA for perfect and imperfect CSI scenarios is described in Section 4.3, while the
study of TSR and PSR protocols with instantaneous capacity derivations are presented in Sections 4.4 and 4.5, respectively. Section 4.6 derives the OP expressions of the PUs and the SUs. Numerical results are presented and discussed in Section 4.7. Finally, the Chapter is concluded in Section 4.8.

*Notation:* Vectors and matrices are denoted by lowercase and uppercase bold symbols, e.g., $\mathbf{x}$ and $\mathbf{X}$, respectively, while $\mathbf{I}$ represents the identity matrix. The superscripts $\cdot^\dagger$ and $\cdot[\iota]$ stand for the Hermitian transpose and time slot index, accordingly. $\text{tr}(\mathbf{X})$ calculates the trace of $\mathbf{X}$. $\|\mathbf{x}\|$ and $|x|$ denote the Euclidean norm of $\mathbf{x}$ and the absolute value of $x$, respectively. $\mathbb{E}(\cdot)$ and $\mathbb{V}(\cdot)$ are the expectation and variance operators, respectively. $\mathbb{C}^{N\times M}$ is the space of complex $N \times M$ matrices. $\mathcal{CN}(a, b)$ denotes the complex normal distribution with $a$ mean and $b$ variance. Finally, $\text{span}(\mathbf{X})$ and $\text{eig}(\mathbf{X})$ denote the vector space spanned by the column vectors and the eigenvectors of $\mathbf{X}$, accordingly.
4.3 System Model

The proposed system model is shown in Fig. 4.2, where a MIMO CRN consisting of one SN and L PU-pairs is presented. Each primary transmitter \( T_i, i \in \{1, 2, ..., L\} \), communicates to its corresponding receiver \( PR_j, j \in \{1, 2, ..., L\} \), by causing interference to the other PRs and SUs. The SN consists of \( S, R \) and \( D \). An energy-constrained \( R \) can scavenge the energy from the desired and interfering signals. It is assumed that \( R \) utilizes all the harvested energy at each EH time interval to maintain the \( S-D \) communication within the half-duplex DF relaying. Other nodes than \( R \), i.e., \( T_i, PR_j, S \) and \( D \) are assumed to be provided with external power sources. It is also assumed that all nodes support MIMO antennas, where each appropriate channel remains constant through a transmission time slot \( \tau \), with \( \tau \in \{1, 2\} \), and varies independently in every time period.

In the PN, the intra-network channels between \( T_i \) and \( PR_j \) are denoted by \( H_{j,i}^{[l]} \in \mathbb{C}^{N_j \times M_l}, \forall i, j \in \{1, 2, ..., L\} \), where \( PR_j \) and \( T_1 \) nodes are deployed with \( N_j \) and \( M_l \) antennas, respectively. In the SN, channels between \( S-R \) and \( R-D \) transmission links are described by \( H_{R,S} \in \mathbb{C}^{N_R \times N_S} \) and \( H_{D,R} \in \mathbb{C}^{N_D \times N_R} \), accordingly. \( N_S, N_R \) and \( N_D \) denote the numbers of antennas at \( S, R \) and \( D \), respectively. Inter-network interference channels are denoted by \( H_{j,R} \in \mathbb{C}^{N_j \times N_R} \), \( H_{j,S} \in \mathbb{C}^{N_j \times N_S} \) and \( H_{R,i} \in \mathbb{C}^{N_R \times M_l} \).

Moreover, every channel link can be characterized by the corresponding distance and the path loss exponent\(^1\), i.e., \( d_{m,n} \) and \( \tau_{m,n}, \forall n \in \mathcal{A} = \{1, 2, ..., L, S, R\} \) and \( \forall m \in \mathcal{B} = \{1, 2, ..., L, R, D\} \). Each entry of \( H \) is assumed to be an i.i.d Rayleigh block fading modeled as \( \mathcal{CN}(0, 1) \).

It is assumed that a global CSI is available at each node\(^2\). Each transmit node \( l \) with the power of \( P_l \) employs a precoding BF matrix \( V_l \in \mathbb{C}^{M_l \times N_l} \), with \( \text{tr}\{V_l V_l^\dagger\} = 1 \), \( \forall l \in \mathcal{A} \), where \( f_l \) is the number of data streams that is sent from the corresponding transmit node. Furthermore, every receiver employs the interference suppression matrix \( U_m \in \mathbb{C}^{N_m \times f_m}, \forall m \in \mathcal{B} \), where \( f_m \) denotes the number of data streams. Thus, \( PR_j \), within the time slot \( \tau \), obtains the following signal

\[
y_j^{[l]} = \sum_{i=1, i \neq j}^{L} \sqrt{P_i d_{j,i}^{-\tau_{m,n}}} H_{j,i}^{[l]} V_{i}^{[l]} s_i + F^{[l]} + n_j^{[l]} + \sqrt{\frac{P_j}{d_{j,j}^{-\tau_{m,n}}}} U_{j}^{[l]} H_{j,j}^{[l]} V_{j}^{[l]} s_j, \quad (4.1)
\]

\(^1\)In wireless communications, the power attenuation increases logarithmically with the distance between the transmitter and the receiver [156].

\(^2\)In different works in the literature, it is common to accept that any primary node exploits pilot signals which can be systematically transmitted to maintain the channel estimation and synchronization [157, §12.3.1], [158]. Saying so, each secondary node is able to attain own channel estimation between the primary/secondary transmitters and itself, by the condition that the secondary node participates in this training process.
where \( \tilde{n}_j^{[i]} = U_j^{[i]^\dagger} n_j^{[i]} \) denotes the AWGN vector with the effective zero mean at the output of the beamformer, with \( \mathbb{E}\{\tilde{n}_j^{[i]} \tilde{n}_j^{[i]^\dagger}\} = \sigma_n^2 I \). Moreover, \( \mathbb{E}\{s_j s_i^\dagger\} = I \), with \( \forall l \in \mathcal{A} \), where the \( s_j \) vector contains symbols of the desired constellation taken from i.i.d. Gaussian input signaling. Thus, all transmit nodes satisfy the average power constraint by meeting all the aforementioned conditions. Then, interference caused by the SN to \( PR_j \) denotes as

\[
F^{[i]} = \begin{cases} \sqrt{\frac{P_j}{d_{j,S}^R}} U_j^{[i]^\dagger} H_{j,S} V_S s_S, & \text{if } \iota = 1, \\ \sqrt{\frac{P_j}{d_{j,R}^R}} U_j^{[2]^\dagger} H_{j,R} V_R s_R, & \text{if } \iota = 2. \end{cases} \tag{4.2}
\]

Within 1st time slot of the relaying process, \( R \) receives the following signal

\[
y_R = \sqrt{\frac{P_S}{d_{R,S}^S}} U_R^{\dagger} H_{R,S} V_S s_S + \sum_{i=1}^{L} \sqrt{\frac{P_i}{d_{R,i}^R}} U_R^{\dagger} H_{R,i} V_i^{[1]} s_i + \tilde{n}_R, \tag{4.3}
\]

where \( \tilde{n}_R = U_R^{\dagger} n_R \) denotes the effective AWGN noise at \( R \) after the applied interference suppression BF matrices.

Furthermore, \( R \) decodes the desired signal \( s_S \) and forwards it to \( D \). Thus, the received signal at \( D \) can be written as

\[
y_D = \sqrt{\frac{P_R}{d_{D,R}^D}} U_D^{\dagger} H_{D,R} V_{R} s_R + \sum_{i=1}^{L} \sqrt{\frac{P_i}{d_{D,i}^D}} U_D^{\dagger} H_{D,i} V_i^{[2]} s_i + \tilde{n}_D, \tag{4.4}
\]

where \( \tilde{n}_D = U_D^{\dagger} n_D \) denotes the effective AWGN noise at \( D \).

All the interference at receivers can be perfectly managed when IA is feasible. Therefore, to cancel those interference, the following conditions should be satisfied at \( PR_j \) [125]

\[
U_j^{[i]^\dagger} H_{j,i}^{[i]} V_i^{[i]} = 0, \quad \forall i, j \in \{1, 2, \ldots, L\} \text{ and } \forall i \neq j, \tag{4.5a}
\]

\[
U_j^{[i]^\dagger} W^{[i]} = 0, \quad \text{where } W^{[i]} = \begin{cases} H_{j,S} V_S, & \text{if } \iota = 1, \\ H_{j,R} V_R, & \text{if } \iota = 2. \end{cases} \tag{4.5b}
\]

\[
\text{rank } (U_j^{[i]^\dagger} H_{j,j}^{[i]} V_j^{[i]}) = f_j, \quad \forall j \in \{1, 2, \ldots, L\}, \tag{4.5c}
\]
and at $R$ and $D$ as

\begin{align}
U^\dagger_R H_{R,i} V_i^{[1]} &= 0, \forall i \in \{1, 2, ..., L\}, \quad (4.6a) \\
\text{rank} \left( U^\dagger_R H_{R,S} V_S \right) &= f_S, \quad \text{and} \quad (4.6b) \\
U^\dagger_D H_{D,i} V_i^{[2]} &= 0, \forall i \in \{1, 2, ..., L\}, \quad (4.6c) \\
\text{rank} \left( U^\dagger_D H_{D,R} V_R \right) &= f_D, \quad \text{respectively}. \quad (4.6d)
\end{align}

When the conditions given in (4.5) are met, there is no need for $PR_j$ to apply the ITC to the SN. Hence, the secondary transmit nodes can broadcast with any available power level [116].

This Chapter concentrates on SWIPT instead of a DoF similarly as in [116] and [134]. Thus, each transmitter is assumed to broadcast only one data stream $f_l, \forall l \in A$, to the dedicated user. Hence, the BF matrices can be replaced by the corresponding vectors. Moreover, as symmetric networks are considered, all nodes are assumed to have the same antenna parameters, i.e., $M_i = N_S = M, N_j = N_R = N_D = N$. Therefore, due to the IA feasibility analysis, the largest number of user pairs in the proposed system should satisfy the following condition [116]

\begin{equation}
L + 1 \leq M + N - 1. \quad (4.7)
\end{equation}

In this chapter, the QoS of $PR_j$ is considered as the service of the PUs in a high priority. The transmission rate can directly reflect the variability of PUs’ QoS, so a rate threshold $R_{th}^j$ is defined according to the QoS requirement of $PR_j$. Hence, the QoS of $PR_j$ can be satisfied if $R_j \geq R_{th}^j$, which means that SUs can not transmit over the licensed band unless the mentioned condition is satisfied.

### 4.3.1 Imperfect CSI

An assumption of the perfect CSI is highly idealistic in practice. Thus, this Chapter examines the effect of imperfect CSI on the IA cancellation at the receiving node. The CSI mismatch can be described by [125]

\begin{equation}
\hat{H} = H + E, \quad (4.8)
\end{equation}

where $\hat{H}$ is the observed mismatched channel, $H \sim \mathcal{CN}(0, I)$ stands for the real channel matrix and $E$ denotes the inaccuracy degree in the estimated CSI which is independent from $H$. The $E$ can be described, considering the nominal transmit SNR $\rho$, as $E \sim \mathcal{CN}(0, \lambda I)$ with $\lambda = \psi \rho^{-\kappa}$, where $\lambda$ is an error variance which represents different acquisition scenarios for any constant values of $\kappa \geq 0$ and $\psi > 0$. Considering this
model, the perfect CSI can be obtained when $\kappa \to \infty$ for $\rho > 1$. Hence, the real channel matrix, conditioning on $\hat{H}$, can be written as [159]

$$
H = \frac{1}{1 + \lambda} \hat{H} + \tilde{H},
$$

(4.9)

where $\tilde{H} \sim \mathcal{CN}(0, \frac{1}{1+\lambda} I)$ is independent of $\hat{H}$.

### 4.4 Time-Switching Relaying

Fig. 4.3 depicts the TSR protocol, where the required time for the $S-D$ information transmission is taken as $T$ and the time fractions of $\alpha T$, with $0 \leq \alpha \leq 1$, is allotted for EH, whereas the other two equal time proportions $(1 - \alpha)T/2$ are allocated for the $S-R$ and $R-D$ data transmission sessions, accordingly [127]. In the case of the PN, $T_i$ conveys its information to the respective receiver over time $T$, which can be divided into two time periods, i.e., $(1 + \alpha)T/2$ and $(1 - \alpha)T/2$, to align with the time structure of the SN. Due to the proposed TSR system model, within the EH time period, $R$ receives the following signal

$$
y_{R}^{EH} = \sqrt{\frac{P_S}{d_{R,S}^{\text{TH}}}} H_{R,S} v_{s_s} + \sum_{i=1}^{L} \sqrt{\frac{P_i}{d_{R,i}^{\text{TH}}}} H_{R,i} v_{i}^{[1]} s_i + n_R.
$$

(4.10)

Then, by neglecting the relatively small energy scavenged from the noise, the harvested energy at $R$ can be written as [127]

$$
E_{H}^{\text{TSR}} = \eta \alpha T \left( \frac{P_S \|H_{R,S} v_s\|^2}{d_{R,S}^{\text{TH}}} + \sum_{i=1}^{L} \frac{P_i \|H_{R,i} v_i^{[1]}\|^2}{d_{R,i}^{\text{TH}}} \right).
$$

(4.11)
The relation between harvested energy and the relay transmit power can be written as

$$P_{TSR}^{R} = E_{H}^{TSR}/((1 - \alpha)T/2). \quad (4.12)$$

Moreover, by assuming that all the harvested power is utilized for data transmission, we can rewrite the relay transmit power by using (4.11) as

$$P_{TSR}^{R} = \frac{2\alpha \eta}{1 - \alpha} \left( \frac{P_{S} ||H_{R,S}v_{S}||^{2}}{d_{R,S}^{TSR}} + \sum_{i=1}^{L} \frac{P_{i} ||H_{R,i}v_{i}^{[1]}||^{2}}{d_{R,i}^{TSR}} \right), \quad (4.13)$$

where $\eta$ denotes the EH conversion efficiency, with $0 < \eta < 1$. During the $S$-$R$ data transmission phase, considering imperfect CSI in (4.9) and interference suppression vectors, the received information signal at $R$ can be written as

$$y_{R}^{IT} = \sqrt{\frac{P_{S}}{d_{R,S}^{TSR}}} u_{R}^{†} \left( \frac{1}{1 + \lambda} H_{R,S} + \hat{H}_{R,S} \right) v_{S} s_{S} + \sum_{i=1}^{L} \sqrt{\frac{P_{i}}{d_{R,i}^{TSR}}} u_{R}^{†} \left( \frac{1}{1 + \lambda} H_{R,i} + \hat{H}_{R,i} \right) v_{i}^{[1]} s_{i} + \hat{n}_{R} \quad (4.14)$$

The corresponding signal-to-interference-plus-noise ratio (SINR) at $R$, from (4.14), can be written as

$$\gamma_{R} = \frac{\frac{P_{S}}{d_{R,S}^{TSR}} ||u_{R}^{†} \hat{H}_{R,S}v_{S}||^{2}}{\frac{P_{S}}{d_{R,S}^{TSR}} ||u_{R}^{†} \hat{H}_{R,S}v_{S}||^{2} + I_{TSR}^{PN} + \sigma_{\hat{n}_{R}}^{2}}, \quad (4.15)$$

where $\sigma_{\hat{n}_{R}}^{2}$ is the noise power at $R$ and interference from $T_{i}$ is denoted by

$$I_{TSR}^{PN} = \sum_{i=1}^{L} \frac{P_{i} ||u_{R}^{†} \hat{H}_{R,i}v_{i}^{[1]}||^{2}}{d_{R,i}^{TSR}}. \quad (4.16)$$
Similar to (4.14), the signal received by $D$ can be written as

$$
y_D = \frac{1}{1 + \lambda} \sqrt{\frac{P_{TSR}}{d_{D,R}^j}} u_D^\dagger \hat{H}_{D,R} v_{R} S_R + \sqrt{\frac{P_i}{d_{D,i}^j}} u_D^\dagger \hat{H}_{D,i} v_i^{[2]} s_i + \tilde{n}_D.
$$

Further, the respective SINR at $D$ can be written as

$$
\gamma_D = \frac{\frac{P_{TSR}}{d_{D,R}^j} \| u_D^\dagger \hat{H}_{D,R} v_R \|^2}{\frac{P_{TSR}}{d_{D,R}^j} \| u_D^\dagger \hat{H}_{D,R} v_R \|^2 + G_{PN} + \sigma_{\tilde{n}_D}^2},
$$

where $\sigma_{\tilde{n}_D}^2$ denotes the noise power at $D$ and interference from $T_i$ is denoted by

$$
G_{PN} = \sum_{i=1}^L \frac{P_i \| u_D^\dagger \hat{H}_{D,i} v_i^{[2]} \|^2}{d_{D,i}^j}.
$$

Furthermore, the SINR for $PR_j$ is derived as

$$
\gamma_j = \frac{P_j \| u_j^{[\ell]} \hat{H}_{j,j} v_j^{[\ell]} \|^2}{d_{j,j}^\ell (1 + \lambda)^2 \left( A^{[\ell]} + B_{TSR}^{[\ell]} + \sigma_{n_j}^{[\ell]} \right)},
$$

where the intra-network interference from another $T_i$ due to the CSI mismatch is written as

$$
A^{[\ell]} = \frac{P_j \| u_j^{[\ell]} \hat{H}_{j,j} v_j^{[\ell]} \|^2}{d_{j,j}^\ell} + \frac{P_i \| u_i^{[\ell]} \hat{H}_{i,j} v_i^{[\ell]} \|^2}{d_{i,j}^\ell},
$$

while the inter-network interference from SN is denoted by

$$
B_{TSR}^{[\ell]} = \begin{cases} \frac{P_{SR}^{[\ell]}}{d_{j,s}^\ell} \| u_j^{[\ell]} \hat{H}_{j,s} v_S \|^2, & \text{if } \ell = 1, \\
\frac{P_{TSR}^{[\ell]}}{d_{j,s}^\ell} \| u_j^{[2]} \hat{H}_{j,R} v_R \|^2, & \text{if } \ell = 2. 
\end{cases}
$$

### 4.4.1 Capacity for TSR

The instantaneous capacity for $PR_j$ can be written as [160]

$$
C_j = \log_2(1 + \gamma_j),
$$
where $\gamma_j$ denotes the SINR of $PR_j$. Now, after substituting (4.20) into (4.23) and considering the exploited two-time slots, the instantaneous capacity for the $P_j$ is derived as

$$ C_{TSR}^j = 2 \sum_{t=1}^{2} j[t] \log_2 \left( 1 + \frac{P_j d_{j,j}^2 (1+\lambda)^2}{A[t] + B_{TSR}^{|j|} + \sigma_{n_j}^2} \right), \quad (4.24) $$

where $J[t] = \begin{cases} 1 + \alpha^2, & \text{if } t = 1, \\ 1 - \alpha^2, & \text{if } t = 2. \end{cases}$

Moreover, the end-to-end instantaneous capacity for $D$ can be calculated by the weakest channel between the $S-R$ and $R-D$ transmission channels, which can be written as

$$ C_D = \frac{1}{2} \log_2 \left( 1 + \min(\gamma_R, \gamma_D) \right). \quad (4.25) $$

Hence, by using (4.15), (4.18) and (4.25), the instantaneous capacity expression in (4.25) can be rewritten for the TSR protocol as

$$ C_{TSR}^D = \frac{1 - \alpha^2}{2} \log_2 \left[ 1 + \min \left( \frac{P_S d_{R,S}^2 (1+\lambda)^2}{2} \| u_R^j H_{R,S} v_S \|^2 + I_{PN} + \sigma_{n_R}^2, \frac{P_{TSR}^D d_{D,R}^2 (1+\lambda)^2}{2} \| u_D^j H_{D,R} v_R \|^2 + G_{PN} + \sigma_{n_D}^2 \right) \right]. \quad (4.26) $$

As the derivation of (4.26) is complicated, the instantaneous capacity can be assessed only numerically.

### 4.5 Power-Splitting Relaying

In the PSR protocol, the total time period of $T$ is divided into two equal time fractions to support the $S-R$ and $R-D$ data transmissions [138] (see Fig. 4.4). During the first half of the time period, $R$ utilizes a coefficient $\theta$ ($0 < \theta < 1$) to determine a part of the obtained signal power to harvest, while the remaining signal portion $(1 - \theta)$ is used for the $S-R$ data transmission purpose. Thus, within the EH time period, the obtained signal at $R$ can be written as

$$ y_{EH}^R = \sqrt{\frac{\theta P_S d_{R,S}^2}{d_{R,S}} H_{R,S} v_S s_S} + \sum_{i=1}^{L} \sqrt{\frac{\theta P_i}{d_{R,i}^2}} H_{R,i} v_i[i] s_i + \sqrt{\theta} n_R. \quad (4.27) $$
Then, $R$ harvests the following signal [127]

$$E^{PSR}_H = \frac{\eta T}{2} \left( \frac{P_S \|H_{R,S}v_S\|^2}{d_{R,S}^{H}} + \sum_{i=1}^{L} \frac{P_i \|H_{R,i}v_i^{[1]}\|^2}{d_{R,i}^{H}} \right).$$ \hspace{1cm} (4.28)

The relationship between the relay transmit power and the harvested energy for PSR protocol can be described by [127]

$$P^{PSR}_R = \frac{2E^{PSR}_H}{T}.$$ \hspace{1cm} (4.29)

Now, the transmit power at $R$ can be rewritten as

$$P^{PSR}_R = \eta T \left( \frac{P_S \|H_{R,S}v_S\|^2}{d_{R,S}^{H}} + \sum_{i=1}^{L} \frac{P_i \|H_{R,i}v_i^{[1]}\|^2}{d_{R,i}^{H}} \right).$$ \hspace{1cm} (4.30)

Furthermore, using (4.9) and applying interference suppression vector $u_R$, the received signal at $R$ is calculated by

$$y^{IT}_R = \frac{1}{1 + \lambda} \sqrt{\frac{(1 - \theta)P_S}{d_{R,S}^{F}}} u_R^\dagger H_{R,S}v_S s_S + \sqrt{\frac{(1 - \theta)P_S}{d_{R,S}^{F}}} u_R^\dagger H_{R,S}v_S s_S$$

$$+ \sum_{i=1}^{L} \sqrt{\frac{(1 - \theta)P_i}{d_{R,i}^{F}}} u_R^\dagger H_{R,i}v_i^{[1]} s_i + \tilde{n}_R.$$ \hspace{1cm} (4.31)

Hence, the corresponding SINR at $R$ can be derived by

$$\gamma_R = \frac{\frac{P_S(1-\theta)}{d_{R,S}^{F}(1+\lambda)^2} \|u_R^\dagger H_{R,S}v_S\|^2}{I^{PSR} + I^{PN}_R + \sigma^2_{n_R}},$$ \hspace{1cm} (4.32)
where interference from $T_i$ is denoted by

$$I_{PSR}^{PN} = \sum_{i=1}^{L} (1 - \theta_i) P_i ||u_{R,i}^\dagger \hat{H}_{R,i} v_i^1||^2 d_{R,i}^{-1}. \tag{4.33}$$

Further, $D$ receives the following signal

$$y_D = \frac{1}{1 + \lambda} \sqrt{P_{PSR}^{D}} u_D^\dagger \hat{H}_{D,R} v_{RS} + \tilde{n}_D$$

$$+ \sqrt{P_{PSR}^{D}} u_D^\dagger \hat{H}_{D,R} v_{RS} + \sum_{i=1}^{L} \sqrt{P_i} u_D^\dagger \hat{H}_{D,i} v_i^2 s_i. \tag{4.34}$$

Then, the corresponding SINR of $D$ from (4.34) is calculated by

$$\gamma_D = \frac{P_{PSR}^{D}}{d_{D,R}^{2}} ||u_D^\dagger \hat{H}_{D,R} v_R||^2$$

$$+ G_{PN} + \sigma_{\tilde{n}_D}^2. \tag{4.35}$$

In the case of the PN, the SINR at $PR_j$ can be as

$$\gamma_{[j]} = \frac{P_j ||u_{[j]}^\dagger \hat{H}_{[j]} v_{[j]}||^2}{d_{j,j}^{2}(1 + \lambda)^2 \left(A_{[j]} + B_{PSR}^{[j]} + \sigma_{\tilde{n}_j}^2\right)}, \tag{4.36}$$

where the interference from the SN is expressed as

$$B_{PSR}^{[j]} = \begin{cases} \frac{P_j}{d_{j,j}^{2}} ||u_{j}^\dagger \hat{H}_{j,S} v_S||^2, & \text{if } i = 1, \\ \frac{P_{PSR}}{d_{j,R}^{2}} ||u_{j}^\dagger \hat{H}_{j,R} v_R||^2, & \text{if } i = 2. \end{cases} \tag{4.37}$$

### 4.5.1 Capacity for PSR

The instantaneous capacity for $PR_j$ can be written using (4.36) as

$$C_{PSR}^{j} = \frac{1}{2} \sum_{i=1}^{2} \log_2 \left( 1 + \frac{P_j}{d_{j,j}^{2}(1 + \lambda)^2} \left| \left| u_{j}^\dagger \hat{H}_{j,j} v_{j}^i \right| \right|^2 \right). \tag{4.38}$$
Furthermore, using (4.25), (4.32) and (4.35), the instantaneous end-to-end capacity expression for $D$ is derived by

$$C_{PSR_D} = \frac{1}{2} \log_2 \left[ 1 + \min \left( \frac{P_S(1-\theta)}{d_{R,S}^2(1+\lambda)^2} \left| \begin{bmatrix} u_R^\dagger H_{R,S} v_S \end{bmatrix} \right|^2 \right) + \left( \frac{P_S(1-\theta)}{d_{R,S}^2(1+\lambda)^2} \right) \left| \begin{bmatrix} u_R^\dagger H_{R,S} v_S \end{bmatrix} \right|^2 + P_{PSR}^R \sigma_n^2 \right]^{\frac{1}{2}}.$$  \hspace{1cm} (4.39)

Similarly as in the TSR case, the average instantaneous capacity performance for the PSR protocol can be obtained numerically.

### 4.6 Outage Probability Analysis

#### 4.6.1 Outage Probability of PSR

In wireless communications, the signal is considered to be in an outage if an obtained rate at a receiver is lower than a predefined rate threshold [161]. If the predefined rate threshold of $PR_j$ is $R_{th}^j$ (bits/s/Hz), then the OP of $PR_j$, for one data stream per user and considering the perfect CSI, can be expressed by

$$P_{out} = \Pr \left[ C_{PSR}^j < R_{th}^j \right]$$  \hspace{1cm} (4.40)

Hence, within time slot $\iota$, the OP of $PR_j$ is derived by

$$P_{out(j)}^{|\iota|} = \Pr \left[ \frac{1}{2} \log_2 \left( 1 + \frac{P_j |h_{j,j}^{|\iota|}|^2}{d_{j,j}^{|\iota|} \sigma_n^j} \right) < R_{th}^j \right],$$  \hspace{1cm} (4.41)

where $h_{j,j}^{|\iota|} = u_j^{|\iota|} H_{j,j} v_j^{|\iota|}$.

As it is assumed that each user has a single data stream, then $v_j^{|\iota|}$ and $u_j^{|\iota|}$, with $\forall \iota \in \{1,2\}$, $j \in \{1,2,\ldots,L\}$, are i.n.i.d. unitary vectors which follow $CN(0,1)$ and independent from $H_{j,j}$. Thus, $h_{j,j}^{|\iota|} = u_j^{|\iota|} H_{j,j} v_j^{|\iota|}$ also follows $CN(0,1)$ [116]. Moreover, the variance of $h_{j,j}^{|\iota|}$ can be described by

$$\nabla \left( \left| h_{j,j}^{|\iota|} \right| \right) = \mathbb{E} \left( \left| h_{j,j}^{|\iota|} \right|^2 \right) = \mathbb{E} \left( \left| h_{j,j}^{|\iota|} \right|^2 \right).$$  \hspace{1cm} (4.42)

From (4.42), it is easily can be derived that $\left| h_{j,j}^{|\iota|} \right|^2$ follows an exponential distribution with $CN(1,1)$ [116].
Proposition 3 The closed-form OP expression for \( PR_j \) within time slot \( \iota \) is written as

\[
P_{\text{out}(j)}^{[\iota]} = 1 - \exp \left( -\xi_j^j \frac{d_{j,j}^{[\iota]} \sigma_{n_j}^{[\iota]} }{P_j} \right),
\]

where \( \xi_j^j = 2^{2R_j} - 1 \) is the respective SNR threshold at \( PR_j \).

Proof: See Appendix B.1 for details.

Now, regarding the SN where it has EH at \( R \), the OP expression of \( D \) is derived, using (4.39) and (4.40), as

\[
P_{\text{out}(D)} = \Pr \left[ \frac{1}{2} \log_2 \left( 1 + \min(\gamma_R, \gamma_D) \right) < R_{\text{th}}^D \right].
\]

(4.44)

Proposition 4 The closed-form OP expression for \( D \) is written as

\[
P_{\text{out}(D)} = 1 - \exp \left( -\xi_D \left( \frac{d_{R,S}^{R,S} \sigma_R^{2} }{(1 - \theta) P_S} + \frac{d_{D,R}^{D,R} \sigma_D^{2} }{P_{PSR}^D} \right) \right).
\]

(4.45)

Proof: See Appendix B.2 for details.

4.6.2 Outage Probability of TSR

Using the capacity expression for \( PR_j \) in (4.24), the OP of \( PR_j \) in the TSR protocol can be written as

\[
P_{\text{out}(j)}^{[\iota]} = \Pr \left[ J^{[\iota]} \log_2 \left( 1 + \gamma_j^{[\iota]} \right) < R_{\text{th}}^j \right] = \Pr \left[ J^{[\iota]} \log_2 \left( 1 + \frac{P_j |h_{j,j}^{[\iota]}|^2}{d_{j,j}^{[\iota]} \sigma_{n_j}^{[\iota]} } \right) < R_{\text{th}}^j \right]
\]

\[
= \Pr \left[ |h_{j,j}^{[\iota]}|^2 < \frac{\frac{\pi^2}{2} - 1}{d_{j,j}^{[\iota]} \sigma_{n_j}^{[\iota]} } \right].
\]

(4.46)

Then, by following the same steps as in Appendix B.1, the closed-form expression for the OP of \( PR_j \) in the TSR, within the time slot \( \iota \), can be expressed by

\[
P_{\text{out}(j)}^{[\iota]} = 1 - \exp \left( -\frac{\frac{\pi^2}{2} - 1}{d_{j,j}^{[\iota]} \sigma_{n_j}^{[\iota]} } \right).
\]

(4.47)
Chapter 4. Wireless Powered CRN with Interference Alignment

Further, using (4.26) and (4.40), the OP of $D$ for the TSR protocol can be executed as

$$P_{\text{out}(D)} = \Pr \left[ \frac{1 - \alpha}{2} \log_2 \left( 1 + \min(\gamma_R, \gamma_D) \right) < R_{\text{th}}^D \right].$$ (4.48)

Then, following similar steps as in Appendix B.2, the OP in (4.48) can be written in its closed-form as

$$P_{\text{out}(D)} = 1 - \exp \left( -v_{\text{th}}^D \left( \frac{d_{R,S}^R \sigma_R^2}{P_S} + \frac{d_{D,R}^D \sigma_D^2}{P_{\text{TSR}}^R} \right) \right),$$ (4.49)

where $v_{\text{th}}^D = \frac{2 \eta^2 R_{\text{th}}^D}{2 - \alpha - 1} - 1$.

4.6.3 Bit-Error Rate

The BER of the data decoded at the receiver $m$ for a symbol-by-symbol hard detection of the BPSK constellation with the zero-forcing equalization is expressed by [122, 162]

$$\text{BER}_m = \int_0^\infty Q(\sqrt{2\gamma_m} f_{\gamma_m}(\gamma_m)) d\gamma_m, \ \forall m \in \mathcal{B},$$ (4.50)

where $Q(x) = 1/\sqrt{2\pi} \int_x^\infty \exp(-t^2/2)dt$ and $f_{\gamma_m}(\gamma_m)$ denote the Gaussian-$Q$ function and the PDF of $\gamma_m$, accordingly.

4.7 Simulation Results and Discussion

In this section of the chapter, the simulation results for the proposed system model is presented. For the sake of simplicity, it is considered that the network consists of two primary user pairs and one cooperative SN. The parameters of the network are set as the next (unless it is specified otherwise). Each transmit node, except $R$, transmits with the power of $P$, whereas $R$ utilizes the harvested power for data transmission. The optimal $\alpha^*$ and $\theta^*$ values (for the case $\eta = 0.8$) for both TSR and PSR protocols over various CSI settings can be derived through differentiation, i.e., $dC_{\text{TSR}}^D/d\alpha = 0$ and $dC_{\text{PSR}}^D/d\theta = 0$. The number of antennas at each node is taken as $M = N = 4$ (due to symmetric networks scenario). Moreover, each receive node in the PN and the SN is considered to have an equal rate threshold. Furthermore, the Rayleigh block fading is assumed and the nodes are located in the equal distance from each other, i.e., $d = 3$ m [138], and the path loss exponent is taken to be $\tau = 2.7$ (for urban area environments). The CSI mismatch parameters given by $(\kappa, \psi)$ are set as $(0, 0.001), (0, 0.05), (0.75, 10), (1, 10)$ and $(1.5, 15)$. 
Fig. 4.5 illustrates the simulated results for the instantaneous capacity of the SN as a function of $\alpha$ and $\theta$ within various CSI settings at 20 dB. Eqs. (4.26) and (4.39) are taken for the capacity calculation of the TSR and PSR protocols, respectively. In the PSR protocol, when $\theta$ is too small, the capacity performance is low due to the fact that the less power of the obtained signal is devoted to the EH purpose which results in very low available transmit power at $R$. On the other hand, in the case of $\theta$ tends to 1, the capacity performance degrades because of the small power portion allotted for information decoding, whereas major fraction of power is inefficiently dedicated to the EH purpose. Similarly, the same performance pattern can be seen in $\alpha$ for the TSR-based system. From the plot it can be seen that, for the perfect CSI case, the optimal values of the TS fraction and the PS factor are considered to be as $\alpha^* = 0.2$ and $\theta^* = 0.75$, accordingly. For imperfect CSI, values of $(\alpha^*, \theta^*)$ were determined as the next: for $\kappa = 0$, $\psi = 0.001 \rightarrow (0.25, 0.84)$; for $\kappa = 0$, $\psi = 0.05 \rightarrow (0.3, 0.97)$; for $\kappa = 0.75$, $\psi = 10 \rightarrow (0.35, 0.99)$; for $\kappa = 1$, $\psi = 10 \rightarrow (0.35, 0.97)$; for $\kappa = 1.5$, $\psi = 15 \rightarrow (0.3, 0.96)$.

Fig. 4.6 reveals the result of the CSI mismatch on the capacity performance of $PR_j$ and $D$. It can be noted that the impact of $\psi$ to capacity performance is significant when $\kappa = 0$. When SNR = 20 dB and $\psi = 0.001$, the capacity performance loss of $PR_j$ is 1.82 bit/s/Hz whereas that of $D$ for the PSR and TSR-based systems is observed to be as 0.66 and 0.57 bit/s/Hz, accordingly. These capacity losses of $PR_j$ and $D$ is effected by CSI mismatch due to the higher level of interference from primary and secondary transmit nodes. Overall, the capacity losses of the TSR is smaller than that of the PSR case, nevertheless, the PSR provides more bits/s/Hz than the TSR. On the other hand,
when $\kappa \neq 0$, the capacity shows poorer performance in the low SNR region. It is worth to mention that the dependence on the SNR results in the capacity growth when the SNR increases. In addition, it is noticed that the capacity performance improves as $\kappa$ grows.

The impact of imperfect CSI parameters on the BER performance\(^3\) of the $PR_j$ and

---

\(^3\)For the sake of brevity, the BER performance of $PR_j$ only in the 1\(^{st}\) time slot is illustrated. The BER results of $PR_j$ in the 2\(^{nd}\) time slot will perform better than those of $PR_j$ in the 1\(^{st}\) time slot due to the
the SUs is shown in Fig. 4.7. When $\kappa = 0$, it is noticed that the channel error variance becomes SNR-independent and the BER performance degrades while the value of $\psi$ is increased. Moreover, the BER performance of $PR_j$ and SUs start their saturations at 23 dB and 27 dB, respectively, when $\psi = 0.001$. In addition, when $\psi = 0.05$, the BER curves of $PR_j$ and SUs start to saturate after 17 dB and 24 dB, respectively (see Fig. 4.7a). Furthermore, it is worth to mention that $\psi$ does not cause any impact on the BER performance at lower SNRs, i.e., the effect of $\psi$ on the BER performance of $PR_j$ and $D$ starts at 3 dB and 4 dB, respectively. It can be explained by the fact that, the CSI quality approaches the perfect CSI when $\psi$ tends to 0. From here, it means that the small value of $\psi$ does not cause much effect on the error performance at lower SNRs. On the other hand, when $\kappa \neq 0$, no BER performance saturation is observed as the channel error variance becomes SNR-dependent (see Fig. 4.7b). It is noticed that, when the value of $\kappa$ is increased, it leads to the better performance of the BER curves. Hence, when transmit SNR is set to be equal to 30 dB, the BER performance results of $PR_j$, $R$ and $D$ for $(\kappa = 0.75, \psi = 10)$, $(\kappa = 1, \psi = 10)$ and $(\kappa = 1.5, \psi = 15)$ are equal to $(0.0371, 0.0373$ and 0.1), $(0.0073, 0.0076$ and 0.024), $(0.00035, 0.0005$ and 0.0013), accordingly. In overall, results showed that the BER performance of $PR_j$ performs better compared to those of the SUs. This happens due to the signal’s power coefficient of $(1 - \theta)$ dedicated to data detection at $R$. In other words, the lower fraction of the allocated power for data detection at $R$ causes a higher probability of the incorrect data detection. Now, by comparing the BER performance of the SN users, it can be seen that $R$ shows outperformance over $D$. It is due to the fact that $R$ relays the decoded data to $D$ with a certain number of errors, which in turn, affects and degrades the error rate of $D$. In addition, the BER performance of $D$ additionally degrades due to mismatched CSI caused by the SN and the PNs. Finally, it can be noticed that, in imperfect CSI cases, the BER performance of $R$ shows almost the same result as that of $PR_j$ at higher SNRs.\footnote{This can be explained due to the error variance $\lambda$ of the mismatched channel. The SINRs of $R$ and $PR_j$ can be approximately expressed as $\text{SINR}_R = \frac{(1-\theta)\rho}{\rho^L(1-\theta)(1+L)+1}$ and $\text{SINR}_{PR_j}^{[1]} = \frac{\rho}{\rho^L(1+L)+1}$, respectively, where $\rho = \frac{P}{\sigma_n^2}$ indicates the nominal transmit SNR. Therefore, at higher transmit power level, the SINR of $R$ approaches the SINR of $PR_j$, i.e., $\lim_{\rho \to \infty} \Delta = 0$, where $\Delta = \left| \text{SINR}_R - \text{SINR}_{PR_j}^{[1]} \right|$.}

A deeper analysis of the effect of $\kappa$ and $\psi$ on the BER in the PSR-based system at 20 dB is illustrated in Fig. 4.8. It is clearly seen from the plot that an increase of $\kappa$ leads to the improvement of the BER performance. Contrary, an increase of the value of $\psi$ results in the degradation of the BER curves. In the first sub-figure of Fig. 4.8, it can be seen that the BER performance of various settings of $\psi$ approach 0.0019 at certain $\kappa$ values. When the second sub-figure shows how the BER performance for various values of $\kappa$ degrades
Figure 4.7. BER versus SNR for the PSR protocol under different CSI scenarios:

a) BER performance for perfect CSI and SNR-independent CSI mismatch cases ((0, 0.001) and (0, 0.005)).

b) BER performance for SNR-dependent CSI mismatch cases ((0.75, 10), (1, 10) and (1.5, 15)).
as the value of $\psi$ raises. Besides, it is noticed that, when the values of $\kappa$ are small, the BER performance degrades more sharply.

Fig. 4.9 depicts analytical values of the OP for $PR_j$ and $D$ in Section 4.6 and their simulated results. The channel settings consider perfect CSI and $d = 1$ m and $d = 3$ m at 25 dB. It can be noted from the plot that the OP degrades by raising the rate threshold. Moreover, $D$ attains worse OP compared with $PR_j$. Moreover, the simulation results verify the analytical ones by showing the same results, which proves the correctness of the derived analytical results. In Fig. 4.9a, the OP results for the PSR-based system is illustrated. As it is seen from there, the OP of $PR_j$ in the 1st and 2nd time slots show perfect match as both time slots have equal time fractions of $T/2$. Furthermore, when $d = 1$ m, $PR_j$ and $D$ approach the outage at 5.4 and 5 bits/s/Hz, accordingly. On the other hand, when $d = 3$ m, $PR_j$ is in the outage at 3.4 bits/s/Hz, whereas $D$ obtains that at 2 bits/s/Hz. From here, it can be concluded that the increase in the distance between nodes leads to the degradation of the OP due to the channel path loss effect. Moreover, the comparison between the OP of $PR_j$ and $D$ shows that the OP performance of $D$ corrupts more as the distance increases due to the fact that the effect of the distance on the OP is doubled in the two-hop communication of the SN. (see Eq. (4.45)). In Fig. 4.9b, the OP results obtained for the TSR-based system is plotted. When $d = 1$ m, during the 1st time slot, $PR_j$ obtains the outage at 5.8 bits/s/Hz, while $PR_j$ reaches the outage at 5 bits/s/Hz within the 2nd time slot. This phenomenon can be explained by the simple fact that, $PR_j$ is allocated with a time portion of $(1 + \alpha)T/2$ for the signal reception in the 1st time slot, whereas the time portion of $(1 - \alpha)T/2$ is allocated for $PR_j$ within the 2nd time slot for the same purpose (see Fig. 4.3). In addition, $D$ experiences the outage at 4.8 bits/s/Hz,
which is lower for 0.2 bits/s/Hz rate compared with the OP of the PSR-based system.

Now, using settings of $L = 5$, the impact of primary interference on the EH performance at $R$ is studied. Thus, Fig. 4.10 illustrates the harvested power at $R$ in means of various distances between $PR_j$ and $R$. The distance between $S$ and $R$ is set as $d = 3$ m,

![Graph](image)

**Figure 4.9.** The OP of the PN and the SN with different values of data rate threshold at 25 dB with the perfect CSI case: a) The OP of the system in the PSR protocol. b) The OP of the system in the TSR protocol.
while the distance between $PR_j$ to $R$ differs from 2 m to 5 m. The harvested energy is prorated with regard to the transmit power $P$. From the plot, it is noted that when $L = 5$, the harvested power is the highest. In other words, $R$ harvest more power when the number of coming interference sources is higher. Furthermore, it is also seen that the most power is harvested from $S$ and the power harvested from primary interference becomes lower when $PR_j$ is located further from $R$. It can be explained by the fact that, in longer distances, the effect of the path-loss on the signal quality is stronger. Finally, the
comparison of the EH performance for optimal TSR and optimal PSR protocols shows that $R$ harvests more power in the latter protocol than in the former one.

### 4.8 Chapter Summary

This Chapter investigated the CRN consisting of $L$ pairs of PUs and one SN, which composes $S$, the energy-restricted $R$ and $D$. The interference received at the PUs and the SN was managed using the IA technique by applying precoding and interference suppression BF matrices. Furthermore, the capacity and BER performance of the EH-based DF CRN by deploying the TSR and PSR protocols were analyzed. In addition, along with perfect CSI, five sets of imperfect CSI parameters $(\kappa, \psi)$ given by $(0, 0.001)$, $(0, 0.05)$, $(0.75, 10)$, $(1, 10)$ and $(1.5, 15)$ were considered to investigate the effect of the CSI quality on the performance of the system capacity and the BER. The simulated results showed that the system performance can be improved by increasing the values of $\kappa$, while an increase of $\psi$ degraded the system performance. Moreover, an optimal selection of the EH time fraction in the TSR protocol and the PS fraction in the PSR protocol were found to be important in achieving the best system performance. From the results, it can be concluded that the optimized PSR protocol obtained better performance over the optimized TSR scheme in terms of capacity and BER performance. In addition, it was observed that an increase in the distance between nodes results in an outage degradation. Then, the derived closed-form expressions for the OP analysis of the SN showed that the PSR protocol outperformed the TSR protocol which validity was confirmed by good agreement with simulation results. Finally, it was shown that the increase in the number of interfering primary nodes allows $R$ to harvest more power, while the increase of interfering nodes’ distance results in lower harvested power.

The contributions of this Chapter can be categorized into two main parts:

- We showed that by using the IA technique in underlay cooperative CR networks, interference at the primary receivers can be efficiently canceled as a result secondary transmit nodes are not required to limit their transmit power. Consequently, SNs can obtain better system performance in terms of the OP and the BER.

- Not similar to many works related to interference managing where unwanted interference has been canceled or removed, we used intra- and inter-network interference for EH purpose to feed the power-limited relay node.
Chapter 5

Cognitive Radio Non-Orthogonal Multiple Access Networks

NOMA is considered to be one of the promising multiple access techniques for 5G networks. This Chapter studies a dual-hop underlay CR-NOMA networks. Closed-form expressions for the OP of NOMA secondary destination users are evaluated for DF and AF relaying techniques. Furthermore, power allocation factors for different distances of secondary NOMA users are found to satisfy OP fairness for all users. Moreover, the OP for NOMA is compared with that for conventional orthogonal MA (OMA) to show the supremacy of the former MA technique. All derived analytical expressions are validated with Monte Carlo simulations.

5.1 Introduction

5.1.1 Cooperative NOMA

In NOMA, stronger users, by having preliminary information about messages of the other users, can act as relays. Hence, a cooperative NOMA was presented in [61], where users with stronger channel gains were assigned relay roles to forward the signals to the users with weaker channel gains. The results showed that the cooperative NOMA outperforms both the conventional cooperative OMA and non-cooperative NOMA. Moreover, the cooperative NOMA with a dedicated relay was studied in [163, 164]. Further, in [163], the OP results showed that the cooperative NOMA outperforms the cooperative OMA regarding a coding gain while, in [164], a NOMA scheme with a two-way relay that supports both uplink and downlink was proposed, where the results showed that the proposed system achieves better sum rate compared to the OMA scheme. Furthermore, in [165], the PA scheme in cooperative NOMA systems for fixed and adaptive relaying cases was studied. In addition, the benefits of cooperative NOMA networks were widely studied in various network scenarios [166, 167].
It is already known that due to the distinct characteristic of the underlay CR paradigm, the SN transmissions should not interfere with the PN communication. This fact makes the power control at secondary transmit nodes more critical and due to that traditional underlay CR is often limited to short-range communications [168]. To address these challenges, a cooperative underlay NOMA architecture has been proposed [84, 87, 169], where a relay helps to improve the radio coverage and minimize an outage for underlay NOMA networks.

5.2 Contribution of the Chapter

The main contribution of the Chapter can be concluded as

- The synergy of cooperative underlay CR and NOMA is studied in terms of the outage performance of NOMA users applying both DF and AF relaying methods over Rayleigh and Nakagami-$m$ fading channels.

- The power allocation factors are found for NOMA secondary users in order to guarantee outage fairness (OF) among NOMA users.

- Considering the effect of imperfect CSI on the outage performance, we derived a general closed-form solution for the OP when the number of secondary NOMA nodes is extended to $K$ users.

- The comparison between the proposed cooperative CR-NOMA and the conventional cooperative CR-OMA shows the supremacy of the former.

5.3 Organization of the Chapter

This Chapter studies a downlink dual-hop CR-NOMA model with the imposed ITC at $PR$. The chapter considers several different scenarios for the proposed system model. In Section 5.4, we propose a system model where only one secondary node is limited (Case 1) by maximum transmit power. The OP is investigated as a performance metric for the considered system model applying DF and AF relaying methods over Rayleigh and Nakagami-$m$ fading channels. Here, the closed-form OP expression is derived for two NOMA secondary users. Then, in Section 5.5, the scenario where $PR$ applies the ITC to both $S$ and $R$ (Case 2) is considered. By considering the effect of the imperfect CSI on the outage performance, we derive a general closed-form solution for the OP when the number of secondary NOMA nodes is extended to $K$ users. All derived analytical results are verified by Monte-Carlo simulations. Furthermore, in order to obtain OF among NOMA users, the power allocation factors are found depending on the distance
between the relay and secondary NOMA users. Finally, the obtained OP performance for NOMA is compared to that for conventional cooperative OMA to show the supremacy of the former.

5.4 Outage analysis of dual-hop CR-NOMA: Case 1

In this part, the outage performance will be evaluated for CR-NOMA by assuming just only one secondary transmit node is constrained from the maximum allowable transmit power due to its close location to PR.

5.4.1 CR-NOMA with two secondary NOMA users

5.4.1.1 System Model

Consider the scenario with a downlink dual-hop underlay CR-NOMA network shown in Fig. 5.1. The network consists of $S$, $R$, two destination users ($D_1$ and $D_2$) and PR. It is assumed that the direct $S$-to-$D_1$ and $S$-to-$D_2$ links do not exist. Another assumption is that all nodes are equipped with a single antenna and operate in a half-duplex mode. Fading-channel gains, i.e., $h_{SP}$, $h_{SR}$, $h_1$ and $h_2$, follow the i.n.i.d. Rayleigh block fading model. Moreover, it is assumed that perfect CSI is available at each node. Furthermore, the corresponding distances between nodes are expressed by $d_{SP}$, $d_{SR}$, $d_1$ and $d_2$. According to the system model, $S$ causes interference to $PR$, whereas $R$, which operates in DF and AF modes, cannot interfere with the PN due to its remote location from $PR$ (see Fig. 5.1). Another assumption is that the SN can communicate only if $PR$ does not receive harmful interference from $S$. Hence, $S$ is restricted by transmit power as \[ P_S \leq \min \left( \frac{I_d}{|h_{SP}|^2}, \bar{P}_S \right), \] (5.1)
where $\tau$ is the path-loss exponent, $P_S$ and $I$ indicate maximum average transmit power available at $S$ and the ITC at $PR$, accordingly. By considering this power restriction, $S$ transmits the next superimposed signal to $D_1$ and $D_2$ through the aid of $R$ in two-time slots

$$x_S = \sqrt{\alpha_1 P_S} x_1 + \sqrt{\alpha_2 P_S} x_2,$$

(5.2)

where $x_1$ and $x_2$, with $E(|x_i|^2) = 1$, $i \in \{1, 2\}$, denotes the intended data to $D_1$ and $D_2$, respectively, $\alpha_1$ and $\alpha_2$ stand for the PA factors for corresponding users, with $\alpha_1 + \alpha_2 = 1$. Further, during the 1st time slot, the signal received at $R$ can be written as

$$y_R = h_{SR} \left( \sqrt{\frac{\alpha_1 P_S}{d_{SR}}} x_1 + \sqrt{\frac{\alpha_2 P_S}{d_{SR}}} x_2 \right) + n_R,$$

(5.3)

where $n(\cdot) \sim \mathcal{CN}(0, \sigma_R^2)$ denotes the AWGN at each receiving node$^1$. Moreover, channels and the PA factors are ordered as $h_1 < h_2$ and $\alpha_1 > \alpha_2$, which means that $D_1$ is allocated with the higher power portion due to its channel weaknesses.

### 5.4.1.2 Decode-and-Forward Relaying Scheme

Here, it is considered the case when $R$ employed the DF relaying scheme. Saying so, in the 2nd time slot, $R$ decodes and removes $x_1$ from the received signal in (5.3) before detecting $x_2$, according to the principle of NOMA [61]. Thus, at $R$, the SINR of decoding $x_1$ is derived as

$$\gamma_{R,1} = \frac{|h_{SR}|^2 \alpha_1 P_S}{|h_{SR}|^2 \alpha_2 P_S + \sigma^2} = \frac{|h_{SR}|^2 \alpha_1 \rho_S}{|h_{SR}|^2 \alpha_2 \rho_S + d_{SR}},$$

(5.4)

where $\rho_S = \frac{P_S}{\sigma^2}$ is the source transmit SNR. Further, on the condition$^2$ that $\gamma_{R,1} \geq \xi_1$, where $\xi(\cdot)$ denotes the predefined SNR threshold at $D(\cdot)$, the data detected in (5.4) is removed from the remaining signal and the SNR to detect $x_2$ at $R$ can be written as

$$\gamma_{R,2} = \frac{|h_{SR}|^2 \alpha_2 \rho_S}{d_{SR}}.$$

(5.5)

Further, $R$ relays the detected superimposed signal $x_R$ to $D_1$ and $D_2$

$$x_R = \sqrt{\beta_1 P_R} \tilde{x}_1 + \sqrt{\beta_2 P_R} \tilde{x}_2,$$

(5.6)

$^1$For mathematical tractability and without loss of generality, it is assumed that the noise power at each receiving node is equal to $\sigma^2$, i.e., $\sigma^2_P = \sigma^2_R = \sigma^2_1 = \sigma^2_2 = \sigma^2$ throughout the paper.

$^2$This condition provides succesfull implementation of the SIC at $R$ to remove $x_1$. 
where $P_R$ stands for the relay transmit power, $\tilde{x}_1$ and $\tilde{x}_2$ indicate decoded and forwarded data to corresponding receivers whereas $\beta_1$ and $\beta_2$, with $\beta_1 + \beta_2 = 1$, satisfying $\beta_1 > \beta_2$, represent the PA factors at $R$. Hence, the received signal at $D_1$ and $D_2$ can be respectively written as

$$y_1 = h_1 \left( \sqrt{\frac{\beta_1 P_R}{d_1^2}} \tilde{x}_1 + \sqrt{\frac{\beta_2 P_R}{d_1^2}} \tilde{x}_2 \right) + n_1 \tag{5.7}$$

and

$$y_2 = h_2 \left( \sqrt{\frac{\beta_1 P_R}{d_2^2}} \tilde{x}_1 + \sqrt{\frac{\beta_2 P_R}{d_2^2}} \tilde{x}_2 \right) + n_2, \tag{5.8}$$

where $n_1$ and $n_2$ denote the AWGN noise terms at $D_1$ and $D_2$, accordingly. Moreover, $D_2$ applies the SIC by detecting $\tilde{x}_1$ at the same time considering its own data $\tilde{x}_2$ as a background noise. Thus, the SINR of detecting $x_1$ at $D_2$ can be formulated as follows

$$\gamma_{2,1} = \frac{|h_2|^2 \beta_1 \rho_R}{|h_2|^2 \beta_2 \rho_R + d_2^2}, \tag{5.9}$$

where $\rho_R = \frac{P_R}{\sigma^2}$ stands for the transmit SNR at $R$. Further, on the condition that $\gamma_{2,1} \geq \xi_1$, i.e., the SIC is successfully implemented at $D_2$ to remove $x_1$, the data detected in (5.9) is removed from the remaining signal and the SNR to detect $x_2$ at $D_2$ can be written as

$$\gamma_{2,2} = \frac{|h_2|^2 \beta_2 \rho_R}{d_2^2}. \tag{5.10}$$

Due to the channel ordering, $D_1$ is allocated with higher power fraction. Thus, $D_1$ detects $\tilde{x}_1$ by treating $\tilde{x}_2$ as a background noise with the next SINR

$$\gamma_1 = \frac{|h_1|^2 \beta_1 \rho_R}{|h_1|^2 \beta_2 \rho_R + d_1^2}. \tag{5.11}$$

Then, the achievable rates of $x_1$ and $x_2$ can be written as [170]

$$R_1 = \frac{1}{2} \log_2 \left( 1 + \min (\gamma_{R,1}, \gamma_{2,1}, \gamma_1) \right) \tag{5.12}$$

and

$$R_2 = \frac{1}{2} \log_2 \left( 1 + \min (\gamma_{R,2}, \gamma_{2,2}) \right), \tag{5.13}$$

accordingly, where all SNR values involved in each min function are independent RVs.
5.4.1.3 Amplify-and-Forward Relaying Scheme

Not similar to the DF relaying, where \( R \) detects and forwards the signal from \( S \) to secondary destinations, in the AF protocol, \( R \) amplifies the signal received from \( S \) and forwards to both secondary destinations. Hence, \( D_1 \) and \( D_2 \) respectively receive the following signals

\[
y^{AF}_1 = \sqrt{\frac{P_R}{d_1^2}} h_1 G y_R + n_1 = \sqrt{\frac{P_S P_R}{d_{SR} d_1^2}} h_{SR} h_1 \sum_{i=1}^{2} \sqrt{\alpha_i} x_i G + \sqrt{\frac{P_R}{d_1^2}} h_1 n_R G + n_1 \tag{5.14}
\]

and

\[
y^{AF}_2 = \sqrt{\frac{P_R}{d_2^2}} h_2 G y_R + n_2 = \sqrt{\frac{P_S P_R}{d_{SR} d_2^2}} h_{SR} h_2 \sum_{i=1}^{2} \sqrt{\alpha_i} x_i G + \sqrt{\frac{P_R}{d_2^2}} h_2 n_R G + n_2, \tag{5.15}
\]

where \( G \) denotes the amplification factor at \( R \) which can be formulated as

\[
G = \sqrt{\frac{1}{\frac{P_S}{d_{SR}} |h_{SR}|^2 + \sigma^2}}. \tag{5.16}
\]

Then, due to higher value of \( \alpha_1 \), \( D_1 \) can treat \( x_2 \) as a background noise and detect \( x_1 \) by the following SINR

\[
\gamma^{AF}_1 = \frac{\frac{P_S}{d_{SR}} |h_{SR}|^2 \frac{P_R}{d_1^2} |h_1|^2 G^2 \alpha_1}{\frac{P_S}{d_{SR}} |h_{SR}|^2 \frac{P_R}{d_1^2} |h_1|^2 G^2 \alpha_2 + \frac{P_R}{d_1^2} |h_1|^2 G^2 \sigma^2 + \sigma^2}
= \frac{\frac{P_S}{d_{SR}} |h_{SR}|^2 \frac{P_R}{d_1^2} |h_1|^2 \alpha_1}{\frac{P_S}{d_{SR}} |h_{SR}|^2 \frac{P_R}{d_1^2} |h_1|^2 \alpha_2 + \frac{P_S}{d_{SR}} |h_{SR}|^2 + G^2}
= \frac{n_s Y n_1 X \alpha_1}{n_s Y n_1 X \alpha_2 + n_1 X + n_s Y + 1}, \tag{5.17}
\]

where \( n_s = \frac{P_S}{d_{SR}}, \ n_1 = \frac{P_R}{d_1^2}, \ \alpha_1 = |h_1|^2 \) and \( Y = |h_{SR}|^2 \). Then, according to the principle of NOMA, \( D_2 \) detects \( x_1 \) by treating its own signal as a background noise, which SINR can be derived as

\[
\gamma^{AF}_{2,1} = \frac{n_s Y n_2 Q \alpha_1}{n_s Y n_2 Q \alpha_2 + n_2 Q + n_s Y + 1}, \tag{5.18}
\]

where \( n_2 = \frac{P_R}{d_2^2} \) and \( Q = |h_2|^2 \) is an exponential distribution. Further, on the condition that \( \gamma_{2,1} \geq \xi_1 \), the detected \( x_1 \) in (5.18) is removed and \( D_2 \) detects its own data from the remaining signal, which SNR can be written as

\[
\gamma^{AF}_2 = \frac{n_s Y n_2 Q \alpha_2}{n_2 Q + n_s Y + 1}. \tag{5.19}
\]
Furthermore, the achievable rates\(^3\) of \(D_1\) and \(D_2\) can be respectively derived by

\[
R_{AF}^{1} = \frac{1}{2} \log_2 \left(1 + \gamma_{AF}^1\right),
\]

(5.20)

\[
R_{AF}^{2} = \frac{1}{2} \log_2 \left(1 + \gamma_{AF}^2\right).
\]

(5.21)

### 5.4.1.4 Outage Performance Analysis

#### 5.4.1.5 Outage for the Decode-and-Forward Relaying Mode

The OP of \(x_1\) for DF mode can be formulated as

\[
P_{out,1}(\xi_1) = 1 - \Pr \left[\min \left(\gamma_{R,1}, \gamma_{2,1}, \gamma_1\right) > \xi_1\right] = 1 - \Pr \left[\gamma_{R,1} > \xi_1\right] \Pr \left[\gamma_{2,1} > \xi_1\right] \Pr \left[\gamma_1 > \xi_1\right] = F_{\gamma_{2,1}}(\xi_1) + (1 - F_{\gamma_{2,1}}(\xi_1)) \left(F_{\gamma_1}(\xi_1) + F_{\gamma_{R,1}}(\xi_1) - F_{\gamma_1}(\xi_1)F_{\gamma_{R,1}}(\xi_1)\right).
\]

(5.22)

**Proposition 5** The closed-form OP expression for \(x_1\) is written as

\[
P_{out,1}(\xi_1) = 1 - e^{-\xi_1 \frac{\bar{\rho}_S \alpha^2_1 + \rho_R \beta^2_1}{2 \rho_S \rho_R}} + \frac{\xi_1 \bar{d}_{SR}^2 e^{-\frac{\rho_P (\rho_I \alpha^2_2 + \rho_R \beta^2_2 \xi_1 + \rho_S \alpha^2_1)}{2 \rho_S \rho_R}}}{\bar{\rho}_S \alpha^2_2 \rho_R + \xi_1 \bar{d}_{SR}^2},
\]

(5.23)

where, \(\xi_1 < \frac{\alpha_1}{\alpha_2}\) and \(\xi_1 < \frac{\beta_1}{\beta_2}\), otherwise, \(P_{out,1}(\xi_1) = 1\).

**Proof:** See Appendix C.1 for details.

The OP of the signal \(x_2\) can be written as

\[
P_{out,2}(\xi_2) = 1 - \Pr \left[\min \left(\gamma_{R,2}, \gamma_{2,2}\right) > \xi_2\right] = 1 - \Pr \left[\gamma_{R,2} > \xi_2\right] \Pr \left[\gamma_{2,2} > \xi_2\right] = F_{\gamma_{2,2}}(\xi_2) + F_{\gamma_{R,2}}(\xi_2) - F_{\gamma_{2,2}}(\xi_2)F_{\gamma_{R,2}}(\xi_2).
\]

(5.24)

**Proposition 6** The closed-form OP expression for \(x_2\) is written as

\[
P_{out,2}(\xi_2) = 1 - e^{-\xi_2 \frac{\bar{\rho}_S \alpha^2_1 + \rho_R \beta^2_1}{2 \rho_S \rho_R}} + \frac{\xi_2 \bar{d}_{SR}^2 e^{-\frac{\rho_P (\rho_I \alpha^2_2 + \rho_R \beta^2_2 \xi_2 + \rho_S \alpha^2_1)}{2 \rho_S \rho_R}}}{\alpha_2 \rho_I \bar{d}_{SP} + \xi_2 \bar{d}_{SR}^2}.
\]

(5.25)

**Proof:** See Appendix C.2 for details.

---

\(^3\)In the DF mode, we considered the achievable rate of the messages in order to derive the OP. However, in the AF mode, RVs \(\gamma_{AF}^1\) and \(\gamma_{AF}^2\) are correlated to each other, which makes the derivation of a closed-form solution for the OP of \(x_1\) very complex or even intractable. Thus, we concentrate on the OP derivation of \(D_1\) and \(D_2\) for the AF scenario.
5.4.1.6 Outage for the Amplify-and-Forward Relaying Mode

Not similar to the DF relaying, the intended data in the AF relaying is detected only at the final destination, i.e., $D_1$ and $D_2$. Therefore, the OP for $D_1$ can be expressed as

$$P_{\text{out},1}^{\text{AF}}(\xi_1) = 1 - \text{Pr} \left[ \gamma_1^{\text{AF}} > \xi_1 \right] = 1 - (1 - F_{\gamma_1}^{\text{AF}}(\xi_1)) = \Delta F_{\gamma_1}^{\text{AF}}(\xi_1).$$ (5.26)

**Proposition 7** The exact OP expression of $D_1$ can be written as

$$P_{\text{out},1}^{\text{AF}}(\xi_1) = 1 - e^{-\frac{u_1(\bar{u}_1+1)}{n_1}} \sqrt{\frac{2}{r_1 K_1}} \left( 2 \sqrt{r_1} \right) \left( 1 - e^{-c} \right) - \frac{n_1 e^{-c \bar{u}_1}}{\bar{u}_1 (n_p \bar{u}_1 + 1)}$$

$$\times \sum_{k=0}^{\infty} \left( -\frac{c \bar{u}_1 (n_p \bar{u}_1 + 1)}{n_1} \right)^k \frac{k!}{G_{13}^{13}} \begin{pmatrix} n_1(\bar{u}_1 + 1) \left| 0, k-1, k \right. \\ \bar{u}_1 (n_p \bar{u}_1 + 1) \end{pmatrix},$$ (5.27)

where $\xi_1 < \frac{\alpha_1}{\alpha_2}$, otherwise, $P_1(\xi_1) \sim 1$.

**Proof:** See Appendix C.3 for details. ■

Furthermore, the OP of $D_2$ can be expressed by

$$P_{\text{out},2}^{\text{AF}}(\xi_2) = 1 - \text{Pr} \left[ \gamma_2^{\text{AF}} > \xi_2 \right] = 1 - (1 - F_{\gamma_2}^{\text{AF}}(\xi_2)) = \Delta F_{\gamma_2}^{\text{AF}}(\xi_2).$$ (5.28)

**Proposition 8** The exact OP of $D_2$ can be written as

$$P_{\text{out},2}^{\text{AF}}(\xi_2) = 1 - e^{-\frac{u_2(\bar{u}_2+1)}{n_2}} \sqrt{\frac{2}{r_2 K_1}} \left( 2 \sqrt{r_2} \right) \left( 1 - e^{-c} \right) - \frac{n_2 e^{-c \bar{u}_2}}{\bar{u}_2 (n_p \bar{u}_2 + 1)}$$

$$\times \sum_{k=0}^{\infty} \left( -\frac{c \bar{u}_2 (n_p \bar{u}_2 + 1)}{n_2} \right)^k \frac{k!}{G_{13}^{13}} \begin{pmatrix} n_2(\bar{u}_2 + 1) \left| 0, k-1, k \right. \\ \bar{u}_2 (n_p \bar{u}_2 + 1) \end{pmatrix},$$ (5.29)

where $r_2 = \frac{u_2(n_p \bar{u}_2 + 1)}{n_2}$.

**Proof:** See Appendix C.4 for details. ■

5.4.1.7 Numerical Results

This part of the Chapter represents numerical and simulation results for the OP over Rayleigh fading channels. The adopted system parameters for numerical and simulation results are taken as (unless otherwise stated): $d_1 = \{1.5d, 2d, 3d, 4d, 5d\}$, $d_{SP} =$

\[d \text{ is assumed to be a unity in order to focus on OP results.}\]
Chapter 5. Cognitive Radio Non-Orthogonal Multiple Access Networks

Figure 5.2. The OP of $x_1$ and $x_2$ versus SNR performance when $d_1 = \{1.5d, 2d, 3d, 5d\}$, $\alpha_1 = 0.8$ and $\alpha_2 = 0.2$.

$d_{SR} = d_2 = d$, $\tau = 2.7$, $I = 20$ dB, $\alpha_1 = \beta_2 = 0.8$, $\alpha_2 = \beta_2 = 0.2$, and $\xi = \xi_1 = \xi_2 = 3$ dB.

**Numerical Results for the Decode-and-Forward Relaying mode**

In Fig. 5.2, the OP results for $x_1$ and $x_2$ obtained using (5.23) and (5.25), accordingly, are illustrated. One observation is that the outage of $x_2$ shows better performance than that of $x_1$ at lower SNR values (until 20 dB). The latter is due to the fact that $D_2$ cancels interference from $D_1$ by applying the SIC, while $D_1$ detects the desired signal without canceling interference from $D_2$ which increases the probability of error in data detection. Another observation is that the OP $x_1$ in different $d_1$ cases outperforms that of $x_2$ at higher SNR. This can be explained as follows: at higher SNR values, $D_1$ receives a stronger signal to detect the desired data which, in turn, provides less possibility for an outage. To compare the OP performance of $x_1$ at different $d_1$, it is seen that the increase of $d_1$ leads to the rise of the outage curve. Moreover, the simulation results match with the analytical ones, which proves the correctness of the derived closed-form OP expressions of $x_1$ and $x_2$. 
Figure 5.3. The OP of $x_1$ and $x_2$ versus the SNR threshold at 20 dB when $d_1 = \{1.5d, 2d, 3d, 5d\}$, $\alpha_1 = 0.8$ and $\alpha_2 = 0.2$.

Figure 5.4. The OP of $x_1$ and $x_2$ versus PA factors at 20 dB when $d_1 = \{1.5d, 2d, 3d, 5d\}$.

The analytical values of the OP for $x_1$ and $x_2$ with various $d_1$ distances and SNR threshold values are shown in Fig. 5.3. It is observed that curves of the OP raise by increasing the SNR threshold. When $d_1 \geq 2d$, $x_1$ shows worse outage performance compared to $x_2$. When $d_1 = 1.5d$ and $\xi < 3$ dB, the OP of $x_1$ achieves better performance.
Table 5.1. OF-based PA factors for different $d_1$ for the DF mode.

<table>
<thead>
<tr>
<th>$d_1$</th>
<th>1.5d</th>
<th>2d</th>
<th>3d</th>
<th>5d</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\alpha_1^<em>, \beta_1^</em>$</td>
<td>0.8</td>
<td>0.83</td>
<td>0.91</td>
<td>0.97</td>
</tr>
<tr>
<td>$\alpha_2^<em>, \beta_2^</em>$</td>
<td>0.2</td>
<td>0.17</td>
<td>0.09</td>
<td>0.03</td>
</tr>
</tbody>
</table>

Figure 5.5. The OP of $x_1$ and $x_2$ versus the transmit SNR performance with OF-based PA factors when $d_1 = \{2d, 3d, 5d\}$.

than that of $x_2$. It is due to the fact that higher value of power allocated to $D_1$ compare to $D_2$. However, when $\xi > 3$ dB, the OP curve of $x_1$ dramatically increases reaching an outage at 6 dB.

The OF-based PA factors $\alpha_1^*$ and $\beta_1^*$ at 20 dB is shown in Fig. 5.4. It is clearly seen that, when $\xi > \frac{\alpha_1}{\alpha_2}$, signal $x_1$ is in outage for all values of $d_1$. However, when $\xi < \frac{\alpha_1}{\alpha_2}$, the increase of $\alpha_1$ leads to an improvement of the outage performance of $x_1$ at the same time decreasing that of $x_2$. The observed OF-based PA factors for different $d_1$ values are given in Table 5.1.

Fig. 5.5 illustrates the outage performance results with OF-based $\alpha_1^*, \beta_1^*$ and $\alpha_2^*, \beta_2^*$ values in the case of different $d_1$. It is observed that, in all $d_1$ cases, the outage performance of $x_2$ outperforms that of $x_1$ at values of SNR lower than 20 dB. At other SNR values, the OP of $x_1$ is stronger than that of $x_2$. Moreover, when $d_1$ is increased, the saturation of outage curve of $x_1$ starts at higher SNRs due to the fact that larger $\alpha_1$ is needed when $D_1$ is located far away from $R$. 
Fig. 5.6 shows the results of the OP versus the SNR threshold with OF-based PA factors for \( d_1 \). It is observed that the OF-based PA factors improved the OP performance of \( x_1 \) in all \( d_1 \) cases providing fairness for both users.
The comparison of the OP for cooperative NOMA to that for cooperative OMA is illustrated in Fig. 5.7. We consider two cases, with no ITC, e.g., $I = \infty$ and when $I = 20$ dB. In the proposed system model, the cooperative NOMA requires two-time slots to deliver the intended messages to $D_1$ and $D_2$, whereas four time periods are needed for cooperative OMA to serve two destination users. Hence, the data requirements for cooperative OMA is set as 2-fold of that for cooperative NOMA for a fair comparison of two MA techniques. Moreover, it is assumed that transmit power levels at $S$ and $R$ are equal and normalized to $P$, e.g., $P_S = P_R = P$. Therefore, in the OMA case, the power allocated for each data transmission at $S$ and $R$ is equal to $\frac{1}{2}P$. The other parameters of the OMA system are identical with those of the NOMA system. As it is seen from Fig. 5.7, both messages of cooperative NOMA outperform the OMA ones for all SNR regimes when no ITC implemented. When $I = 25$ dB, the OP curves for NOMA messages start to saturate at lower SNR regimes compared to those for the OMA ones. This phenomenon can be explained by using the simple fact that in the OMA case, the transmitted power $\frac{1}{2}P$ increases the ITC at $PR$ regarding (5.1).

**Numerical Results for the Amplify-and-Forward Relaying mode**

The OP results for $D_1$ and $D_2$ of AF CR-NOMA are shown in Fig. 5.8. It is observed that the OP for $D_2$ performs better than that for $D_1$ in all $d_1$ distances. It is due to the fact that $D_2$ removes the unwanted message of $D_1$ by applying the SIC technique which improves the probability of errors in the data detection. On the other hand, the OP saturation for $D_2$ starts at lower SNRs compared with that for $D_2$. Moreover, it is noticed that the OP of $D_1$ degrades as $d_1$ increases. However, a saturation of the outage curve of $D_1$ starts at higher SNRs when the value of $d_1$ increases. This phenomenon can be explained by the fact that higher $\alpha_1$ is required when $D_1$ is located farther from $R$. Moreover, Fig. 5.8 also compares the OP results derived for NOMA with simulated results for OMA in the cases without ITC, i.e., $P_I = \infty$, and when $P_I = 20$ dB. As it is seen in no ITC regime, both NOMA users outperform the OMA users in terms of the OP for all SNR values. However, when the ITC is applied ($P_I = 20$ dB), OMA users’ OP curves start their saturation at higher SNR levels compared to those for the NOMA users.

In Fig. 5.9, the OP is plotted versus the transmit SNR for $D_1$ and $D_2$ when $d_1 = 3d$. We compare the OP of the proposed AF mode with that of the DF mode. Moreover, we plot high SNR approximation using $e^{-t} = 1 - t$ and $K_1(\omega) \sim 1/\omega$ for both AF and DF modes. It can be observed from the plot that the DF relaying performs better than the AF one. Furthermore, the OP of the high SNR approximation of the AF mode is consistent with that of the DF mode, which means that the maximum optimal OP of the AF cannot exceed the outage performance of the DF mode. In addition, Monte Carlo simulations results are consistent with the derived analytical results, which confirm the accuracy of
Chapter 5. Cognitive Radio Non-Orthogonal Multiple Access Networks

Figure 5.8. The OP of $D_1$ and $D_2$ versus the transmit SNR when $d_1 = \{2d, 4d\}$, $\alpha_1 = 0.8$, $\alpha_2 = 0.2$ and $\tau = 0$.

Figure 5.9. The OP of $D_1$ and $D_2$ versus the transmit SNR when $d_1 = 3d$, $\alpha_1 = 0.8$, $\alpha_2 = 0.2$ and $\tau = 0$. 
Table 5.2. The observed OF-based PA factors for the AF mode.

<table>
<thead>
<tr>
<th>(d)</th>
<th>(1.5d)</th>
<th>(2d)</th>
<th>(4d)</th>
</tr>
</thead>
<tbody>
<tr>
<td>(\alpha_1)</td>
<td>0.79</td>
<td>0.84</td>
<td>0.945</td>
</tr>
<tr>
<td>(\alpha_2)</td>
<td>0.21</td>
<td>0.16</td>
<td>0.055</td>
</tr>
</tbody>
</table>

Figure 5.10. The OP of \(D_1\) and \(D_2\) versus the PA factors at 20 dB when \(d_1 = \{1.5d, 2d, 4d\}\) and \(\tau = 0\).

the derived exact and approximated OP expressions.

The numerical results on the OF-based PA factors for \(D_1\) and \(D_2\) are shown in Fig. 5.10. Hence, all observed values of OF-based \(\alpha_1^*\) and \(\alpha_2^*\) for different \(d_1\) distances are illustrated in Table 5.2.

Fig. 5.11 shows the OP versus the transmit SNR with OF-based \(\alpha_1^*\) and \(\alpha_2^*\) regarding different \(d_1\). It is observed from the figure that, for all \(d_1\) values, \(D_2\) shows better outage performance compared to \(D_1\) at lower SNRs \((< 22)\) dB. However, at higher SNR levels, \(D_1\) is superior to \(D_2\) in terms of the outage performance. Furthermore, when the value of \(d_1\) rises, the OP saturation of \(D_1\) commences at higher SNR levels. This phenomenon can be explained by using the simple fact that \(D_1\) requires a larger value of \(\alpha_1\) when it is located farther from \(R\). Finally, at about 22 dB, the OP performance of both users are the same, which confirms that OF-based PA factors provide OP fairness for users.

Fig. 5.12 illustrates the results on the OP versus the predefined SNR threshold considering OF-based \(\alpha_1^*\) and \(\alpha_2^*\) for various \(d_1\) at 20 dB transmit SNR. It is worth mentioning that the OF-based PA factors improve the outage performance of \(D_1\) (for all \(d_1\) values)
Figure 5.11. The OP of $D_1$ and $D_2$ versus the transmit SNR with OF-based PA factors when $d_1 = \{1.5d, 2d, 4d\}$ and $\tau = 0$.

Figure 5.12. The OP of $D_1$ and $D_2$ versus the predefined SNR threshold with OF-based PA factors at 20 dB when $d_1 = \{1.5d, 2d, 4d\}$ and $\tau = 0$. 
and provide fairness among NOMA users, especially, at the predefined SNR threshold of 3 dB.

5.4.2 CR-NOMA over Nakagami-\(m\) distribution

In the wireless communication, the received signal strength or power fluctuate due to the fading effect. It occurs due to multiple copies of the same transmitted signal reaching the receiver over various paths. Accordingly, the received signal power becomes RV, which can be described through different distributions, e.g., Rayleigh, Rician, Nakagami-\(m\), etc. The Nakagami-\(m\) distribution has been broadly used for theoretical modeling of the channel fading, particularly in the wireless environment after the legendary work by Nakagami was published in [171]. The Nakagami-\(m\) is an empirical model covering a wide range of multipath fading models through the introduction of the parameter \(m\). Thus, this channel model can give more precise results for representing the channel when it is used to model the communication systems [172, 173].

In this part, a downlink dual-hop underlay CR-NOMA network is studied for the Nakagami-\(m\) distribution. The considered system model is the same as in the previous sub-chapter despite the transmit power constraint is applied at the DF based \(R\) instead of \(S\) (see Fig. 5.13). Thus, it is assumed that \(S\) does not interfere with \(PR\) due its sufficient remoteness. Moreover, channel coefficients, i.e., \(h_t, \forall t \in \{SR, RP, 1, 2\}\), experience quasi-static non-independent and identically distributed Nakagami-\(m\) fading which PDF can be written as

\[
f_{h_t}(r) = \frac{2\lambda_t^{m_t}}{\Gamma(m_t)} r^{2m_t-1} \exp \left( -\lambda_t r^2 \right), \quad \lambda_t = \frac{m_t}{\bar{r}_t},
\]

(5.30)

It worth to mention that the Nakagami-\(m\) fading can describe different multipath channels by using the fading parameter \(m\). For example, the one-sided Gaussian fading and the Rayleigh fading can be described by setting \(m = \frac{1}{2}\) and \(m = 1\), respectively. Additionally, the Nakagami-\(m\) fading frequently used to describe the indoor [174] and urban [175] multipath environments. According to the proposed system model, \(R\) is restricted from maximum transmit power as the secondary transmission is only permitted if \(PR\) receive a tolerable level of interference caused by the SN. Therefore, the transmit power restriction at \(R\) can be written as [161]

\[
P_R \leq \min \left( \frac{I_{d_{RP}}^R}{|h_{RP}|^2}, P_R \right),
\]

(5.31)
where \( \bar{P}_R \) is the maximum allowed transmit power at \( R \). Then, \( S \) broadcasts the superimposed signal to \( R \), which can be written as followings

\[
y_R = \sqrt{\frac{P_S}{d_{SR}}} h_{SR} (\sqrt{a_1 x_1} + \sqrt{a_2 x_2}) + n_R. \tag{5.32}
\]

In this system model, the power allocation factors are assumed to be as \( a_1 < a_2 \) and channel ordering as \( h_1 > h_2 \), which means that the channel of \( D_1 \) is stronger than that of \( D_2 \). Hence, \( D_2 \) is allocated with higher power portion. After \( R \) receives the signal from \( S \), it implements the SIC scheme first to decode \( x_2 \) with the SINR of

\[
\gamma_{R,2} = \frac{|h_{SR}|^2 a_2 P_S}{|h_{SR}|^2 a_1 P_S + \sigma_R^2 d_{SR}^R}. \tag{5.33}
\]

Further, on the condition that \( \gamma_{R,2} \geq \xi_2 \), i.e., the SIC is successfully implemented at \( R \) to remove \( x_2 \). Then, the data detected in (5.4) is removed from the remaining signal and the SNR to detect \( x_1 \) at \( R \) can be written as

\[
\gamma_{R,1} = \frac{|h_{SR}|^2 a_1 P_S}{\sigma_R^2 d_{SR}^R}. \tag{5.34}
\]

Furthermore, \( R \) forwards the detected superimposed signal of \( x_R = \sum_{i=1}^2 \sqrt{b_i} P_R \tilde{x}_i \) to both secondary users, where \( b_1 < b_2 \). Thus, in the 2\textsuperscript{nd} time slot, the received signals at \( D_1 \) and \( D_2 \) are respectively expressed by followings

\[
y_i = \sqrt{\frac{P_R}{d_i^R}} h_1 (\sqrt{b_1} \tilde{x}_1 + \sqrt{b_2} \tilde{x}_2) + n_i \tag{5.35}
\]
and
\[ y_2 = \sqrt{\frac{P_R}{d_2^2}} h_2 \left( \sqrt{b_1 \tilde{x}_1} + \sqrt{b_2 \tilde{x}_2} \right) + n_2. \] (5.36)

Further, \( D_1 \) applies the SIC technique to detect \( \tilde{x}_2 \) whereas treating its own signal \( \tilde{x}_1 \) as a background noise. Therefore, the SNR of detecting \( \tilde{x}_2 \) can be written by
\[ \gamma_{1,2} = \frac{|h_1|^2 b_2 \rho_R}{|h_1|^2 b_1 \rho_R + d_1^2}. \] (5.37)

Then, on the condition that \( \gamma_{1,2} \geq \xi_2 \), i.e., the SIC is successfully implemented at \( D_1 \) to remove \( x_2 \), the data detected in (5.37) is removed from the remaining signal and the SNR to detect \( \tilde{x}_1 \) at \( D_1 \) can be written as
\[ \gamma_{1,1} = \frac{|h_1|^2 b_1 \rho_R}{d_1^2}. \] (5.38)

Regarding \( D_2 \), it decodes \( \tilde{x}_2 \) while \( \tilde{x}_1 \) is treated as a background noise, which SINR can be expressed as
\[ \gamma_{2,2} = \frac{|h_2|^2 b_2 \rho_R}{|h_2|^2 b_1 \rho_R + d_2^2}. \] (5.39)

Now, the achievable rates for \( x_1 \) and \( x_2 \) are derived by [170]
\[ R_1 = \frac{1}{2} \log_2 \left( 1 + \min (\gamma_{R,1}, \gamma_{1,1}) \right) \] (5.40)
and
\[ R_2 = \frac{1}{2} \log_2 \left( 1 + \min (\gamma_{R,2}, \gamma_{1,2}, \gamma_{2,2}) \right), \] (5.41)
respectively.

### 5.4.2.1 Outage Performance Analysis

This section will evaluate the OP for \( x_1 \) and \( x_2 \) for Nakagami-\( m \) distribution. The OP of \( x_1 \) can be expressed by followings
\[ P_{\text{out},1}(\xi_1) = 1 - \Pr[\min (\gamma_{R,1}, \gamma_{1,1}) > \xi_1] = 1 - \Pr[\gamma_{1,1} > \xi_1] \Pr[\gamma_{R,1} > \xi_1] \]
\[ = F_{\gamma_{1,1}}(\xi_1) + F_{\gamma_{R,1}}(\xi_1) - F_{\gamma_{1,1}}(\xi_1)F_{\gamma_{R,1}}(\xi_1). \] (5.42)
Proposition 9 The closed-form OP expression for \( x_1 \) can be written as

\[
P_{\text{out}, 1}(\xi_1) = 1 - \frac{\Gamma(m_{SR}, \gamma_{SR}) \lambda_{SR} \xi d_d}{\Gamma(m_{SR})} \left[ \frac{\Gamma(m_1, \gamma_{1X} d_d)}{\Gamma(m_1)} \left( 1 - \frac{1}{\Gamma(m_{RP}, \lambda_{RP} \xi_{RP} d_{RP})} \right) \right] + \frac{\lambda_{RP} d_{RP}}{\Gamma(m_{RP})} \sum_{i=0}^{m_{RP} - 1} \left( \frac{\xi d_d}{\lambda_{RP} \xi_{RP} d_{RP}} \right)^i \frac{c_i}{i!} \Gamma(m_{RP} + i, \rho) \left( \frac{\rho d_{RP}}{\xi_{RP}} c_i \right) \]  

(5.43)

**Proof:** See Appendix C.5 for details.

Similarly as \( x_1 \), the OP of \( x_2 \) is expressed by

\[
P_{\text{out}, 2}(\xi_2) = 1 - \Pr[\min(\gamma_{R,2}, \gamma_{1,2}, \gamma_{2,2}) > \xi_2] = 1 - \Pr[\gamma_{R,2} > \xi_2] \Pr[\gamma_{1,2} > \xi_2] \Pr[\gamma_{2,2} > \xi_2] = F_{\gamma_{1,2}}(\xi_2) + (1 - F_{\gamma_{1,2}}(\xi_2)) (F_{\gamma_{2,2}}(\xi_2) + F_{\gamma_{R,2}}(\xi_2) - F_{\gamma_{2,2}}(\xi_2) F_{\gamma_{R,2}}(\xi_2)).
\]

(5.44)

Proposition 10 The closed-form OP expression for \( x_2 \) is given in C.6.

**Proof:** See Appendix C.6 for details.

### 5.4.2.2 Numerical Results

In this section, numerical and simulation results of the OP expressions for Nakagami-\( m \) fading will be presented. The system parameters are adopted as follows: \( d_{RP} = d_{SR} = d_1 = d \), \( d_2 = 2d \), \( 4d \), \( 6d \), \( \tau = 3 \), \( I = 25 \) dB and the corresponding SNR threshold \( \xi = \xi_1 = \xi_2 = 1 \) dB. Moreover, parameters of Nakagami-\( m \) distribution are taken as \( m = m_{RP} = m_{SR} = m_1 = m_2 = 1 \), \( R = \hat{r}_{RP} = \hat{r}_{SR} = \hat{r}_1 = \hat{r}_2 = \sqrt{2} \).

Fig. 5.14 illustrates the OP versus the transmit SNR for both messages of NOMA users. Here, it is assumed that \( a_2 = b_2 = 0.8 \). It is clearly seen from the plot that the OP of \( x_1 \) shows better performance compared to that of \( x_2 \) for low SNR levels (below 30 dB). On the other hand, after 30 dB, the OP of \( x_2 \) perform better. This can be explained by the fact that \( D_1 \) implements the SIC method in order to decode the desired message (which intuitively provides us with higher SNR value) whereas \( D_2 \) decodes \( x_2 \) by treating \( x_1 \) as a background noise. The OP curve of \( x_1 \) starts saturating at 25 dB and then, after 35 dB, it becomes flat due to the fact that the OP of \( x_1 \) in (5.43) cannot outperform that of \( x_1 \) observed at \( R \) (as the relay transmit power is constrained). Moreover, it can be seen that the OP of \( x_2 \) remains its improvement till it reaches the best performance defined by the OP of \( x_2 \) attainable at \( R \). Furthermore, it is noticed that the increase of \( d_2 \) results in the
Figure 5.14. The OP versus the transmit SNR for $x_1$ and $x_2$ when $d_2 = \{2 \text{ m}, 4 \text{ m}, 6 \text{ m}\}$, $a_1 = b_1 = 0.2$ and $a_2 = b_2 = 0.8$.

degradation of the OP performance. However, the saturation of the OP curves happens at higher SNR levels when the value of $d_2$ is higher.

In the previous results, it was assumed that $a_1 = b_1 = 0.2$ and $a_2 = b_2 = 0.8$. According to the NOMA concept, different PA coefficients are applied to the user terminals located at the different distances from the transmitter, i.e., $R \rightarrow D_1$ and $R \rightarrow D_2$. However, this is not valid for the $S \rightarrow R$ link and thus we can find OF-based PA coefficients $a_1$ and $a_2$ to provide equal outage performance for $x_1$ and $x_2$ since we found that at high SNR the OP curves of $x_1$ and $x_2$ saturate at the OP values defined by the outage observed at $R$. Therefore, in Fig. 5.15, we plot the OP of $x_1$ and $x_2$ as a function of $a_1$ attainable at $D_1$ and $D_2$, respectively. The following system parameters are adopted: $d_{RP} = d_1 = 2d$, $d_2 = 2d$ and $d_{SR} = \{d; 2d\}$. The observation SNR point is 30 dB. The increase of $a_1$ results in the OP improvement of $x_1$ while deteriorating the performance of the message $x_2$. For both distances, each pair of the corresponding OP curves can be characterized by the same intersection point which can be simply determined using Mathematica software. For the defined system parameters, the OF-based power coefficient $a^*_1$ equals 0.307 which provides the equal OP for both users (will be used for the next results discussion).

In Fig. 5.16, the OF-based OP results for $a^*_1 = 0.307$ and $a^*_2 = 0.693$ are depicted. It is noted that, compared with the results in Fig. 5.14, the saturation of the OP curves starts at 0.0128 for both $x_1$ and $x_2$ observed at the corresponding destination users for $b^*_1 = 0.089$ and $b^*_2 = 0.911$, i.e., the OP of $x_1$ improves while the OP of $x_2$ degrades. In the case when $d_{SR}$ is increased, the saturation starts at higher SNR levels. It can be explained by
Figure 5.15. The OP versus $a_1$ for $x_1$ and $x_2$ at 30 dB when $d_{SR} = \{d, 2d\}$.

Figure 5.16. The optimized OP versus the transmit SNR for $x_1$ and $x_2$ when $d_2 = \{2d, 4d, 6d\}$. 
the simple fact that more transmit power is required to overcome the path-loss attenuation and support the required performance.

5.5 Outage analysis of dual-hop CR-NOMA: Case 2

In this part, we consider a dowlink underlay CR-NOMA model consisting of \( K \) NOMA secondary destination users. It is assumed that both \( S \) and \( R \) are restricted by the maximum transmit power. This assumption is practical and provides extra useful insights into the system performance. Besides, considering the ITC results in the appearance of the additional RVs, which, in turn, significantly complicates the already complex analysis. Moreover, we consider imperfect CSI conditions and applied interference from primary transmitters to secondary receive nodes. The exact generalized closed-form expressions for the OP of the secondary destination users are derived and verified by Monte-Carlo simulations.

5.5.1 DF CR-NOMA with \( K \) secondary NOMA users

5.5.1.1 System Model

A downlink dual-hop DF underlay CR-NOMA network is shown in Fig. 5.17, where the system model consists of the \( PR \) and the SN with \( S \), \( R \) and \( K \) NOMA secondary destination users (\( D_1, ..., D_{K-1}, D_K \)). Fading-channel gains, i.e., \( h_{SP}, h_{SR}, h_{RP}, h_1, ..., h_{K-1} \) and \( h_K \), follow i.n.i.d. Rayleigh block fading model. In all previous system models in this chapter, it is assumed perfect CSI. However, due to the unpredictable characteristics of the wireless medium, in practice, the wireless CSI becomes imperfect. Considering this, the real channels, by assuming imperfect CSI and minimum mean square estimation error model, can be written by \([68, 176]\)

\[
h_t = \hat{h}_t + \tilde{h}_t, \ \forall t \in \{SP, SR, 1, ..., K - 1, K\}, \tag{5.45}
\]

where \( \hat{h}_t \) denotes the estimated channel coefficient with \( \mathcal{CN}(0, \sigma^2_{\hat{h}_t}) \) while \( \tilde{h}_t \) stands for the error term, which is modeled as a complex Gaussian distributed RV with \( \mathcal{CN}(0, \sigma^2_{\tilde{h}_t}) \), where \( \sigma^2_{\hat{h}_t} \) is the variance. The distances between nodes are expressed by \( d_{SP}, d_{SR} \) and \( d_1, ..., d_{K-1}, d_K \). Moreover, the PN interference\(^4\) received at the SN users can be indicated by \( P_I \). It is assumed that \( S \) and \( R \) can interfere with \( PR \). Hence, the transmit power level

\(^4\)In general, CSI of primary transmitters is not available at secondary receivers. Thus, with regard to the central limit theorem [177], secondary receivers can manage all interference signals coming from primary transmitters as the AWGN noise with \( \mathcal{CN}(0, \nu \sigma^2) \), where \( \nu \) is the scaling coefficient of \( P_I \). For instance, considering HetNet systems, when the receiver obtains interference from many sources, the total interference signals from these \( N \) interfering nodes can be approximated as a sum of \( n \) terms, which, with the respect to the CLT, can be further approximated by a complex Gaussian distribution [178], [179, §3.9.2].
at the secondary transmit node $L$, $\forall L \in \{S, R\}$, is restricted as

$$P_L \leq \min \left( \frac{I_{L_P}^{d_L}}{\|h\|^2}, \bar{P}_L \right), \quad (5.46)$$

where $\bar{P}_L$ denotes the maximum average allowed transmit power level at $L$.

Now, by considering the aforementioned power restrictions, $S$ conveys the superimposed signal to $K$ secondary NOMA destination users with the help of $R$ in two-time slots, which can be written as

$$x_S = \sum_{p=1}^{K} \sqrt{\alpha_p P_S} x_p, \quad (5.47)$$

where $x_p$, with $\mathbb{E}(\|x_p\|^2) = 1$, is the intended message to $D_p$ while $\alpha_p$ denotes the PA factor, with $\sum_{p=1}^{K} \alpha_p = 1$. Furthermore, regarding the system model, the channel gains of destination NOMA users are ordered as $h_1 < \ldots < h_{K-1} < h_K$, while $\alpha_1 > \ldots > \alpha_{K-1} > \alpha_K$, which means that the channel of $D_1$ is weaker than the channel of $D_{K-1}$. Accordingly, a higher power portion is allotted to $D_1$. 

Figure 5.17. A downlink underlay dual-hop CR-NOMA network.
In the 1st time slot, $R$ obtains the superimposed NOMA signal as follows

$$y_R = \left( \hat{h}_{SR} + \tilde{h}_{SR} \right) \sqrt{\frac{P_S}{d_{SR}} \sum_{p=1}^{K} \sqrt{\alpha_p x_p}} + P_I + n_R$$

$$= \hat{h}_{SR} \sqrt{\frac{P_S}{d_{SR}} \sum_{p=1}^{K} \sqrt{\alpha_p x_p}} + \tilde{h}_{SR} \sqrt{\frac{P_S}{d_{SR}} \sum_{p=1}^{K} \sqrt{\alpha_p x_p}} + P_I + n_R. \quad (5.48)$$

For mathematical tractability and without loss of generality, it is assumed that each SN receive node obtains the same $P_I$ from the primary transmitters.

Further, $R$ applies the SIC with the following decoding order of the users’ messages: first of all, the message of the user 1 ($j = 1$) is decoded at the same time treating the other messages ($j = 2, ..., K$) as a background noise. Further, $R$ removes the decoded message from the obtained signal. In the next phase, $R$ decodes the message of the user 2 by treating the remained messages ($j = 3, ..., K$) as a background noise, and so on and so forth. In the final stage, the message of the user $K$ can be decoded at $R$ without any inter-user interference. Therefore, the SINR at $R$ to decode $x_j$ is derived by

$$\gamma_{R,j} = \frac{\left| \hat{h}_{SR} \right|^2 \alpha_j \rho_S}{\left| \hat{h}_{SR} \right|^2 \rho_S \Theta + \zeta_R d_{SR}^\nu}, \quad (5.49)$$

where $\zeta_R = \frac{\rho_S \sigma_{\tilde{h}_{SR}}^2}{\hat{h}_{SR}} + \nu + 1$ and $\Theta = \sum_{p=j+1}^{K} \alpha_p$. Then, on the condition that $\gamma_{R,j} \geq \xi_j$, i.e., the SIC is successfully implemented at $R$ to remove $x_j$, the messages detected in (5.49) are removed from the received signal and the SNR to detect $x_K$ at $R$ can be written as

$$\gamma_{R,K} = \frac{\left| \hat{h}_{SR} \right|^2 \alpha_K \rho_S}{\zeta_R d_{SR}^\nu}. \quad (5.50)$$

Then, within the 2nd time slot, the detected superimposed signal at $R$ is further forwarded to all secondary NOMA destination users, which is shown by

$$x_R = \sum_{p=1}^{K} P_R \beta_p \tilde{x}_p, \quad (5.51)$$

where $\tilde{x}_p$ is the detected message of the corresponding NOMA user at $R$; $\beta_p$, with $\sum_{p=1}^{K} \beta_p = 1$.\]
1, satisfying $\beta_1 > \ldots > \beta_{K-1} > \beta_K$, denotes the relay PA factor. Then, the received superimposed signal at $D_j$ can be written as

$$y_j = \hat{h}_j \sqrt{\frac{P_R}{d_j}} \sum_{p=1}^{K} \sqrt{\beta_p} \tilde{x}_p + \hat{h}_j \sqrt{\frac{P_R}{d_j}} \sum_{p=1}^{K} \sqrt{\beta_p} \tilde{x}_p + P_f + n_j, \forall j \in \{1, \ldots, K - 1, K\}.$$  

(5.52)

Moreover, $D_k$ uses the SIC in order to decode the undesired signal of $D_j$ ($1 \leq j \leq k \leq K$) by following the same method as in $R$. Hence, the SINR of decoding the message $j$ at $D_k$ can be expressed as

$$\gamma_{k,j} = \frac{|\hat{h}_k|^2 \beta_j \rho_R}{|\hat{h}_k|^2 \rho_R \Phi + \zeta_k d_k}, \forall k \in \{1, \ldots, K\},$$  

(5.53)

where $\Phi = \sum_{p=j+1}^{K} \beta_p$ and $\zeta_k = \frac{\rho R \sigma^2 h_k x_k}{d_k} + \nu + 1$. Finally, if the messages of $K - 1$ users are detected successfully, the $D_K$ can detect its own message $K$ without any inter-user interference, which SNR is written by

$$\gamma_K = \frac{|\hat{h}_K|^2 \beta_K \rho_R}{\zeta_K d_K},$$  

(5.54)

where $\zeta_K = \frac{\rho R \sigma^2 h_K x_K}{d_K} + \nu + 1$. The achievable rates for the messages dedicated to $D_j$ ($1 \leq j \leq k < K$) and $D_K$ can be expressed by [170]

$$R_j = \frac{1}{2} \log_2 \left(1 + \min (\gamma_{R,j}, \gamma_{k,j})\right)$$  

(5.55)

and

$$R_K = \frac{1}{2} \log_2 \left(1 + \min (\gamma_{R,K}, \gamma_K)\right),$$  

(5.56)

respectively.

### 5.5.1.2 Outage Performance Analysis

The OP for the proposed system model is investigated in this part. The message $x_j$ is considered to be in an outage if the achievable rate of $x_j$ is below a predefined target rate $R_{th,j}$ (which corresponding receive SNR threshold is $\xi_j = 2^{2R_{th,j}} - 1$), i.e., $P_{out} = \ldots$
Pr \[ R_j < R_{\text{th}} \]. Hence, the OP of \( x_j \) is derived by followings

\[
P_{\text{out},j}(\xi_j) = 1 - \Pr \left[ \min (\gamma_{R,j}, \gamma_{k,j}) > \xi_j \right]
\]

\[
= F_{\gamma_{R,j}}(\xi_j) + \left( 1 - F_{\gamma_{R,j}}(\xi_j) \right) \left( -1 \right)^n \prod_{t} F_{\gamma_{t,j}}(\xi_j),
\]

(5.57)

where \( t = \{ m(0), m(1), \ldots, m(n) \} \), \( n = K - j \), \( m(0) = j, j + 1, \ldots, K \), \( m(1) = j + 1, \ldots, K \), \( m(n) = j + n, \ldots, K \) and \( m(n) \neq m(n+1) \). For instance, if \( K = 3 \), the OP of \( x_2 \) \( (j = 2) \) is derived from (5.57) as follows: \( n = 1 \), \( m(0) = 2, 3 \) and \( m(1) = 3 \) \((m(0) \neq m(1)) \). Hence, firstly, the CDFs of \( F_{\gamma_{2,2}} \) and \( F_{\gamma_{3,2}} \) are summed, then the product of the both CDFs are subtracted (since \( m(0) \neq m(1) \rightarrow m(0) = 2 \) and \( m(1) = 3 \)), and the resulted equation can be written by followings

\[
P_{\text{out},2}(\xi_2) = 1 - \Pr \left[ \min (\gamma_{R,2}, \gamma_{3,2}, \gamma_{2,2}) > \xi_2 \right]
\]

\[
= 1 - \Pr [\gamma_{R,2} > \xi_2] \Pr [\gamma_{3,2} > \xi_2] \Pr [\gamma_{2,2} > \xi_2]
\]

\[
= F_{\gamma_{R,2}}(\xi_2) + \left( 1 - F_{\gamma_{R,2}}(\xi_2) \right) \left( F_{\gamma_{3,2}}(\xi_2) + F_{\gamma_{2,2}}(\xi_2) - F_{\gamma_{2,2}}(\xi_2) F_{\gamma_{3,2}}(\xi_2) \right).
\]

(5.58)

**Proposition 11** The closed-form expression for the CDF of the RV \( \gamma_{R,j} \) in (5.49) is expressed as

\[
F_{\gamma_{R,j}}(\xi_j) = \begin{cases} 
1 - \left( e^{-\xi_j \rho R \frac{\sigma R}{\sigma S}} - \frac{G_{R,j} e^{\frac{\rho R \sigma R + \xi_j \rho R \sigma S}{2\mu S}}}{\mu + G_{R,j}} \right), & \xi_j < \frac{\alpha_j}{\sigma}, \\
1, & \text{otherwise}.
\end{cases}
\]

(5.59)

**Proof:** See Appendix C.7 for details. \( \blacksquare \)

**Proposition 12** The closed-form expression for the CDF of the RV \( \gamma_{k,j} \) in (5.53) is expressed as

\[
F_{\gamma_{k,j}}(\xi_j) = \frac{K!}{(K-k)!(k-1)!} \sum_{i=0}^{K-k} \frac{(-1)^i}{k+i} \binom{K-k}{i} \sum_{n=0}^{k+i} \frac{(-1)^n}{n} \binom{k+i}{n} L_{i}(i) V_{i}(n)
\]

\[
\times \left( 1 - e^{-\frac{\mu R \rho R}{2\mu R}} \right) e^{-\frac{\xi_j \rho R \sigma R}{2\mu R}} + \frac{e e^{\frac{\rho R \sigma R \gamma_{k,j} + \xi_j \rho R \sigma S}{2\mu S}} e^{\frac{\rho R \sigma R}{2\mu R}}}{\epsilon + G_{k,j} \sigma R} \right)
\]

(5.60)
Proof: See Appendix C.8 for details.

Now, using (5.59) and (C.51), the OP expression in the example shown in (5.58) can be expressed in its closed form as

$$P_{\text{out},2}(\xi_2) = 1 - \Lambda_{R,2} + \sum_{p=2}^{\infty} L_p(i) V_p(n) \left[ \sum_{p=2}^{\infty} L_p(i) V_p(n) \right]$$

where $\xi_2 < \alpha_2$ and $\xi_2 < \beta_2$, otherwise, $P_{\text{out},2}(\xi_2) = 1$.

The OP of $x_K$, similar to $x_j$, can be written as follows

$$P_{\text{out},K}(\xi_K) = 1 - Pr[\min(\gamma_{R,K}, \gamma_K) > \xi_K] = F_{\gamma_R,K}(\xi_K) + F_{\gamma_{R,K}}(\xi_K) - F_{\gamma_R}(\xi_K) F_{\gamma_K}(\xi_K),$$

where $\xi_K$ is the receive SNR threshold at $D_K$. Now, following the same steps as in Appendix C.7, the CDF of $\gamma_{R,K}$ can be written in a closed-form as

$$F_{\gamma_{R,K}}(\xi_K) = 1 - \left( e^{-\frac{\xi_K \alpha_{R,K} d_{p,R,S}^2}{2 \alpha_{R,K} p_S}} - \frac{\mathcal{G}_{R,K} e^{-\frac{\alpha_{R,K} p_S^2}{2 \alpha_{R,K} p_S \beta_{R,K}}}}{\alpha_{R,K} + \mathcal{G}_{R,K}} \right),$$

In addition, by applying order statistics for the strongest channel [180, Eq. (19)], the CDF of $\gamma_{R,K}$ can be written in a closed-form as

$$F_{\gamma_K}(\xi_K) = V_K(m) \left( 1 - e^{-\frac{\xi_K \alpha_K d_{K,R}^2}{2 \alpha_K p_R}} \right)^K,$$

Then, following the same steps as in Appendix C.8, the CDF of $\gamma_K$ can be written in a closed-form as

$$F_{\gamma_K}(\xi_K) = V_K(m) \left( e^{-\frac{\xi_K \alpha_K d_{K,R}^2}{2 \alpha_K p_R}} + \frac{\beta_K e^{-\frac{\alpha_K p_S \beta_{K,R}^2}{2 \alpha_K p_S \beta_{K,R}}}}{\beta_K + \mathcal{G}_{K,R} m} \right),$$
Figure 5.18. The OP of messages $x_1$ and $x_2$ versus the transmit SNR when $d_1 = \{1.5d, 2d, 3d\}$, $\alpha_1 = 0.8$, $\alpha_2 = 0.2$, $\nu = 0$ and $\sigma_h^2 = 0$.

where $\mathcal{V}_K(m) = \sum_{m=0}^{K} \binom{K}{m} (-1)^m$. Finally, substituting (5.63) and (5.65) into (5.62), the OP of $x_K$ can be written in its closed-form as

$$P_{\text{out},K} = 1 - \Lambda_{R,K} \left[ 1 - \mathcal{V}_K(m) \left( I e^{-\xi_K^R \frac{G_{Km}}{\beta_K R^R}} + \beta_K e^{-\frac{\frac{\beta_K G_{Km}}{R^R} (\frac{\nu_K + \nu_{Km}}{\beta_K R^R} + \bar{G}_{Km})}{\beta_K + \bar{G}_{Km}}} \right) \right].$$

Now, using (5.55) and (5.56), the average throughput of the messages devoted to $D_j$ ($j < \forall k \leq K$) and $D_K$ are respectively written as

$$C_j = R_j (1 - P_{\text{out},j}), \quad (5.67)$$
$$C_K = R_K (1 - P_{\text{out},K}). \quad (5.68)$$

5.5.1.3 Numerical Results

In this section, numerical results on the OP over the Rayleigh fading is presented. As a special case, it is considered that the system model consists of two secondary NOMA destination users ($K = 2$), i.e., $D_1$ and $D_2$. Similarly in the previous system models, the next system parameters are considered: $d_1 = \{1.5d, 2d, 3d\}$, $d_{SP} = d_{SR} = d_{RP} = d_2 = d$, equal source and relay transmit power, i.e., $P_S = P_R = P$, $I = 20$ dB, $\alpha_i =$
Fig. 5.19. The OP of messages $x_1$ and $x_2$ versus the SNR threshold at 20 dB when $d_1 = \{1.5d, 2d, 3d\}$, $\alpha_1 = 0.8$, $\alpha_2 = 0.2$, $\nu = 0$ and $\sigma^2_{\tilde{h}_i} = 0$.

\[ \beta_i, \forall i \in \{1, 2\}, \nu = 0.5, \xi = \xi_1 = \xi_2 = 3 \text{ dB}, \sigma^2_{\tilde{h}_i} = \{0.001, 0.005\}, \text{ with } \forall \nu \in \{SP, SR, RP, 1, 2\}, \text{ and } \tau = 2.7 \text{ (for urban area)}. \]

Fig. 5.18 illustrates the OP results for messages $x_1$ and $x_2$ obtained using (5.57) and (5.66), respectively. The PA factors are taken as $\alpha_1 = \beta_1 = 0.8$ and $\alpha_2 = \beta_2 = 0.2$. It is observed that the OP of $x_2$ performs better than that of $x_1$ at all SNR values. This phenomenon is explained by the fact that $D_2$ applies the SIC to remove the message of $D_1$, whereas $D_1$ detects its message without eliminating the message of $D_2$. The comparison of the OP performance of message $x_1$ at different $d_1$ shows that the outage performance degrades as the distance $d_1$ increases. In addition, Monte Carlo simulations show a consistency with the analytical ones, which confirms the correctness of the derived closed-form expressions for the OP of messages $x_1$ and $x_2$.

In Fig. 5.19, we demonstrate the OP performance versus the SNR threshold for messages $x_1$ and $x_2$ with different values of $d_1$. From the plot, it is seen that the OP of both messages degrades by increasing the SNR threshold value. It is noticed that the increase of the distance $d_1$ degrades the OP of message $x_1$. For example, when $d = 1.5d$ and $d = 3d$, message $x_1$ reaches an outage at 6 dB and 5.6 dB, respectively. The message $x_2$ shows the best result by reaching the OP of 0.23 at 6 dB.

Fig. 5.20 shows the simulated results on the OF-based PA factors for messages $x_1$ and
Figure 5.20. The OP of messages $x_1$ and $x_2$ versus PA factors at 20 dB when $d_1 = \{1.5d, 2d, 3d\}$, $\nu = 0$ and $\sigma^2_{\tilde{h}_i} = 0$.

Table 5.3. The observed OF-based PA factors ($K$ users).

<table>
<thead>
<tr>
<th>$d_1$</th>
<th>1.5d</th>
<th>2d</th>
<th>3d</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\alpha_1^<em>, \beta_1^</em>$</td>
<td>0.885</td>
<td>0.93</td>
<td>0.965</td>
</tr>
<tr>
<td>$\alpha_2^<em>, \beta_2^</em>$</td>
<td>0.115</td>
<td>0.07</td>
<td>0.035</td>
</tr>
</tbody>
</table>

$x_2$ at 20 dB. It is seen from the plot that, for all $d_1$ values, the OP of message $x_1$ is equal to 1 when $\psi > \frac{\alpha_1^*}{\alpha_2^*}$. On the other hand, when $\psi < \frac{\alpha_1^*}{\alpha_2^*}$, an increase of the value of $\alpha_1$ leads to an improvement of the OP of message $x_1$, by the same time, decreasing that of message $x_2$. The observed OF-based PA factors for different $d_1$ values are illustrated in Table 5.3.

Fig. 5.21 demonstrates the outage performance versus the transmit SNR with OF-based $\alpha_1^*, \beta_1^*$ and $\alpha_2^*, \beta_2^*$ considering various $d_1$. From the plot, we observe that for all values of $d_1$, message $x_2$ shows an outperformance over message $x_1$ in terms of the OP at SNR levels higher than 17 dB. However, at lower SNR levels, the OP difference of both messages is not considerable, which shows that OF-based PA factors provide OP fairness for both messages.

The results on the outage performance versus the SNR threshold using OF-based $\alpha_1^*, \beta_1^*$ and $\alpha_2^*, \beta_2^*$ for different $d_1$ at the transmit SNR of 20 dB are illustrated in Fig. 5.22. It is observed that the OF-based PA factors improve the OP of $x_1$ in all $d_1$ values providing fairness for both messages.

In Fig. 5.23, the comparison of the CR-NOMA with CR-OMA in terms of the outage performance is demonstrated. In order to show the impact of the ITC on the OP of messages $x_1$ and $x_2$, two scenarios are considered: with no ITC, i.e., $I = \infty$ and when
Figure 5.21. The OP of messages $x_1$ and $x_2$ versus the transmit SNR with OF-based PA factors when $d_1 = \{1.5d, 2d, 3d\}$, $\nu = 0$ and $\sigma^2_{h_i} = 0$.

Figure 5.22. The OP of messages $x_1$ and $x_2$ versus the SNR threshold with OF-based PA factors at 20 dB when $d_1 = \{1.5d, 2d, 3d\}$, $\nu = 0$ and $\sigma^2_{h_i} = 0$. 
\[ I = 20 \text{ dB}. \] It is obvious from Fig. 5.23 that both messages of CR-NOMA obtain better outage performance compared to those messages of the CR-OMA for all SNR levels. It is also noticed that saturation of outage curves for CR-NOMA starts at lower SNR regimes compared to those for the CR-OMA. This phenomenon can be explained by the fact that, regarding (5.46), the source and relay transmit power of \( \frac{1}{2} P \) in the CR-OMA network increases the level of the ITC at \( PR \), which consequently raises the SNR value where the OP saturation starts.

Fig. 5.24 illustrates the effect of the imperfect CSI and PN interference on the OP performance of NOMA messages. Here, we consider the case when \( I = \infty \). It can be seen from the plot that the OP of both messages degrades when imperfect CSI and PN interference exist. When the PN interference is inserted, the OP of both messages degrades linearly. It is also noticed that when \( \sigma_{h_i}^2 = 0.001 \), the effect of the channel error variance on the outage performance is insignificant at SNR levels below 15 dB. However, the impact of the imperfect channel on the OP curves of both messages becomes more significant at moderate SNRs. Moreover, due to the impact of the channel error variance, the OP becomes SNR-independent at higher SNR levels and starts saturating at 48 dB for message \( x_1 \) and at 45 dB for message \( x_2 \). In addition, when \( \sigma_{h_i}^2 = 0.005 \), the channel error variance starts to degrade the OP of messages \( x_1 \) and \( x_2 \) at 17 dB and 12 dB, respectively, while, the OP saturation for the same messages begins at 45 dB and 40 dB, accordingly.

Fig. 5.25 shows the results on the average throughput obtained from (5.67) and (5.68). It is clearly shown that the message \( x_2 \) outperforms the message \( x_1 \) in terms of the average
Figure 5.24. The OP of messages $x_1$ and $x_2$ versus the transmit SNR with imperfect CSI and PN’s interference when $\alpha_1 = 0.8$ and $d_1 = 2d$.

Figure 5.25. The average throughput versus the transmit SNR for NOMA messages with $\alpha_1 = 0.8$ and $d_1 = 2d$. 
throughput performance. The message $x_2$ achieves the maximum average throughput of 3.1 bps/Hz and 2.04 bps/Hz at 39 dB for the perfect CSI and when $\sigma^2_{h_i} = 0.005$, respectively. After that, the curves of the average throughput start saturating because of the ITC. Hence, it shows that imperfect CSI degrades the average throughput by 1.06 bps/Hz at 39 dB. On the other hand, the saturation of the average throughput of the message $x_1$ begins at 30 dB by achieving 0.71 bps/Hz.

5.5.2 System Model for AF CR-NOMA

The proposed system model in this section is the extension of Fig. 5.1, where only $S$ was restricted by the maximum transmit power. Now, it is assumed that $PR$ applies the ITC to both secondary transmit nodes $S$ and $R$ as they cause interference to the PN. Moreover, $PTs$ cause interference to the SN (See Fig. 5.26). The channel between $R$ and $PR$ is denoted by $h_{RP}$, which also follow i.n.i.d. Rayleigh block fading model.

During the 1st time slot, $R$ obtains the following signal

$$ y_R^{AF} = \sqrt{\frac{P_S}{d_{SR}^2}} h_{SR} \sum_{i=1}^{2} \sqrt{\alpha_i} x_j + P_I + n_R, \quad (5.69) $$
In the 2nd time slot, based on the AF protocol, first, $R$ amplifies the received signal with an amplification factor $G = \frac{1}{\sqrt{\frac{P_s}{d_{SR} h_{SR}^2} + \sigma^2}}$ and, then, forwards it to $U_j$, $j \in \{1, 2\}$. Thus, the received signal at $U_j$ can be written as

$$y_{j}^{AF} = \sqrt{\frac{P_s P_R h_{SR} h_j}{d_{SR}^2 d^2_j}} \sum_{i=1}^{2} \sqrt{\alpha_i} x_j G + \sqrt{\frac{P_s}{d_{SR} h_j}} n_R G + \sqrt{\frac{P_s}{d_{SR} h_j}} P_I G + P_I + n_j, \quad (5.70)$$

Then, as due to higher value of $\alpha_1$, $D_1$ detects its own signal by treating $x_2$ as a background noise, which SINR can be written as

$$\gamma_1^{AF} = \frac{\frac{P_s}{d_{SR} h_{SR}} |h_{SR}|^2 \epsilon_1}{n_s Y n_1 X \alpha_1 + n_1 X \bar{\nu} + n_s Y \bar{\nu} + \bar{\nu}} \quad (5.71)$$

where $\bar{\nu} = \nu + 1$. Then, $D_2$ treats its own signal as a noise when detects $x_1$ by the SINR

$$\gamma_{2,1}^{AF} = \frac{n_s Y n_1 Q \alpha_1}{n_s Y n_2 Q \alpha_1 + n_2 Q \bar{\nu} + n_s Y \bar{\nu} + \bar{\nu}}. \quad (5.72)$$

After successfully removing $x_1$ using the SIC, $d_2$ detects its own message $x_2$ from the remaining signal with the SINR

$$\gamma_2^{AF} = \frac{n_s Y n_2 Q \alpha_2}{n_2 Q \bar{\nu} + n_s Y \bar{\nu} + \bar{\nu}}. \quad (5.73)$$

### 5.5.3 Outage Analysis

The OP of $D_1$ can be written as

$$P_{out,1}^{AF}(\xi_1) = 1 - \Pr \left[ \gamma_1^{AF} > \xi_1 \right] = 1 - \left(1 - F_{\gamma_1}^{AF}(\xi_1) \right). \quad (5.74)$$
Proposition 13 The exact OP expression of $D_1$ can be written as

$$
\begin{align*}
P_{\text{out,1}}^{\text{AF}}(\xi_1) &= (1 - e^{-C_S}) (1 - e^{-C_R}) \left( 1 - e^{-\frac{u_{s,1}(n_1 + n_2)}{n_1}} 2\sqrt{r_1} K_1 (2\sqrt{r_1}) \right) \\
&+ (1 - e^{-C_R}) \left( e^{-C_S} - e^{-\frac{C_S n_1 (u_{p,1} + 1) + n_p u_{p,1}}{n_1} \Omega_{B_2}} \right) + (1 - e^{-C_S}) (e^{-C_R} (1 - e^{-u_{s,1}}) \\
&+ e^{-C_R (u_{s,1} + 1)} u_{s,1} + 1 - e^{-\frac{C_R (m_1 + n_{s,1}) + m_u u_{s,1}}{m_1}} \frac{C_R}{\Upsilon_C} \sum_{k=0}^{\infty} (-\Upsilon_C)^k \frac{G_3}{k!} \frac{1}{3} \left( C_R \frac{\Theta_C}{\Upsilon_C} \left( 0, k - 1, k \right) \right) \\
&+ e^{-C_R - C_S - \Theta_{D_1}} \times \Psi(0) \frac{\pi}{C_R} M G_2 \left( \frac{\Upsilon_{D_2}}{M}, \frac{\zeta}{M} \right) 0 - \frac{1}{2} \left| 0 \right. 1 1 0 2 1 \\
&+ \Theta_{D_1} \sum_{k=1}^{\infty} \Psi(k) C_R^{k-1} \sum_{n=0}^{k-1} \left( \begin{array}{c} k - 1 \\ n \end{array} \right) \left( \frac{1}{C_R \zeta} \right) \sum_{m=0}^{\infty} (-\zeta C_R r)^m m! \\
&\times G_4 \left( \frac{\Upsilon_{D_2}}{\zeta} \right) \left| -n, m - n - 1, 2m - n - 2, 2m - n - 1 \right. 0, m - n - 1, 2m - n - 1, m - n \right),
\end{align*}
$$

where $\theta_1 < \frac{\alpha_1}{\alpha_2}$, otherwise, $P_1(\theta_1) \sim 1$.

Proof: See Appendix C.9 for details.

The OP derivation of $D_2$ can be found by following the similar approach as in Appendix C.9.

5.5.4 Numerical Results

We adopt the following system parameters to present the results for the above investigation: $P_I = 20$ dB, $\alpha_1 = 0.8, \alpha_2 = 0.2, \theta_1 = \theta_2 = 3$ dB, $\eta = 2.7, d_1 = 2d, d_{SP} = d_{SR} = d_2 = d, \alpha_1 = 0.8, \alpha_2 = 0.2$. Without loss of generality, $d$ is assumed to be unity in order to focus on the OP results.

Fig. 5.27 shows the OP results for $D_1$ and $D_2$. It is observed that the OP for $D_2$ performs better than that for $D_1$. It is due to the fact that $D_2$ removes the unwanted message of $D_1$ by applying the SIC technique which results in an improvement of the error probability in the message detection. Moreover, the OP results for both users in the NOMA system show better performance compared to the same users in the OMA scenario. Finally, it is noticed that primary interference ($\nu = 0.2$) considerably degrades the OP of secondary nodes.
Figure 5.27. The OP of Users 1 and 2 versus the transmit SNR when \( d_1 = \{2d, 4d\}, \alpha_1 = 0.8, \alpha_2 = 0.2 \) and \( \tau = 0 \).

5.6 Chapter Summary

This Chapter proposed the dual-hop underlay CR-NOMA networks with the imposed ITC at \( PR \) for different system model scenarios. The OP was investigated as the main performance metric for the considered system model applying DF and AF relaying methods over Rayleigh and Nakagami-\( m \) fading channels. First, the OP performance was evaluated to the scenario with two NOMA secondary users and the OF-based PA factors were found for both users to guarantee OP fairness between NOMA users. Then, considering the effect of the imperfect CSI on the outage performance, we derived a general closed-form OP expression of SUs when the number of secondary NOMA nodes was extended to \( K \) users. In addition, the obtained OP performance for CR-NOMA was compared to that for conventional CR-OMA to show the supremacy of the former. Finally, all derived analytical results were verified by Monte-Carlo simulations, which validated the accuracy of the performance analysis.
Chapter 6

Wireless Powered CR-NOMA

In this chapter, we study the OP of the dual-hop underlay CR network consisted of a secondary relay node with energy harvesting and two NOMA secondary destination users ($D_1$ and $D_2$). Moreover, the OF-based power allocation factors are found for different distances of $D_1$ in order to satisfy OP fairness for both $D_1$ and $D_2$. Additionally, the CR-NOMA is compared with CR-OMA in terms of the OP. Finally, the obtained analytical expressions are corroborated by Monte Carlo simulations.

6.1 Introduction

CR-NOMA is a promising technique that can not only improve the spectral efficiency but also can increase user connectivity in wireless communication networks [59]. Aside from the spectrum, another key objective of future 5G networks is to maximize energy efficiency. Hence, one of the promising methods to enhance energy efficiency is wireless EH [132]. In the EH-enabled networks, energy-limited nodes can harvest RF signals, then convert them into the DC [140]. The motivation for the application of SWIPT to NOMA can be illustrated in works [181–183]. In these works, SWIPT is applied for NOMA uplink transmission, where users harvest energy from the base station and then send their information to the base station simultaneously by utilizing the NOMA policy. The authors in [184] studied resource allocation for uplink SWIPT - NOMA transmission, where PA and the durations for power and information transfer are jointly designed to eliminate the doubly near-far effect.

Note that SWIPT is also can be applied to cooperative NOMA communication [185]. As it was already mentioned before, cooperative NOMA can cause effective aid to the users with weak channel conditions, by using the strong users as relay nodes. But, in real time communications, these users may not desire to perform as relay nodes, considering the fact that the procedure can consume a considerable amount of power which shorten their battery life. On the other hand, by implementing SWIPT, the strong users can harvest energy from signals conveyed by the base station and use that harvested energy to further forward signals of weak users. Thus, the authors in [185] studied SWIPT in cooperative
NOMA, where users are spatially randomly located and NOMA users which are located near to the source act as EH relays to assist far NOMA users. Moreover, inspired by the idea in [185], the authors in [186] investigated cooperative SWIPT-NOMA in terms of a transmitter beamforming design, a power splitting ratio optimization and a receiver filter design in order to maximize the rate of the relay node. Moreover, the authors in [187] studied the NOMA system with a wireless powered near user with the TS, the PS and generalized schemes, where the achievable rate regions of the TS and PS were derived in closed-form expressions. Furthermore, the OP of cooperative relaying transmissions in two-user SWIPT-NOMA networks was studied in [188], where the authors proposed a best-near best-far user selection scheme to use the EH-based near users as relay nodes. The authors in [189] proposed antenna selection method to study the OP for EH-based NOMA relaying networks over Nakagami-$m$ fading. In addition, the authors in [140] studied the EH-based cooperative CRN with multiple primary users and one secondary network, where interference at receive nodes is canceled with the IA technique. The derived outage performance results showed that the PS protocol outperforms the TS one.

### 6.2 Contributions of the Chapter

The main contributions of the Chapter are as follows:

- We study the OP for the cooperative underlay CR-NOMA networks with enabled the SWIPT technology.
- The OP for the proposed system model with EH-enabled $R$ calculated numerically. In addition, the OF-based PA factors are calculated depending on NOMA users’ channel quality. Thus, the proper evaluated PA factors can guarantee fairness among NOMA users in terms of the OP.
- We investigate how the PA factor in the PS mode and the distance of nodes affect on the harvested power level.
- Cooperative CR-OMA networks are taken as a benchmark to illustrate the better performance of the cooperative CR-NOMA.

The remainder of this chapter is organized as follows. Section 6.3 describes the system model with evaluated achievable rates. In Section 6.4, the OP analysis of both secondary destination users are presented. Moreover, in Section 6.5, numerical results are presented and discussed. Finally, Section 6.6 concludes the Chapter.
6.3 System Model

We consider a downlink dual-hop DF underlay CR-NOMA network consisting of PR and a SN with $S$, an energy-constrained $R$ operating in a half-duplex mode and two destination NOMA users ($D_1$ and $D_2$) as shown in Fig. 6.1. Channels between nodes, i.e., $h_{SP}$, $h_{SR}$, $h_1$ and $h_2$, are assumed to follow an i.n.i.d. Rayleigh block fading model with $CN(0, 1)$. Moreover, a global CSI is assumed to be available at each node. The corresponding distances between nodes are denoted by $d_{SP}$, $d_{SR}$, $d_1$ and $d_2$. Regarding the system model, $S$ and $R$ cause interference to PR, Thus, the SN communication is available only if PR does not receive harmful interference from the SN. Hence, the transmit power constraints are imposed at $S$ and $R$ regarding (5.46). Hence, considering the transmit power restriction, the superimposed signal $\sqrt{P_S}\alpha_1 x_1 + \sqrt{P_S}\alpha_2 x_2$ is conveyed from $S$ to $D_1$ and $D_2$ via the assistance of $R$ within two time slots. Without loss of generality, we assume that the channels of destination NOMA users are ordered as $h_1 < h_2$. Based on the ordering of the users’ channels, the corresponding PA factors are ordered as $\alpha_1 > \alpha_2$.

6.3.1 Power-Splitting Relaying

The PSR protocol is illustrated in Fig. 6.2, where, during the 1st time slot, a coefficient $\theta$, with $(0 < \theta < 1)$, of the received signal power at $R$ is utilized for the EH purpose, while the remaining signal $(1 - \theta)$ is used for the data detection at $R$. Hence, the signal received
at $R$ for the EH purpose can be written as

$$y_{R}^{EH} = h_{SR} \left( \sqrt{\frac{\alpha_1 P_S}{d_{SR}^r}} x_1 + \sqrt{\frac{\alpha_2 P_S}{d_{SR}^r}} x_2 \right) + \sqrt{\theta} n_R, \quad (6.1)$$

where $n_R \sim \mathcal{CN}(0, \sigma_R^2)$ stands for the AWGN at $R$. Therefore, neglecting the relatively small power harvested from the noise, the harvested energy at $R$ can be derived by [140]

$$E_H = \frac{\eta \theta T}{2} |h_{SR}|^2 \left( \frac{\alpha_1 P_S}{d_{SR}^r} + \frac{\alpha_2 P_S}{d_{SR}^r} \right) = \frac{\eta \theta T |h_{SR}|^2 P_S}{2 d_{SR}^r}, \quad (6.2)$$

where $\eta$ denotes the EH conversion efficiency, with $(0 < \eta < 1)$ and $T$ is the total time required for the end-to-end transmission. Regarding the fact that transmit power at $R$ relates to the EH as $P_R = 2E_H/T$, the transmit power at $R$ can be expressed as [127]

$$P_R = \frac{\eta \theta |h_{SR}|^2 P_S}{d_{SR}^r}. \quad (6.3)$$

Moreover, the signal received for the $S - R$ information transmission can be written as

$$y_{R}^{IT} = \sqrt{1 - \theta} h_{SR} \left( \sqrt{\frac{\alpha_1 P_S}{d_{SR}^r}} x_1 + \sqrt{\frac{\alpha_2 P_S}{d_{SR}^r}} x_2 \right) + n_R. \quad (6.4)$$

Furthermore, $R$ decodes and removes $x_1$ from (6.4), and $x_2$ can be detected from the remained signal. The SINR and SNR to decode $x_1$ and $x_2$ at $R$ can be derived as

$$\gamma_{R,1} = \frac{(1 - \theta) \alpha_1 \rho_S |h_{SR}|^2}{(1 - \theta) \alpha_2 \rho_S |h_{SR}|^2 + d_{SR}^r} \quad (6.5)$$

and

$$\gamma_{R,2} = \frac{(1 - \theta) \alpha_2 \rho_S |h_{SR}|^2}{d_{SR}^r}, \quad (6.6)$$

respectively.
Furthermore, during the 2\textsuperscript{nd} time slot, \( R \) forwards the detected superimposed signal \( \sqrt{P_R \beta_1} \tilde{x}_1 + \sqrt{P_R \beta_2} \tilde{x}_2 \) to \( D_1 \) and \( D_2 \). Therefore, \( D_1 \) and \( D_2 \) receive the following signals

\[
y_1 = h_1 \left( \sqrt{\frac{\beta_1 P_R}{d_1^r}} \tilde{x}_1 + \sqrt{\frac{\beta_2 P_R}{d_1^r}} \tilde{x}_2 \right) + n_1 \tag{6.7}
\]
and

\[
y_2 = h_2 \left( \sqrt{\frac{\beta_1 P_R}{d_2^r}} \tilde{x}_1 + \sqrt{\frac{\beta_2 P_R}{d_2^r}} \tilde{x}_2 \right) + n_2, \tag{6.8}
\]
respectively, where \( n_1 \) and \( n_2 \) stand for the AWGN noise terms at \( D_1 \) and \( D_2 \), respectively. Furthermore, \( D_2 \) implements SIC by detecting \( \tilde{x}_1 \) while considering its own data \( \tilde{x}_2 \) as a noise. The SINR of which can be written as

\[
\gamma_{2,1} = \frac{\eta \theta \beta_1 \rho_S |h_2|^2 |h_{SR}|^2}{\eta \theta \beta_2 \rho_S |h_2|^2 |h_{SR}|^2 + d_2^r d_{SR}^r}. \tag{6.9}
\]
Furthermore, \( D_2 \) extracts the detected data in (6.9) from the received signal. Then, \( D_2 \) detects its own signal from the remaining signal by the following SNR

\[
\gamma_{2,2} = \frac{\eta \theta \beta_2 \rho_S |h_2|^2 |h_{SR}|^2}{d_2^r d_{SR}^r}. \tag{6.10}
\]
Now, as \( D_1 \) is allocated with higher PA factors, it can detect its message by treating \( \tilde{x}_2 \) as a background noise with the following SINR

\[
\gamma_1 = \frac{\eta \theta \beta_1 \rho_S |h_1|^2 |h_{SR}|^2}{\eta \theta \beta_2 \rho_S |h_1|^2 |h_{SR}|^2 + d_1^r d_{SR}^r}. \tag{6.11}
\]
Then, the end-to-end achievable rates of \( x_1 \) and \( x_2 \) can be respectively written as the followings

\[
\mathcal{R}_1 = \frac{1}{2} \log_2 \left( 1 + \min (\gamma_{R,1}, \gamma_{2,1}, \gamma_1) \right) \tag{6.12}
\]
and

\[
\mathcal{R}_2 = \frac{1}{2} \log_2 \left( 1 + \min (\gamma_{R,2}, \gamma_{2,2}) \right). \tag{6.13}
\]
6.4 Outage Probability

This section investigates the OP for $x_1$ and $x_2$. Therefore, the OP of $x_1$ can be derived as

$$P_{\text{out},1}(\xi_1) = 1 - \Pr[\min(\gamma_{R,1}, \gamma_{2,1}) > \xi_1],$$

(6.14)

where RVs $\gamma_{R,1}$, $\gamma_{2,1}$ and $\gamma_1$ are dependent to each other. Similarly, the OP of $x_2$ can be expressed as

$$P_{\text{out},2}(\xi_2) = 1 - \Pr[\min(\gamma_{R,2}, \gamma_{2,2}) > \xi_2],$$

(6.15)

where RVs $\gamma_{R,2}$ and $\gamma_{2,2}$ are dependent to each other.

Due to the dependence of RVs to each other in (6.14) and (6.15), the analytical derivation of the OP becomes very complicated which is the out of the scope of this Chapter. Hence, the OP for $x_1$ and $x_2$ will be investigated only numerically.

6.5 Numerical Results

In this section, the OP of the proposed system model is evaluated over Rayleigh fading channels. We adopt the following system parameters, unless stated otherwise: $\alpha_j = \beta_j$, $\xi_j = 3$ dB $\forall j \in \{1, 2\}$, $\eta = 0.8$, $\theta = 0.75$, $d_1 = \{1.5d, 3d, 4d\}$, $d_{SP} = d_{SR} = d_2 = d$, $\tau = 2.7$. 

Figure 6.3. The OP of $x_1$ and $x_2$ versus the ransmit SNR threshold at 25 dB when $d_1 = 3d$, $\alpha_1 = 0.8$ and $\alpha_2 = 0.2$. 
Table 6.1. OF-based PA factors for different $d_1$ with EH.

<table>
<thead>
<tr>
<th>$d_1$</th>
<th>1.5$d$</th>
<th>3$d$</th>
<th>4$d$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\alpha_1, \beta_1^*$</td>
<td>0.792</td>
<td>0.885</td>
<td>0.927</td>
</tr>
<tr>
<td>$\alpha_2, \beta_2^*$</td>
<td>0.208</td>
<td>0.115</td>
<td>0.073</td>
</tr>
</tbody>
</table>

Figure 6.4. The OP of $x_1$ and $x_2$ versus PA factors at 20 dB when $d_1 = \{1.5d, 3d, 5d\}$.

In Fig. 6.3, the OP results of $x_1$ and $x_2$ are illustrated. It is considered that $\alpha_1 = \beta_1 = 0.8$, $\alpha_2 = \beta_2 = 0.2$, $d_1 = 3d$. Additionally, the asymptotic case without ITC ($I = \infty$) is also plotted to analyze the OP when SUs can transmit with a maximum transmit power. Moreover, the OP results of the conventional cooperative CR-OMA are plotted to compare that of the proposed cooperative CR-NOMA model. From the plotted results, it is clearly seen that $x_2$ outperforms $x_1$ in terms of the OP results. It can be explained by the fact that $D_2$ implements the SIC to remove the interference from $D_1$, while $D_1$ detects its data without canceling the interference which results in worse the OP results. Moreover, in no ITC case, both messages of the cooperative CR-NOMA obtain better OP compared to that of cooperative CR-OMA for all SNR values. When $I = 30$ dB, a saturation of OP curves for NOMA users began at lower SNRs compared with those of the OMA ones. It is due to the fact that, the transmitted power of $\frac{1}{2}P_S$ at $S$ results in an increase of the ITC value at $PR$.

Fig. 6.4 shows the OF-based values of PA factors for $x_1$ and $x_2$ at 20 dB for different values of $d_1$. It is noted that $x_1$ is in outage for all values of $d_1$ when $u > \frac{\alpha_1}{\alpha_2}$. When $u < \frac{\alpha_1}{\alpha_2}$, an increase of the value of $\alpha_1$ improves the OP of $x_1$ and decrease that of $x_2$. Moreover, it is noticed that when $d_1$ is larger, the OP results get worse. Finally, the obser-
Figure 6.5. The OP of $x_1$ and $x_2$ versus the transmit SNR performance with OF-based PA factors when $d_1 = \{1.5d, 3d, 4d\}$.

Figure 6.6. Harvested power versus the $S$-$R$ distance performance with various $\theta$ values.

Fig. 6.5 plots results for the OP with OF-based PA factors when $d_1 = \{1.5d, 3d, 4d\}$. Noticeably, in all $d_1$ cases, the OP of $x_2$ performs better than that of $x_1$ at SNR levels below 27 dB. However, the OP of $x_1$ obtains better results than that of $x_2$ at higher SNRs.
Moreover, the saturation of the OP curve of \( x_1 \) starts at higher SNRs when \( d_1 \) is larger. It is due to the fact that larger \( \alpha_1 \) is needed when \( D_1 \) is located far away from \( R \). Finally, it can be deduced that the OF-based PA factors improved the OP performance of \( x_1 \) in all \( d_1 \) cases providing fairness for both users.

Fig. 6.6 illustrates the harvested power at \( R \) regarding various distances between \( S \) - \( R \), i.e., \( d_{SR} = \{1d, 2d, 5d, 10d, 20d\} \). It is noted that \( R \) harvest less power when the value of \( d_{SR} \) is higher. It can be explained by the simple fact that the path-loss effect in longer distances degrades the signal power more intensively. Moreover, it is noted that the increase of \( \theta \) allows harvesting much power.

### 6.6 Conclusion

The SWIPT is another candidate technology for the future 5G networks which can maximize energy efficiency. Hence, this chapter studied the SWIPT in dual-hop underlay CR-NOMA networks with the DF relaying over the Rayleigh distribution. The OP for cooperative DF underlay CR-NOMA networks with EH-enabled \( R \) calculated numerically. Moreover, we found OF-based PA factors depending on NOMA users’ channel quality. Thus, numerical results showed that the proper evaluated PA factors can guarantee fairness of both NOMA destination users in terms of the OP. Furthermore, cooperative CR-OMA networks were used as a benchmark to illustrate the better performance of the cooperative CR-NOMA. The comparison of two EH modes, i.e, TS and PS, in Section 3, showed that the PS mode is superior. Therefore, in this Chapter, we studied the PS mode, where it is shown that the increase of \( \theta \) allows harvesting more power, while the increase of \( d_{SR} \) results in less harvested power.
Chapter 7

Conclusions and Future Work

5G is the next major evolution of cellular communication systems with performance improvements of several orders of magnitude over 4G. Hence, the main set of 5G requirements is recognized in the industry as follows: higher spectral and energy efficiency, lower end-to-end latency, massive connectivity, etc. [154]. Due to these high demands on performance improvements in mobile communications, the candidate technologies that can be implemented in 5G are still under consideration. The techniques that jointly studied in this thesis are the potential candidates for 5G. For example, IA-embedded networks can improve the spectral efficiency by mitigating the interference while the CR and NOMA can improve spectrum scarcity and provide massive connectivity. Furthermore, 5G networks can become more energy efficient by applying the SWIPT with different EH protocols.

In this thesis, we investigated the CR networks with EH and NOMA techniques. This concluding chapter summarizes all the key contributions from different chapters and suggests several interesting future research directions.

7.1 Conclusions

Emerging cellular technologies proposed for use in 5G communications will accommodate a wide range of usage scenarios with diverse link requirements. This will necessitate operation over a versatile set of wireless channels, i.e., indoor, outdoor, line-of-sight (LOS) and non-LOS. To bridge the gap between practical and theory channels, Chapter 3 considered the generalized $\alpha-\mu$ fading distribution, which includes several other distributions as special cases, i.e., Rayleigh, Gamma, Exponential, Nakagami-$m$, One-sided Gaussian, and Weibull. This Chapter studied the underlay CR network over generalized $\alpha-\mu$ fading distribution, where the primary receiver applies interference temperature constraint to secondary transmit nodes. Based on the used dual-hop CR network model, the closed-form OP expression for secondary destination node was derived analytically. Furthermore, the impact of the interference temperature constraint on the system outage and SER performance were investigated. From the derived results, we showed that the applied interference constraint degrades the outage and SER performances, which could
be clearly seen from the outage saturation. In addition, Monte Carlo simulations show a consistency with the analytical results, which confirms the correctness of the derived closed-form expression.

Chapter 4 investigated a dual-hop DF wireless powered CR network with IA over Rayleigh fading channels. An introducing CR networks presents two new types of interference, i.e., inter-network interference and intra-network interference. In order to mitigate the negative effect of intra- and inter-network interference terms at the primary and secondary receivers, we used an IA technique by applying precoding and interference suppression beamforming matrices. Here we showed that by using IA technique in underlay cooperative CR networks, interference at primary receivers can be efficiently canceled, and, as a result, secondary transmit nodes are not required to limit their transmit power. Consequently, SNs obtained better system performance in terms of the OP and BER. Moreover, unlike many works in the current literature related to the interference management where unwanted interference has been canceled or removed, we used inter- and intra-network interference for EH purposes to feed the power-limited relay node. In addition, an optimal selection of the EH time fraction in the TSR protocol and the PS fraction in the PSR protocol was found to be important in achieving the best system performance. Moreover, the derived closed-form expressions for the OP of the SN in different CSI mismatch scenarios showed that the PSR protocol outperformed the TSR protocol which validity was confirmed by good agreement with simulation results. Finally, it was shown that the increase in the number of interfering primary nodes allows $R$ to harvest more power, while the increase of interfering nodes’ distance results in lower harvested power.

CR and NOMA will provide more efficient utilization of wireless spectrum in the future. Moreover, the design principle of CR-NOMA networks is perfectly aligned to functional requirements of 5G networks. Thus, Chapter 5 focused on a downlink CR-NOMA model with an imposed interference constraint at the primary receiver. The outage probability was investigated as the main performance metric for the considered system model applying DF and AF relaying methods over Rayleigh and Nakagami-$m$ fading channels. First, the OP was evaluated for the scenario with two secondary users and the OF-based power allocation factors were found for both users to guarantee OP fairness between NOMA users. Then, to attain more practical insight, we considered the effect of the imperfect CSI on the outage performance and derived a general closed-form solution for the outage performance when the number of secondary non-orthogonal nodes was extended to $K$ users. Hence, in this chapter, we showed that CR-NOMA networks not only achieve massive connectivity, but also obtain better outage performance compared to the conventional CR-OMA. Finally, all derived analytical results were verified by Monte-Carlo
simulations, which validated the accuracy of the performance analysis.

Finally, in Chapter 6, we studied the wireless energy EH in underlay CR-NOMA networks with the DF relaying. The OP for cooperative DF underlay CR-NOMA networks with EH-enabled relay calculated numerically. Moreover, we optimized the power allocation factors depending on the NOMA users’ channel quality. Thus, numerical results showed that the properly evaluated power allocation factors can guarantee fairness among NOMA destination users in terms of the OP. Furthermore, cooperative CR-OMA networks were used as a benchmark to illustrate the better performance of the cooperative CR-NOMA. The PS mode showed that the increase of power splitting coefficient allows harvesting more power, while the increase of the source-relay distance results in less harvested power.

7.2 Future Work

Here we suggest some possible research extensions to the work in this thesis as follows.

- In Chapter 3, we studied the outage and SER performance of the underlay CR network considering an ideal transceiver hardware. However, physical transceiver hardware introduces impairments that distort the emitted and received signals in any communication system. Hence, in practice, hardware suffers from various types of impairments, i.e., I/Q imbalance, phase noise and high power amplifier (HPA) nonlinearities [190]. The authors in [191] showed how I/Q imbalance attenuates the amplitude and rotates the phase of the desired constellation, which lead to an SER floor. Hence, as hardware impairments have a deleterious impact on the achievable performance, it would be worthwhile to consider the proposed system model with the aforementioned hardware impairments scenarios.

- Chapter 4 and 6 studied SWIPT technique in CR networks considering linear RF harvesting, which exhibit a linear relationship between input RF and output harvested power. However, rectennas, due to the presence of diodes, exhibit highly nonlinear behavior with limited sensitivity due to the need for bias [192]. Thus, it would be worthwhile to evaluate the RF harvesting model in the thesis considering nonlinearity, sensitivity and saturation of the RF harvesting circuits as well as other facts from the relevant microwave literature.

- In Chapter 5, we investigated CR-NOMA networks assuming that users are able to perform the perfect SIC, and thus completely remove the inter-user interference from the weaker signals. However, in practical systems, the decoding error might be inevitable during the procedure of the SIC, as a result, the interference that are not
removed completely decreases the quality of the received signal [193]. Hence, imperfect SIC has a great impact on the performance of NOMA, especially in massive access systems [194]. Imperfect SIC in NOMA systems has attracted the attention, but is still an open issue. Therefore, studying the proposed CR-NOMA with imperfect SIC is very promising as it can provide extra useful practical insights.
Appendix A

Appendix for Chapter 3

A.1 Proof of Proposition 1

Here, the derivation of (3.21) is provided. Since $|h_{SR}|^2$ and $|h_{SP}|^2$ in (3.14) are independent from each other, (3.14) can be further reformulated by

$$\Lambda_1 = F_{|h_{SR}|^2} \left( \frac{\xi \sigma^2_R}{P_S} \right) \cdot F_{|h_{SP}|^2} \left( \frac{I}{P_S} \right) = \left( 1 - e^{-\frac{\lambda \xi \sigma^2_R}{r_S}} \right) \cdot \left( 1 - e^{-\frac{\lambda I}{r_S}} \right). \quad (A.1)$$

Then, $\Psi$ in (3.15) can be derived as follows

$$\Psi_1 = \int_{\frac{I}{\bar{P}_S}}^{\infty} \int_{\frac{I}{\bar{P}_S}}^{\infty} f_{|h_{SP}|^2}(y) f_{|h_{SR}|^2}(z) \, dz \, dy = \int_{\frac{I}{\bar{P}_S}}^{\infty} \int_{\frac{I}{\bar{P}_S}}^{\infty} \lambda e^{-\lambda y} \left( 1 - e^{-\frac{\lambda I}{r_S}} \right) \, dy$$

$$= \int_{\frac{I}{\bar{P}_S}}^{\infty} \lambda e^{-\lambda y} \, dy - \int_{\frac{I}{\bar{P}_S}}^{\infty} \lambda e^{-\lambda (I + \xi \sigma^2_R)} \, dy = e^{-\frac{\lambda I}{r_S}} - \frac{1}{\bar{P}_S} - I e^{-\frac{\lambda (I + \xi \sigma^2_R)}{r_S}}. \quad (A.2)$$

Then, by substituting (A.1) and (A.2) into (3.13), the CDF of $\varnothing_S$ can be derived in a closed form as

$$F_{\varnothing_S}(\xi) = 1 - e^{-\frac{\lambda \xi \sigma^2_R}{r_S}} + \xi \sigma^2_R e^{-\frac{\lambda (I + \xi \sigma^2_R)}{r_S}}. \quad (A.3)$$

Similarly to $F_{\varnothing_S}(\xi)$, the closed-form expression for $F_{\varnothing_R}(\xi)$ can be derived by following the same steps as in (3.13) - (A.2) and written as

$$F_{\varnothing_R}(\xi) = 1 - e^{-\frac{\lambda \xi \sigma^2_D}{r_R}} + \xi \sigma^2_D e^{-\frac{\lambda (I + \xi \sigma^2_D)}{r_R}}. \quad (A.4)$$

Finally, by using (A.3) and (A.4) and after some mathematical manipulations, the closed-form expression for the OP of $D$ can be computed as in (3.21). ■
A.2 Proof of Proposition 2

From (3.14), the CDFs of $|h_{SR}|^2$ and $|h_{SP}|^2$ can be respectively derived as followings

$$F_{|h_{SR}|^2} (\xi) = \int_0^{\xi} \alpha_{SR} \mu_{SR} \gamma_{\frac{\alpha_{SR} \mu_{SR}}{2}} e^{-\mu_{SR} \left( \frac{\xi}{\gamma} \right)^{\frac{\alpha_{SR}}{2}}} \frac{d\xi}{2\Gamma(\mu_{SR})} = \frac{\gamma_{\text{inc}} \left( \mu_{SR}, \mu_{SR} \left( \frac{\xi}{\gamma} \right)^{\frac{\alpha_{SR}}{2}} \right)}{\Gamma(\mu_{SR})}$$  \quad (A.5)

and

$$F_{|h_{SP}|^2} = \int_0^{I_{\gamma P}} \alpha_{SP} \mu_{SP} \gamma_{\frac{\alpha_{SP}}{2}} e^{-\mu_{SP} \left( \frac{I}{\gamma P} \right)^{\frac{\alpha_{SP}}{2}}} \frac{dI}{2\Gamma(\mu_{SP})} = \frac{\gamma_{\text{inc}} \left( \mu_{SP}, \mu_{SP} \left( \frac{I}{\gamma P} \right)^{\frac{\alpha_{SP}}{2}} \right)}{\Gamma(\mu_{SP})} \quad \gamma_{\text{inc}} \left( \mu_{SR}, \mu_{SR} \left( \frac{\xi y \sigma_R^2 R}{\gamma I P} \right)^{\frac{\alpha_{SR}}{2}} \right) \Gamma(\mu_{SR})} \quad \gamma_{\text{inc}} \left( \mu_{SP}, \mu_{SP} \left( \frac{I}{\gamma P} \right)^{\frac{\alpha_{SP}}{2}} \right) \Gamma(\mu_{SP})} \quad (A.6)

Then, by using (A.5) and (A.6), (3.14) can be further rewritten as

$$\Lambda_2 = \frac{\gamma_{\text{inc}} \left( \mu_{SR}, \mu_{SR} \left( \frac{\xi \sigma_R^2 R}{\gamma I P} \right)^{\frac{\alpha_{SR}}{2}} \right) \gamma_{\text{inc}} \left( \mu_{SP}, \mu_{SP} \left( \frac{I}{\gamma P} \right)^{\frac{\alpha_{SP}}{2}} \right)}{\Gamma(\mu_{SR}) \Gamma(\mu_{SP})} \quad \gamma_{\text{inc}} \left( \mu_{SR}, \mu_{SR} \left( \frac{\xi \sigma_R^2 R}{\gamma I P} \right)^{\frac{\alpha_{SR}}{2}} \right) \Gamma(\mu_{SR})} \Gamma(\mu_{SP})} = \frac{\gamma_{\text{inc}} \left( \mu_{SR}, \mu_{SR} \left( \frac{\xi y \sigma_R^2 R}{\gamma I P} \right)^{\frac{\alpha_{SR}}{2}} \right) \Gamma(\mu_{SR})} \Gamma(\mu_{SP})} \quad (A.7)

Further, $\Psi$ in (3.15) can be computed by

$$\Psi_2 = \int_0^{\infty} f_{|h_{SR}|^2} (y) \int_0^{\infty} f_{|h_{SR}|^2} (z) \frac{dz}{2\Gamma(\mu_{SR})} \frac{d\mu_{SR}}{2\Gamma(\mu_{SP})} e^{-\mu_{SR} \left( \frac{\xi y \sigma_R^2 R}{\gamma I P} \right)^{\frac{\alpha_{SR}}{2}}} \gamma_{\text{inc}} \left( \mu_{SR}, \mu_{SR} \left( \frac{\xi y \sigma_R^2 R}{\gamma I P} \right)^{\frac{\alpha_{SR}}{2}} \right) \Gamma(\mu_{SR}) \Gamma(\mu_{SP})} \quad \gamma_{\text{inc}} \left( \mu_{SR}, \mu_{SR} \left( \frac{\xi y \sigma_R^2 R}{\gamma I P} \right)^{\frac{\alpha_{SR}}{2}} \right) \Gamma(\mu_{SR})} \quad (A.8)

Now, by applying the series expansion of an incomplete gamma function [110]

$$\gamma_{\text{inc}} (n, x) = \Gamma (n) - \Gamma (n) e^{-x} \sum_{m=0}^{n-1} \frac{x^m}{m!} \quad (A.9)$$
Ψ_2 can be re-expressed in a closed form as

\[
Ψ_2 = \frac{\Gamma\left(\mu_{SP}, \mu_{SP}\left(\frac{l}{\gamma P_S}\right)^{\alpha_{SP}/2}\right)}{\Gamma(\mu_{SP})} - \frac{\mu_{SP}^{\mu_{SP}}}{\Gamma(\mu_{SP} \gamma_{SP}^{\alpha_{SP}/2})} \\
\times \sum_{i=0}^{\mu_{SR}-1} \frac{\left(\mu_{SR}\left(\frac{\xi \sigma^2_R}{\gamma l}\right)^{\alpha_{SR}/2}\right)^i}{i! \left(\frac{\mu_{SP}}{\gamma \alpha_{SP}/2} + \mu_{SR}\left(\frac{\xi \sigma^2_R}{\gamma l}\right)^{\alpha_{SR}/2}\right)^{\mu_{SP}+i}} \\
\times \Gamma\left(\mu_{SP} + 1, \left(\frac{I}{P_S}\right)^{\alpha_{SR}/2} \left[\frac{\mu_{SP}}{\gamma \alpha_{SP}/2} + \mu_{SR}\left(\frac{\xi \sigma^2_R}{\gamma l}\right)^{\alpha_{SR}/2}\right]\right),
\]  

(A.10)

where \(\Gamma(s, x) = \int_x^\infty t^{s-1}e^{-t}dt\) denotes the upper incomplete gamma function. Further, by summing (A.7) and (A.10), the CDF of \(\mathcal{S}\) for the \(\alpha - \mu\) distribution can be written in a closed form as in (3.26) Then, by following the similar approach as in deriving the CDF of \(\mathcal{S}\), the closed-form expression for the CDF of \(\mathcal{R}\) for the \(\alpha - \mu\) distribution can be expressed as in (3.27)
Appendix B

Appendices for Chapter 4

B.1 Proof of Proposition 3

Eq. (4.41) can be extended as

\[
P_{out}^{(i)} = \Pr \left[ \frac{1}{2} \log_2 \left( 1 + \frac{P_j |h_{j,j}^{(i)}|^2}{\sigma_j^2} \right) < R_{th} \right] = \Pr \left[ |h_{j,j}^{(i)}|^2 < \frac{\xi d_{j,j}^{\nu_j} \sigma_j^2}{P_j} \right].
\]  

(B.1)

By setting the rate threshold \( R_{th} \) to positive value, the OP of \( |h_{j,j}^{(i)}|^2 \) can be found from the following cumulative distribution function (CDF)

\[
\Pr \left[ |h_{j,j}^{(i)}|^2 \leq r_j^{(i)} \right] = 1 - e^{-r_j^{(i)}}, \quad r_j^{(i)} \geq 0;
\]  

(B.2)

where \( r_j^{(i)} = \frac{\xi d_{j,j}^{\nu_j} \sigma_j^2}{P_j} \).

Finally, the OP of \( PR_j \) can be derived from (B.2) as (4.43).

B.2 Proof of Proposition 4

The OP for relaying SN can be found by using the end-to-end capacity at \( D \). Thus, (4.44) can be further extended as

\[
P_{out}^{(D)} = \Pr \left[ \min(\gamma_R, \gamma_D) < \xi_D \right] = 1 - \Pr \left[ \min(\gamma_R, \gamma_D) > \xi_D \right]
= 1 - \Pr \left[ \gamma_R > \xi_D \right] \Pr \left[ \gamma_D > \xi_D \right]
= 1 - \Pr \left[ \frac{(1 - \rho) P_S |h_{R,S}|^2}{\sigma_{R,S}^2 P_j} > \xi_D \right] \Pr \left[ \frac{P_{PSR} |h_{D,R}|^2}{\sigma_{D,R}^2 P_j} > \xi_D \right]
= 1 - \Pr \left[ |h_{R,S}|^2 > \frac{\xi_D \sigma_{R,S}^2}{(1 - \rho) P_j} \right] \Pr \left[ |h_{D,R}|^2 > \frac{\xi_D \sigma_{D,R}^2}{P_{PSR}^j} \right].
\]  

(B.3)
where $\xi^D = 2^{2R_D} - 1$ is the corresponding SNR threshold at $D$ while $P_R^{PSR}$ is the harvested energy at $R$ given in (4.30). As $|h_{j,j}^{[i]}|^2$ in Appendix B.1, $|h_{R,S}|^2$ and $|h_{D,R}|^2$ also follow exponential distribution with CDFs

$$
\Pr[|h_{R,S}|^2 \leq r_R] = 1 - e^{-r_R}, \quad r_R \geq 0, \quad (B.4a)
$$

$$
\Pr[|h_{D,R}|^2 \leq r_D] = 1 - e^{-r_D}, \quad r_D \geq 0, \quad (B.4b)
$$

respectively, where $r_R = \frac{\xi^D \sigma_{R,S}^2 \tilde{\sigma}_{R}^2}{(1 - \rho)P_S}$ and $r_D = \frac{\xi^D \sigma_{D,R}^2 \tilde{\sigma}_{D}^2}{P_R^{PSR}}$. Hence, inserting (B.4) into (B.3) gives the OP of $D$ as (4.45).
Appendix C

Appendices for Chapter 5

C.1 Proof of Proposition 5

The CDFs of $\gamma_2$ and $\gamma_1$ are derived using (5.9) and (5.11) as

$$F_{\gamma_2}(\xi_1) = \Pr \left[ |h_2|^2 < \frac{\xi_1 d_2^2}{k \rho_R} \right] = \begin{cases} 1 - e^{-\frac{\xi_1 d_2^2}{2k \rho_R}}, & \xi_1 < \frac{\beta_1}{\beta_2}, \\ 1, & \text{otherwise,} \end{cases}$$  \hspace{1cm} (C.1)

and

$$F_{\gamma_1}(\xi_1) = \Pr \left[ |h_1|^2 < \frac{\xi_1 d_1^2}{k \rho_R} \right] = \begin{cases} 1 - e^{-\frac{\xi_1 d_1^2}{2k \rho_R}}, & \xi_1 < \frac{\beta_1}{\beta_2}, \\ 1, & \text{otherwise,} \end{cases}$$  \hspace{1cm} (C.2)

where $\kappa = \beta_1 - \beta_2 \xi_1$ while $|h_1|^2$ and $|h_2|^2$ follow exponential distribution. Then, the CDF of the RV $\gamma_{R,1}$ can be derived as

$$F_{\gamma_{R,1}}(\xi_1) = \Pr \left[ \frac{|h_{SR}|^2 \alpha_1 \bar{\rho}_S}{|h_{SR}|^2 \alpha_1 \bar{\rho}_S + d_{SR}^2} < \xi_1, \bar{\rho}_S < \frac{\rho_I d_{SP}^2}{|h_{SP}|^2} \right]$$

$$+ \Pr \left[ \frac{|h_{SR}|^2 \alpha_2 \bar{\rho}_S}{|h_{SR}|^2 \alpha_2 \bar{\rho}_S + d_{SR}^2} < \xi_1, \bar{\rho}_S > \frac{\rho_I d_{SP}^2}{|h_{SP}|^2} \right]$$

$$= \Pr \left[ \frac{|h_{SR}|^2 < \xi_1 d_{SR}^2}{\psi \bar{\rho}_S}, |h_{SP}|^2 < \frac{\rho_I d_{SP}^2}{\bar{\rho}_S} \right]$$

$$A_1 + \Pr \left[ \frac{|h_{SR}|^2 < \xi_1 d_{SR}^2}{\psi \rho_I}, |h_{SP}|^2 > \frac{\rho_I d_{SP}^2}{\bar{\rho}_S} \right], \hspace{1cm} (C.3)$$

where $|h_{SP}|^2$ and $|h_{SR}|^2$ also follow exponential distributions, $\psi = \alpha_1 - \alpha_2 \xi_1$, $\bar{\rho}_S = \frac{P_S}{\sigma^2}$ stands for the maximum allowed transmit SNR at $S$ and $\rho_I = \frac{T}{\sigma^2}$ denotes the temperature-constraint-to-noise ratio at $PR$. Now, in the term $A_1$ of (C.3), $|h_{SP}|^2$ and $|h_{SR}|^2$ are
Regarding the term \( B_1 \) in (C.3), it can be further extended as

\[
B_1 = \int_{\bar{\rho}_S}^{\infty} \int_{\bar{\rho}_S}^{\infty} f_{\gamma | \rho_S} f_{\gamma | \rho_s} d\gamma d\rho_s = \int_{\bar{\rho}_S}^{\infty} \frac{1}{2} e^{-\frac{\rho_s d_S^r}{2\rho_S}} \left( 1 - e^{-\frac{\rho_s d_S^r}{2\rho_S}} \right) dy = e^{-\frac{\rho_s d_S^r}{2\rho_S}} - \frac{\psi \rho_s d_s^r e^{-\frac{\rho_s d_s^r}{2\rho_S}}}{\psi \rho_s d_s^r + \xi_1 d_s^r}.
\]

Then, after substituting (C.4) and (C.5) into (C.3) and after some algebraic manipulations, the closed-form expression for (C.3) can be written as

\[
F_{\gamma_{R_1}}(\xi_1) = \begin{cases} 
1 - e^{-\frac{\xi_1 d_S^r}{2\rho_S}} + \frac{\xi_1 d_S^r e^{-\frac{\rho_s d_S^r}{2\rho_S}}}{\psi \rho_s d_s^r + \xi_1 d_s^r}, & \xi_1 < \frac{\alpha_1}{\alpha_2}, \\
1, & \text{otherwise}.
\end{cases}
\]

Finally, the closed-form expression for the OP of \( x_1 \) can be derived by substituting (C.1), (C.2) and (C.6) into (5.22), which can be written as in (5.23).

**C.2 Proof of Proposition 6**

The CDF of the RV \( \gamma_{2,2} \) by using (5.10) is further expressed by

\[
F_{\gamma_{2,2}}(\xi_2) \triangleq F_{\gamma|\rho_2}(\xi_2) = \Pr \left[ \frac{|h_2|^2 \beta_2 \rho_R}{d_2^r} < \xi_2 \right] = \Pr \left[ \frac{|h_2|^2}{\beta_2 \rho_R} < \frac{\xi_2 d_2^r}{\beta_2 \rho_R} \right] = 1 - e^{-\frac{\xi_2 d_2^r}{\beta_2 \rho_R}}.
\]

Furthermore, the CDF of the RV \( \gamma_{R,2} \) is expressed by

\[
F_{\gamma_{R,2}}(\xi_2) = \Pr \left[ \frac{|h_{SR}|^2 \alpha_2 \bar{\rho}_S}{d_{SR}^2} < \xi_2, \bar{\rho}_S < \frac{\rho_s d_s^r}{|h_{SP}|^2} \right] + \Pr \left[ \frac{|h_{SR}|^2 \alpha_2 \rho_I}{|h_{SP}|^2 d_{SR}^2} < \xi_2, |h_{SP}|^2 > \frac{\rho_s d_s^r}{\bar{\rho}_S} \right] = \Pr \left[ \frac{|h_{SR}|^2}{|h_{SP}|^2} < \frac{\xi_2 d_{SR}^r}{\alpha_2 \bar{\rho}_S}, |h_{SP}|^2 < \frac{\rho_s d_s^r}{\bar{\rho}_S} \right] + \Pr \left[ \frac{|h_{SR}|^2}{|h_{SP}|^2} < \frac{\xi_2 d_{SR}^r}{\alpha_2 \rho_I}, |h_{SP}|^2 > \frac{\rho_s d_s^r}{\bar{\rho}_S} \right],
\]

\[
(C.8)
\]
where the term $C_1$ can be reformulated as followings

$$C_1 = F_{|h_{SR}|^2} \left( \frac{\xi_2 d_{SR}^{e}}{\sigma_2 \hat{\rho}_S} \right) F_{|h_{SP}|^2} \left( \frac{\rho_1 d_{SP}^{e}}{\hat{\rho}_S} \right) = \left( 1 - e^{-\frac{\xi_2 d_{SR}^{e}}{2\sigma_2 \hat{\rho}_S}} \right) \left( 1 - e^{-\frac{\rho_1 d_{SP}^{e}}{2\hat{\rho}_S}} \right).$$ \hspace{1cm} (C.9)

Then, the term $D_1$ in (C.8) is extended as follows

$$D_1 = \int_{\rho_1 d_{SP}^{e} \hat{\rho}_S}^{\infty} f_{|h_{SR}|^2}(y) \int_{0}^{\infty} f_{|h_{SP}|^2}(z) dz dy = \int_{\rho_1 d_{SP}^{e} \hat{\rho}_S}^{\infty} 1 - e^{-\frac{\xi_2 d_{SR}^{e}}{2\sigma_2 \hat{\rho}_S}} dy \cdot \frac{\alpha_1 \rho_1 d_{SP}^{e} - \alpha_2 \rho_1 d_{SP}^{e} + \xi d_{SR}^{e}}{\alpha_2 \rho_1 d_{SP}^{e} + \xi d_{SR}^{e}}.$$ \hspace{1cm} (C.10)

Now, (C.9) and (C.10) can be substituted into (C.8), then after some algebraic manipulations, (C.8) can be written by

$$F_{z_{SR}}(\xi_2) = 1 - e^{-\frac{\xi_2 d_{SR}^{e}}{2\sigma_2 \hat{\rho}_S}} + \frac{\xi_2 d_{SR}^{e} e^{-\frac{\alpha_1 \rho_1 d_{SP}^{e} + \xi d_{SR}^{e}}{2\sigma_2 \hat{\rho}_S}}}{\alpha_2 \rho_1 d_{SP}^{e} + \xi d_{SR}^{e}}.$$ \hspace{1cm} (C.11)

Finally, the OP in (5.24), after inserting (C.7) and (C.11), can be written in a closed-form as in (5.25).

### C.3 Proof of Proposition 7

Considering the ITC, the CDF in (5.26), which is also the OP of $D_1$, can be derived as

$$P_{\text{out},1}^{\text{AF}}(\xi_1) = \text{Pr} \left\{ \eta_1 Y_{n_1} X_{n_1} \frac{\alpha_1}{Y_1 \eta_1 \alpha_1 + n_1 X + n_1 Y + 1} < \xi_1, \hat{\rho}_S < \frac{\rho_1 d_{SP}^{e}}{Z} \right\}
+ \text{Pr} \left\{ \eta_1 Y_{n_1} X_{n_1} \frac{Y_1 \eta_1 \alpha_1}{Z_1 \eta_1 \alpha_1} + n_1 X + n_1 Y + 1 < \xi_1, \hat{\rho}_S > \frac{\rho_1 d_{SP}^{e}}{Z} \right\}
= \text{Pr} \left\{ \frac{n_1 X (n_1 Y (\alpha_1 - \alpha_2 \xi_1) - \xi_1)}{Z_1 \eta_1 \alpha_1} < \eta_1 Y_{n_1} X_{n_1} \frac{\alpha_1}{Y_1 \eta_1 \alpha_1 + n_1 X + n_1 Y + 1} < \xi_1, Z < \frac{\rho_1 d_{SP}^{e}}{\hat{\rho}_S} \right\}
+ \text{Pr} \left\{ \frac{n_1 X (n_1 Y (\alpha_1 - \alpha_2 \xi_1) - \xi_1)}{Z_1 \eta_1 \alpha_1} < \frac{n_1 Y_1 \eta_1 \alpha_1}{Z_1 \eta_1 \alpha_1} + \xi_1, Z > \frac{\rho_1 d_{SP}^{e}}{\hat{\rho}_S} \right\}
= \text{Pr} \left\{ Y < \frac{\xi_1}{n_1 \eta_1} \right\} \underset{A}{\text{Pr}} \left\{ X < \frac{Y_{n_1} \eta_1 u_1 + u_1}{n_1 (Y - u_1)} \right\}
+ \text{Pr} \left\{ \frac{Z_1 \eta_1 \alpha_1}{n_1 \eta_1 \alpha_1} < \frac{Z \eta_1 \alpha_1}{n_1 \eta_1 \alpha_1} \right\} \underset{B}{\text{Pr}} \left\{ X < \frac{Y_{n_1} \eta_1 u_1 + Z \eta_1 u_1}{n_1 (Y - Z \eta_1 u_1)} \right\}.$$ \hspace{1cm} (C.12)

where $n_p = \frac{\rho_1 d_{SP}^{e}}{d_{SR}^{e}}$, $c = \frac{\rho_1 d_{SP}^{e}}{\hat{\rho}_S}$, $\rho_1 = \frac{\rho_1 d_{SP}^{e}}{\sigma_2 \hat{\rho}_S}$ and $Z = |h_{SP}|^2$. Moreover, the step $\epsilon$ in (C.12) relies on $\alpha_1 > \alpha_2 \xi_1$, otherwise, $P_{\text{out},1}^{\text{AF}}(\xi_1) = 1$ regardless of the system SNR. Then, we
can further rewrite the term $A$ in (C.12) as

$$
A = \int_0^c f_Z(z)dz \int_0^{u_1} f_Y(y)dy + \int_0^c f_Z(z)dz \int_{u_1}^\infty F_X\left(\frac{yn_u + u_1}{n_1(y - u_1)}\right) f_Y(y)dy.
$$

(C.13)

As the PDFs in the term $A_1$ are independent from each other, we can write it as

$$
A_1 = \int_0^c f_Z(z)dz \int_0^{u_1} f_Y(y)dy = (1 - e^{-c})(1 - e^{-u_1}),
$$

(C.14)

and $A_2$ is written as

$$
A_2 = \int_0^c f_Z(z)dz = (1 - e^{-c}).
$$

(C.15)

Then, the term $A_3$ can be further extended as

$$
A_3 = \int_{u_1}^\infty e^{-y}dy - \int_{u_1}^\infty e^{-\frac{yn_u + u_1}{n_1(y-u_1)}}y dy.
$$

(C.16)

The term $A_{31}$ in (C.16) can be further rewritten as

$$
A_{31} = \int_{u_1}^\infty e^{-y}dy = e^{-u_1}.
$$

(C.17)

Moreover, by using a substitution of $t_1 = y - u_1$, $A_{32}$ in (C.16) can be reformulated by

$$
A_{32} = \int_0^\infty e^{-\frac{t_1}{n_1} + \frac{u_1}{n_1} + \frac{u_1}{n_1(t_1-u_1)}} - t_1 - u_1 dt_1 = e^{-\frac{u_1}{n_1}} \int_0^\infty e^{-\frac{u_1}{n_1} + \frac{u_1}{n_1} + \frac{u_1}{n_1}} \frac{1}{n_1} - t_1 dt_1.
$$

(C.18)

Then, using [121, Eq. (3.324.1)], $A_{32}$ can be further derived as

$$
A_{32} = e^{-\frac{u_1}{n_1}} \frac{2}{n_1} \sqrt{u_1(n_u u_1 + 1)} K_1\left(\frac{2}{n_1} \sqrt{u_1(n_u u_1 + 1)}\right),
$$

(C.19)
where $K_1(\cdot)$ is the modified Bessel function of the second kind of order 1. Further, the term $A$, after substituting (C.14), (C.15), (C.17) and (C.19) into (C.13) is reformulated as

$$A = (1 - e^{-c})(1 - e^{-u_1}) + (1 - e^{-c})$$

$$\times \left( e^{-u_1} - e^{-\frac{u_1(n_1 + n_s)}{n_1}} \right) \left( \sqrt{\frac{u_1(n_u + 1)}{n_1}} K_1 \left( 2 \sqrt{\frac{u_1(n_u + 1)}{n_1}} \right) \right)$$

$$= (1 - e^{-c}) \left( 1 - e^{-\frac{u_1(n_1 + n_s)}{n_1}} \right) 2\sqrt{r_1} K_1 \left( 2\sqrt{r_1} \right), \quad \text{(C.20)}$$

where $r_1 = \frac{u_1(n_u + 1)}{n_1}$.

Now, the term $B$ in (C.12) can be further extended as

$$B = \int_c^\infty f_Z(z) \int_{z \bar{u}_1}^\infty f_Y(y) dy dz + \int_c^\infty f_Z(z) \int_{z \bar{u}_1}^\infty f_Y(y) f_X \left( \frac{n_p y \bar{u}_1 + z \bar{u}_1}{n_1 (y - z \bar{u}_1)} \right) dy dz \bigg|_{B_1}^{B_2}.$$  

The term $B_1$ can be derived as

$$B_1 = \int_c^\infty e^{-c} \int_0^{z \bar{u}_1} e^{-y} dy dz = \int_c^\infty e^{-c} \left( 1 - e^{-z \bar{u}_1} \right) dz$$

$$= \int_c^\infty e^{-c} dz - \int_c^\infty e^{-z(\bar{u}_1 + 1)} dz = e^{-c} - \frac{e^{-c(\bar{u}_1 + 1)}}{\bar{u}_1 + 1}. \quad \text{(C.22)}$$

Further, the term $B_2$ can be rewritten as

$$B_2 = \int_c^\infty e^{-c} \left( \int_{z \bar{u}_1}^\infty e^{-y} dy \right) - \left( \int_{z \bar{u}_1}^\infty e^{-\frac{n_p y \bar{u}_1 + z \bar{u}_1}{n_1(y - z \bar{u}_1)}} dy \right) dz. \quad \text{(C.23)}$$

Then, by using $p_1 = y - z \bar{u}_1$ and [121, Eq. (3.324.1)], the term $B_{21}$ in (C.23) is written as

$$B_{21} = \int_0^\infty e^{-\frac{u_1(n_p z + n_1)}{n_1 p_1}} p_1 d p_1 = e^{-\frac{u_1(n_p + n_1)}{n_1}} \int_0^\infty e^{-\frac{z \bar{u}_1(n_p + n_1)}{n_1 p_1}} p_1 d p_1$$

$$= e^{-\frac{u_1(n_p + n_1)}{n_1}} 2\sqrt{r_1} K_1 \left( 2\sqrt{r_1} \right), \quad \text{(C.24)}$$
where $s_1 = \frac{z u_1(n_p u_1 + 1)}{n_1}$. Now, we rewrite (C.23) using (C.24) as

$$B_2 = \int_c^\infty e^{-z \bar{u}_1} - e^{-\bar{u}_1 z} dz - \int_c^\infty e^{-\frac{z u_1(n_p u_1 + 1)}{n_1}} 2 \sqrt{s_1} K_1(2 \sqrt{s_1}) dz. \tag{C.25}$$

Then, substituting $q = z - c$ and after some algebraic manipulations, the term $B_{22}$ in (C.25) can be rewritten as

$$B_{22} = e^{-c} \frac{n_1 (\bar{u}_1 + 1) + n_p \bar{u}_1}{\bar{u}_1 (n_p \bar{u}_1 + 1)} \sum_{k=0}^{\infty} \left( -\frac{c \bar{u}_1}{n_1} \right)^k \frac{1}{k!} G_{3}^{1,2} \left[ \begin{array}{c} n_1 (\bar{u}_1 + 1) \n_1 \bar{u}_1 + 1 \end{array} \right| \begin{array}{c} 0, k - 1, k \hline 0, k \end{array} \right] \tag{C.26}$$

where $l_1 = \frac{q u_1(n_p u_1 + 1) + c u_1(n_p u_1 + 1)}{n_1}$. Further, after applying [195, Eqs. (12) and (14)] and [196, Eq. (24.1.3)] to $\Lambda$ in (C.26) and by substituting (C.22) - (C.25) into (C.21), the term $B$ can be written in a closed-form expression as

$$B = e^{-c} - \frac{n_1 e^{-\frac{c (\bar{u}_1 + 1) + n_p \bar{u}_1}{\bar{u}_1 (n_p \bar{u}_1 + 1)}}}{\bar{u}_1 (n_p \bar{u}_1 + 1)} \sum_{k=0}^{\infty} \left( -\frac{c \bar{u}_1}{n_1} \right)^k \frac{1}{k!} G_{3}^{1,2} \left[ \begin{array}{c} n_1 (\bar{u}_1 + 1) \n_1 \bar{u}_1 + 1 \end{array} \right| \begin{array}{c} 0, k - 1, k \hline 0, k \end{array} \right] \tag{C.27}$$

Finally, by substituting (C.20) and (C.27) into (C.12), the exact OP for $D_1$ can be written as in (5.27).

### C.4 Proof of Proposition 8

Using (5.19), the OP for $D_2$ can be further rewritten as

$$P_{\text{out},2}^A (\xi_2) = \text{Pr} \begin{cases} \frac{n_s Y n_Q \alpha_2}{n_2 Z + n_s Y + 1} < \xi_2, \bar{\rho}_S < \frac{\rho_{1}}{Z} \\ \frac{n_p Y n_Q \alpha_2}{n_2 Z + n_p Z + Y + 1} < \xi_2, \bar{\rho}_S > \frac{\rho_{1}}{Z} \end{cases}$$

$$= \text{Pr} \left[ n_2 Q (n_s Y \alpha_2 - \xi_2) < n_s Y \xi_2 + \xi_2, Z < c \right] + \text{Pr} \left[ n_2 Q \left( \frac{n_p Y \alpha_2 - \xi_2}{Z} \right) < \frac{n_p Y \xi_2 + \xi_2}{Z}, Z > c \right]$$

$$= \text{Pr} \left[ Y < \frac{\xi_2}{n_s \alpha_2} \triangleq u_2, Q < \frac{Y n_s u_2 + u_2}{n_2 (Y - u_2)}, Z < c \right]$$
The OP derivation of $D_2$ from (C.28) can be derived by following the same step as in (C.13) - (C.27) shown in Appendix (C.3), which is omitted for the sake of brevity. Hence, the OP expression for $D_2$ is written as in (5.29).

**C.5 Proof of Proposition 9**

First, by using the “change of variable” method [115, Eq. (14)], we re-express (5.30) as

$$f_{|h|^2}(r) = \frac{\lambda_t^{m_t}}{\Gamma(m_t)} r^{m_t-1} \exp(-\lambda_t r), \quad g_t > 0,$$  

(C.29)

where $m_t \geq \frac{1}{2}$ and $\hat{r}_t$ stand for the Nakagami-$m$ fading shaping parameter and average channel power gain, accordingly. Now, the CDF from (C.29) can be derived by

$$F_{|h|^2}(r) = 1 - \frac{\Gamma(m_t, \lambda_t r)}{\Gamma(m_t)}, \quad \forall t \in \{SR, RP, 1, 2\}.$$  

(C.30)

Then, using (C.30), the CDF of $\gamma_{R,1}$ can be derived as follows

$$F_{\gamma_{R,1}}(\xi_1) = \Pr \left[ |h_{SR}|^2 < \xi_1 \frac{d_{SR}^r}{a_1 \rho_S} \right] \triangleq F_{|h_{SR}|^2}(\xi_1) = 1 - \frac{\Gamma \left( m_{SR}, \frac{\lambda_{SR} \xi_1 d_{SR}}{a_1 \rho_S} \right)}{\Gamma(m_{SR})}. \quad (C.31)$$

Further, the CDF of $\gamma_{1,1}$ can be written as

$$F_{\gamma_{1,1}}(\xi_1) = \Pr \left[ \frac{|h_1|^2}{b_1^2} \frac{d_{1R}^r}{d_1^r} < \xi_1, \frac{\rho_{1R} d_{1R}^r}{|h_{1R}|^2} \right] + \Pr \left[ \frac{|h_1|^2}{b_1^2} \frac{d_{1R}^r}{d_1^r} < \xi_1, \frac{\rho_{1R} d_{1R}^r}{|h_{1R}|^2} > \frac{\rho_{1R} d_{1R}^r}{\tilde{\rho}_R} \right]$$

$$= \Pr \left[ \frac{|h_1|^2}{b_1^2} \frac{d_{1R}^r}{d_1^r} < \xi_1, \frac{|h_{1R}|^2}{\tilde{\rho}_R} < \frac{\rho_{1R} d_{1R}^r}{\tilde{\rho}_R} \right]$$

$$+ \Pr \left[ \frac{|h_1|^2}{b_1^2} \frac{d_{1R}^r}{d_1^r} < \xi_1, \frac{|h_{1R}|^2}{\tilde{\rho}_R} > \frac{\rho_{1R} d_{1R}^r}{\tilde{\rho}_R} \right], \quad (C.32)$$
Then, applying the series expansion of the lower incomplete Gamma function

where \( \tilde{\rho}_R = \frac{\rho_R}{\sigma_2} \). Now, as \( |h_1|^2 \) and \( |h_{RP}|^2 \) in \( N_1 \) of (C.32) are independent RVs, \( N_1 \) can be recalculated by

\[
N_1 = \int_0^{\rho_{id_{RP}}} f_{|h_{RP}|^2}(x) \int_0^{\frac{\xi_1 d_1^R}{\rho_{id_{RP}}}} f_{|h_1|^2}(y) \, dy
\]

\[
= \left( 1 - \frac{\Gamma(m_{RP}, \frac{\lambda_{RP} \rho_{id_{RP}}}{\tilde{\rho}_R})}{\Gamma(m_{RP})} \right) \left( 1 - \frac{\Gamma(m_1, \frac{\lambda_1 \xi_1 d_1^R}{\tilde{\rho}_R})}{\Gamma(m_1)} \right).
\] (C.33)

Then, \( N_2 \) in (C.32) using (C.30) can be further extended as

\[
N_2 = \int_{\rho_{id_{RP}}}^{\infty} f_{|h_{RP}|^2}(x) \int_0^{\frac{\xi_1 d_1^R}{\rho_{id_{RP}}}} f_{|h_1|^2}(y) \, dy \, dx
\]

\[
= \int_{\rho_{id_{RP}}}^{\infty} \lambda_{m_{RP}}^{m_{RP}} x^{m_{RP}-1} \exp \left( -\lambda_{RP} x \right) \frac{\gamma_{inc}(m_{RP}, \frac{\lambda_1 \xi_1 d_1^R}{\rho_{id_{RP}}})}{\Gamma(m_{RP}) \Gamma(m_1)} \, dx,
\] (C.34)

where \( \gamma_{inc}(s, x) = \int_0^x t^{s-1} e^{-t} \, dt \) denotes the lower incomplete Gamma function [121]. Then, applying the series expansion of the lower incomplete Gamma function \( \gamma_{inc}(s, x) = \Gamma(s) \left( 1 - e^{-x} \sum_{i=0}^{s-1} \frac{x^i}{i!} \right) \) as in [161] and after some algebraic manipulations, (C.34) can be further expressed by followings

\[
N_2 = \frac{\Gamma(m_{RP}, \frac{\lambda_{RP} \rho_{id_{RP}}}{\tilde{\rho}_R})}{\Gamma(m_{RP})} - \frac{\lambda_{m_{RP}}^{m_{RP}}}{\Gamma(m_{RP})} \sum_{i=0}^{m_{RP}-1} \left( \frac{\lambda_1 \xi_1 d_1^R}{\rho_{id_{RP}}} \right)^i \frac{\Gamma(m_{RP} + i, \frac{\rho_{id_{RP}} c_1}{\tilde{\rho}_R})}{\Gamma(m_{RP})},
\] (C.35)

where \( c_1 = \lambda_{RP} + \frac{\lambda_1 \xi_1 d_1^R}{\rho_{id_{RP}}} \). Hence, the CDF of \( \gamma_{1,1} \) given in (C.32) can be rewritten in a closed-form as follows

\[
F_{\gamma_{1,1}}(\xi_1) = 1 - \left( 1 - \frac{\Gamma(m_{RP}, \frac{\lambda_{RP} \rho_{id_{RP}}}{\tilde{\rho}_R})}{\Gamma(m_{RP})} \right) \frac{\Gamma(m_1, \frac{\lambda_1 \xi_1 d_1^R}{\rho_{id_{RP}}})}{\Gamma(m_1)}
\]

\[
- \sum_{i=0}^{m_{RP}-1} \left( \frac{\lambda_1 \xi_1 d_1^R}{\rho_{id_{RP}}} \right)^i \frac{\lambda_{m_{RP}}^{m_{RP}}}{\Gamma(m_{RP})} \frac{\Gamma(m_{RP} + i, \frac{\rho_{id_{RP}} c_1}{\tilde{\rho}_R})}{\Gamma(m_{RP})}.
\] (C.36)

Finally, by substituting (C.31) and (C.36) into (5.42), a closed-form expression for the OP of \( x_1 \) can be expressed as in (5.43).
C.6 Proof of Proposition 10

By using (5.33), the CDF of $\gamma_{R,2}$ is derived as follows

$$F_{\gamma_{R,2}}(\xi_2) = \Pr \left[ \frac{\xi_2 d_{SR}}{\rho_S (a_2 - a_1 \xi_2)} < |h_{SR}|^2 \right] \triangleq F_{|h_{SR}|^2}(\xi_2) = \begin{cases} 1 - \frac{\Gamma \left( m_{SR}, \frac{\lambda_{SR} d_{SR}}{\rho_S (a_2 - a_1 \xi_2)} \right)}{\Gamma (m_{SR})}, & \text{if } u < \frac{a_2}{a_1}, \\ 1, & \text{otherwise.} \end{cases} \quad (C.37)$$

Then, using (5.37), the CDF of $\gamma_{1,2}$ can be derived as

$$F_{\gamma_{1,2}}(\xi_2) = \Pr \left[ \frac{\xi_2 d_1^r}{\rho_R (b_2 - b_1 \xi_2)} < |h_{RP}|^2 < \frac{\rho_I d_{RP}^r}{\rho_R} \right] + \Pr \left[ |h_{1}|^2 < \frac{\xi_2 d_1^r}{\rho_R (b_2 - b_1 \xi_2)}, |h_{RP}|^2 < \frac{\rho_I d_{RP}^r}{\rho_R} \right]. \quad (C.38)$$

Similar as in (C.33), the RVs $|h_1|^2$ and $|h_{RP}|^2$ in $M_1$ are independent from each other, thus, $M_1$ can be rewritten as follows

$$M_1 = \left( 1 - \frac{\left( m_{RP}, \frac{\lambda_{RP} d_{RP}^r}{\rho_R} \right)}{\Gamma (m_{RP})} \right) \left( 1 - \frac{\left( m_1, \frac{\lambda_1 \xi_2 d_1^r}{\rho_R (b_2 - b_1 \xi_2)} \right)}{\Gamma (m_1)} \right). \quad (C.39)$$

Now, the term $M_2$ in (C.38) can be further expressed by following the same steps as in the derivation of $N_2$ (shown in Appendix C.5)

$$M_2 = \frac{\Gamma\left( m_{RP}, \frac{\lambda_{RP} d_{RP}^r}{\rho_R} \right)}{\Gamma (m_{RP})} - \frac{\lambda_{RP} m_{RP} - 1}{\Gamma (m_{RP})} \sum_{i=0}^{m_1} \left( \frac{\lambda_1 \xi_2 d_1^r}{\rho_R (b_2 - b_1 \xi_2)} \right)^i \frac{1}{i!} \times \Gamma\left( m_{RP} + i, \frac{\rho_I d_{RP}^r}{\rho_R} \right) c_2^{-m_{RP} + i}. \quad (C.40)$$
where \(c_2 = \lambda_{RP} + \frac{\lambda_1 \xi_2 d_1^R}{\rho_R d_{RP}(b_2 - b_1 \xi_2)}\). Further, by substituting (C.39) and (C.40) into (C.38), and after some algebraic manipulations, the CDF in (C.38) can be re-expressed as follows:

\[
F_{\gamma_{1,2}}(\xi_2) = 1 - \left( 1 - \frac{\Gamma\left( \frac{m_{RP} \lambda_{RP} d_{RP}}{\rho_R} \right)}{\Gamma(\frac{m_{RP}}{\rho_R})} \right) \frac{\Gamma\left( \frac{m_1 \lambda_1 \xi_2 d_1^R}{\rho_R (b_2 - b_1 \xi_2)} \right)}{\Gamma(m_1)} - \frac{\lambda_{RP}^m}{\Gamma(m_{RP})} \sum_{i=0}^{m_{RP}-1} \Gamma\left( m_{RP} + i, \frac{\rho_R d_{RP}}{\rho_R} \right) \left( \frac{\lambda_1 \xi_2 d_1^R}{\rho_R d_{RP}(b_2 - b_1 \xi_2)} \right)^i i! c_2^{m_{RP}+i}, \text{ if } \xi_2 < \frac{b_2}{b_1}.
\]

Moreover, the CDF of the RV \(\gamma_{2,2}\) can be derived using (5.39) and written as

\[
F_{\gamma_{2,2}}(\xi_2) = \begin{cases} 
\Pr \left[ |h_2|^2 < \frac{\xi_2 d_2^R}{\rho_R (b_2 - b_1 \xi_2)}, |h_{RP}|^2 < \frac{\rho_R d_{RP}}{\rho_R} \right] & \text{in } O_1, \\
\Pr \left[ |h_2|^2 < \frac{\xi_2 d_2^R}{\rho_R d_{RP} (b_2 - b_1 \xi_2)}, |h_{RP}|^2 > \frac{\rho_R d_{RP}}{P_R} \right] & \text{in } O_2.
\end{cases}
\]

Similar as in (C.33) and (C.39), \(|h_2|^2\) and \(|h_{RP}|^2\) in \(O_1\) are independent from each other, thus, \(O_1\) can be further expressed by

\[
O_1 = 1 - \frac{\Gamma\left( m_{RP}, \frac{\lambda_{RP} d_{RP}}{\rho_R} \right)}{\Gamma(\frac{m_{RP}}{\rho_R})} - \frac{\Gamma\left( m_2, \frac{\lambda_2 d_1^R}{\rho_R (b_2 - b_1 \xi_2)} \right)}{\Gamma(m_2)} \left( 1 - \frac{\Gamma\left( m_{RP}, \frac{\lambda_{RP} d_{RP}}{\rho_R} \right)}{\Gamma(\frac{m_{RP}}{\rho_R})} \right).
\]

Then, the term \(O_2\) in (C.42) can be rewritten as

\[
O_2 = \frac{\Gamma\left( m_{RP}, \frac{\lambda_{RP} d_{RP}}{\rho_R} \right)}{\Gamma(\frac{m_{RP}}{\rho_R})} - \frac{\lambda_{RP}^m}{\Gamma(m_{RP})} \sum_{i=0}^{m_{RP}-1} \left( \frac{\lambda_2 d_1^R}{\rho_R d_{RP}(b_2 - b_1 \xi_2)} \right)^i i! c_3^{(m_{RP}+i)} \times \Gamma\left( m_{RP} + i, \frac{\rho_R d_{RP}}{\rho_R} c_3 \right), \text{ if } \xi_2 < \frac{b_2}{b_1},
\]

where \(c_3 = \lambda_{RP} + \frac{\lambda_1 \xi_2 d_1^R}{\rho_R d_{RP}(b_2 - b_1 \xi_2)}\). Then, by substituting (C.43) and (C.44) into (C.42), and after some algebraic manipulations, the CDF of \(\gamma_{2,2}\) can be written in a closed-form as follows

\[
F_{\gamma_{2,2}}(\xi_2) = 1 - \left( 1 - \frac{\Gamma\left( m_{RP}, \frac{\lambda_{RP} d_{RP}}{\rho_R} \right)}{\Gamma(\frac{m_{RP}}{\rho_R})} \right) \frac{\Gamma\left( m_2, \frac{\lambda_2 d_1^R}{\rho_R (b_2 - b_1 \xi_2)} \right)}{\Gamma(m_2)} - \frac{\lambda_{RP}^m}{\Gamma(m_{RP})}
\]
Finally, the OP of $x_2$ can be derived in a closed-form by substituting (C.37), (C.41) and (C.45) into (5.44).

C.7 Proof of Proposition 11

The CDF of the RV $\gamma_{R,j}$ in (5.49) can be further extended as

$$F_{\gamma_{R,j}}(\xi_j) = \Pr \left[ \frac{|\hat{h}_{SR}|^2}{|h_{SR}|^2} < \xi_j, \frac{\varrho S}{\bar{\varrho} S + \zeta R d_{SR}^\varrho} < \frac{\rho_1 d_{SP}^\varrho}{\rho S}, \left| \frac{\hat{h}_{SR}}{|h_{SR}|^2} \right|^2 \right]$$

$$+ \Pr \left[ \frac{|\hat{h}_{SR}|^2}{|h_{SR}|^2} < \xi_j, \frac{\varrho S}{\bar{\varrho} S + \zeta R d_{SR}^\varrho} > \frac{\rho_1 d_{SP}^\varrho}{\rho S}, \left| \frac{\hat{h}_{SR}}{|h_{SR}|^2} \right|^2 > \frac{\rho_1 d_{SP}^\varrho}{\rho S} \right]$$

$$= \Pr \left[ \frac{|\hat{h}_{SR}|^2}{|h_{SR}|^2} < \frac{\xi_j}{\mu}, \left| \frac{\hat{h}_{SR}}{|h_{SR}|^2} \right|^2 > \frac{\rho_1 d_{SP}^\varrho}{\rho S} \right],$$

where $\mu = \alpha_j - \xi_j \Theta, \tilde{\xi}_R = \frac{\rho_1 d_{SP}^\varrho |h_{SR}|^2}{|h_{SR}|^2} + \nu + 1, \mathcal{G}_{R,j} = \frac{\xi_j (\nu+1) d_{SR}^\varrho}{\rho_1 d_{SP}^\varrho}$ and $\mathcal{O}_{R,j} = \xi_j^2 h_{SR}^2$.

Further, as RVs $|h_{SR}|^2$ and $|\hat{h}_{SR}|^2$ in $A_3$ of (4.6) are independent from each other, the term $A_3$ can be re-expressed by follows

$$A_3 = \left( 1 - e^{-\frac{\xi_j d_{SR}^\varrho}{2|\varrho| S}} \right) \left( 1 - e^{-\frac{\rho_1 d_{SP}^\varrho}{\varrho S}} \right)$$

$$= 1 - e^{-\frac{\xi_j d_{SR}^\varrho}{2|\varrho| S}} - e^{-\frac{\rho_1 d_{SP}^\varrho}{\varrho S}} + e^{-\frac{\xi_j d_{SR}^\varrho + \rho_1 d_{SP}^\varrho}{2|\varrho| S}}. \quad (C.47)$$

Then, the term $B_3$ in (4.6) can be further written as

$$B_3 = \int_{\rho_1 d_{SP}^\varrho}^{\infty} \int_{0}^{\frac{\nu \mathcal{G}_{R,j} + \mathcal{O}_{R,j}}{\varrho}} f_{|h_{SP}|^2}(y) \int_{0}^{\frac{\nu \mathcal{G}_{R,j} + \mathcal{O}_{R,j}}{\varrho}} f_{|\hat{h}_{SR}|^2}(z) \mathrm{d}z \mathrm{d}y$$

$$= \int_{\rho_1 d_{SP}^\varrho}^{\infty} f_{|h_{SP}|^2}(y) \left( 1 - e^{-\frac{\nu \mathcal{G}_{R,j} + \mathcal{O}_{R,j}}{2|\varrho| S}} \right) \mathrm{d}y = e^{-\frac{\rho_1 d_{SP}^\varrho}{\varrho S}} - \frac{e^{-\frac{\rho_1 d_{SP}^\varrho}{\varrho S}} \frac{e^{-\frac{\nu \mathcal{G}_{R,j} + \mathcal{O}_{R,j}}{2|\varrho| S}}}{\mu + \mathcal{G}_{R,j}}}{\mu + \mathcal{G}_{R,j}}. \quad (C.48)$$
Finally, after substituting (C.47) and (C.48) into (C.46), and using some algebraic manipulations, the CDF of the RV $\gamma_{R,j}$ can be written in its closed form as in (5.59). ■

C.8 Proof of Proposition 12

Considering the relay power constraint, the CDF of the RV $\gamma_{k,j}$ is written as

$$
F_{\gamma_{k,j}}(\xi) = \Pr \left[ \frac{|\hat{h}_k|^2 \beta_i \rho_R}{|\hat{h}_k|^2 \rho_R} < \xi_j, \rho_R < \frac{\rho_I d_{RP}^2}{|\hat{h}_{RP}|^2} \right]
+ \Pr \left[ \frac{|\hat{h}_k|^2 \beta_i \rho_I d_{RP}^2}{|\hat{h}_k|^2 \rho_R} < \xi_j, \rho_R > \frac{\rho_I d_{RP}^2}{|\hat{h}_{RP}|^2} \right]
= \Pr \left[ |\hat{h}_k|^2 < \frac{\xi_j \zeta_k d_k^2}{\epsilon \rho_R}, |\hat{h}_{RP}|^2 < \frac{\rho_I d_{RP}^2}{\rho_R} \right],
$$

(C.49)

where

$$
\zeta_k = \frac{\rho_I d_{RP}^2 \sigma_h^2}{|\hat{h}_{RP}|^2 d_k^2} + \nu + 1, \quad \epsilon = \beta_j - \xi_j \Phi, \quad G_{k,j} = \frac{\xi_j (\nu+1) d_k^2}{\rho_I d_{RP}^2} \text{ and } \mathcal{O}_{k,j} = \xi_j \sigma_{\hat{h}_k}^2.
$$

In the previous sections, channels of NOMA users have been ordered by their distance. However, in this system model, NOMA channels are ordered regarding order statistics [197]. First, let’s write the CDFs of unordered NOMA channels $|\tilde{h}_k|^2$ in (5.53), which follow an exponential distribution

$$
F_{|\tilde{h}_k|^2}(\xi_j) = \Pr \left[ |\tilde{h}_k|^2 < \frac{\xi_j \zeta_k d_k^2}{\epsilon \rho_R} \right] = \begin{cases} 
1 - e^{-\frac{\xi_j \zeta_k d_k^2}{\epsilon \rho_R}}, & \xi_j < \frac{\beta_j}{\Phi}, \\
1, & \text{otherwise}
\end{cases}
$$

(C.50)

Then, by analyzing order statistics and using [180, Eq. (19)], the CDF of ordered RV $|\hat{h}_k|^2$ is expressed by

$$
F_{|\hat{h}_k|^2}(\xi_j) = \frac{K!}{(K-k)!(k-1)!} \sum_{i=0}^{K-k} \left( \begin{array}{c} M-k \\ i \end{array} \right) \left[ 1 - e^{-\frac{\xi_j \zeta_k d_k^2}{\epsilon \rho_R}} \right]^{k+i} L_k(i),
$$

(C.51)

Further, due to the independence of RVs $|\tilde{h}_k|^2$ and $|\hat{h}_{RP}|^2$ from each other, the term $D_1$ in (C.49) can be extended using binomial expansion as

$$
D_1 = L_k(i) \left( 1 - e^{-\frac{\xi_j \zeta_k d_k^2}{\epsilon \rho_R}} \right)^{k+i} \left( 1 - e^{-\frac{\rho_I d_{RP}^2}{\rho_R}} \right)
$$
Appendix C. Appendices for Chapter 5

C.9 Proof of Proposition 13

\[
\gamma_{k,j} \text{ is derived as in (5.60).} \]

Moreover, the term \( D_2 \) in (C.49), using binomial expansion, can be further extended as

\[
D_2 = \int_{\rho R}^{\infty} f_{\rho R}^2(z) \left[ L_k(i) \left( 1 - e^{-\frac{z \rho_{k,j} + \rho_{k,j}}{2\pi}} \right)^{k+i} \right] dz
\]

\[
= L_k(i) \int_{\rho R}^{\infty} \frac{1}{2} e^{-\frac{z}{2}} \left( 1 - e^{-\frac{z \rho_{k,j} + \rho_{k,j}}{2\pi}} \right)^{k+i} dz
\]

\[
= \frac{1}{2} L_k(i) \int_{\rho R}^{\infty} e^{-\frac{z}{2}} \sum_{n=0}^{k+i} \binom{k+i}{n} e^{-\frac{z (\rho_{k,j} + \rho_{k,j})^n}{2\pi}} dz
\]

\[
= \frac{1}{2} L_k(i) \sum_{n=0}^{k+i} \left( \binom{k+i}{n} e^{-\frac{z (\rho_{k,j} + \rho_{k,j})^n}{2\pi}} \right)
\]

Then, after substituting (C.52) and (C.53) into (C.49), and some algebraic manipulations, a closed-form expression for the CDF of \( \gamma_{k,j} \) is derived as in (5.60).

C.9 Proof of Proposition 13

Considering the source and relay power constraint, (5.74) can be further rewritten as

\[
P_{\text{out},1}(\xi_1) = \text{Pr} \left[ \frac{n_x Y n_1 X \alpha_1}{n_x Y n_1 X \alpha_2 + n_x Y \rho + n_1 Y \rho} < \xi_1, \rho_S < \frac{\rho t d_{SP}}{Z}, \rho_R < \frac{\rho t d_{RP}}{Q} \right] + \text{Pr} \left[ \frac{\rho t d_{SP}}{Z} \right.
\]

\[
+ \left. \frac{n_x Y n_1 X \alpha_1}{n_x Y n_1 X \alpha_2 + n_x Y \rho + n_1 Y \rho} < \xi_1, \rho_R > \frac{\rho t d_{RP}}{Q} \right] + \text{Pr} \left[ \frac{n_x Y m_1 X \alpha_1}{n_x Y m_1 X \alpha_2 + m_1 X \rho + n_1 Y \rho} < \xi_1, \rho_S < \frac{\rho t d_{SP}}{Z}, \rho_R > \frac{\rho t d_{RP}}{Q} \right]
\]

\[
+ \left. \frac{n_x Y m_1 X \alpha_1}{n_x Y m_1 X \alpha_2 + m_1 X \rho + n_1 Y \rho} < \xi_1, \rho_R > \frac{\rho t d_{RP}}{Q} \right] + \text{Pr} \left[ \frac{n_x Y \psi - \xi_1}{n_x Y \psi + \psi} < n_x Y \xi_1 + \psi, Z < C_S, Q < C_R \right]
\]

\[
+ \left. \frac{n_x Y \psi - \xi_1}{n_x Y \psi + \psi} < n_x Y \xi_1 + \psi, Z > C_S, Q < C_R \right]
\]

\[
+ \left. \frac{m_1 X \left( n_x Y \psi - \xi_1 \right)}{m_1 X \left( n_x Y \psi + \psi \right)} < \frac{n_x Y \xi_1 + \psi}{n_x Y \xi_1 + \psi}, Z < C_S, Q > C_R \right]
\]

\[
+ \left. \frac{m_1 X \left( n_x Y \psi - \xi_1 \right)}{m_1 X \left( n_x Y \psi + \psi \right)} < \frac{n_x Y \xi_1 + \psi}{n_x Y \xi_1 + \psi}, Z > C_S, Q > C_R \right]
\]
= \Pr \left[ Y < u_{s,1}, X < \frac{\Delta s}{n_1}, Z < C_S, Q < C_R \right] \\
+ \Pr \left[ Y < Z u_{p,1}, X < \frac{\Delta P}{n_1}, Z > C_S, Q < C_R \right] \\
+ \Pr \left[ Y < u_{s,1}, X < \frac{Q \Delta S}{m_1}, Z < C_S, Q > C_R \right] \\
+ \Pr \left[ Y < Z u_{p,1}, X < \frac{Q \Delta P}{m_1}, Z > C_S, Q > C_R \right] 

(C.54)

where \( u_{v,1} = \frac{\xi_1}{n_v} \), with \( \forall v \in \{ s, p \} \), \( \xi_1 = \bar{\nu} \xi_1 \), \( m_1 = \frac{\rho I d \tau_{RP}}{P} \), \( Q = |h_{RP}|^2 \), \( C_L = \rho I d \tau_{LP} \rho L \), \( \Delta_S = \frac{Y u_{s,1} + u_{s,1}}{Y - u_{s,1}} \) and \( \Delta_P = \frac{Y u_{p,1} + Z u_{p,1}}{Y - Z u_{p,1}} \). Moreover, the step \( a \) in (C.54) relies on \( \alpha_1 > \alpha_2 \xi_1 \), otherwise, \( P_{1}(\xi_1) = 1 \) regardless of the system SNR.

Now, we further extend the term \( A \) in (C.54) as

\[
A = \int_{0}^{C_S} f_Z(z) \, dz \int_{0}^{C_R} f_Q(q) \, dq \int_{0}^{u_{s,1}} f_Y(y) \, dy \\
+ \int_{0}^{C_S} f_Z(z) \, dz \int_{0}^{C_R} f_Q(q) \, dq \int_{u_{s,1}}^{\infty} F_X(\Delta_S) f_Y(y) \, dy, 
\]

(C.55)

Since the PDFs in the terms \( A_1 \) and \( A_2 \) are independent from each other, they can be rewritten as \( A_1 = \left( 1 - e^{-C_S} \right) \left( 1 - e^{-C_R} \right) \left( 1 - e^{-u_{s,1}} \right) \) and \( A_2 = \left( 1 - e^{-C_S} \right) \left( 1 - e^{-C_R} \right) \). Then, the term \( A_3 \) can be further extended as

\[
A_3 = \int_{u_{s,1}}^{\infty} e^{-y} \, dy - \int_{u_{s,1}}^{\infty} e^{-\frac{Y u_{s,1} + u_{s,1}}{n_1} y} \, dy. 
\]

(C.56)

The term \( A_{31} \) in (C.56) can be further rewritten as

\[
A_{31} = \int_{u_{s,1}}^{\infty} e^{-y} \, dy = e^{-u_{s,1}}. 
\]

(C.57)

Moreover, by substituting \( t = y - u_{s,1} \), \( A_{32} \) in (C.56) can be rewritten as follows

\[
A_{32} = \int_{0}^{\infty} e^{-\frac{Y u_{s,1} + u_{s,1}^2 u_{s,1}}{n_1^2} t - t u_{s,1}} \, dt = e^{-u_{s,1}(\frac{1}{n_1} + \frac{1}{n_1})} \int_{0}^{\infty} e^{-\frac{t}{n_1}} \, t^{\frac{1}{n_1} - 1} \, dt. 
\]

(C.58)
Then, using [121, Eq. (3.324.1)], \( A_{32} \) can be further derived as

\[
A_{32} = e^{-\frac{u_{s,1}(n_1+n_2)}{n_1}} 2\sqrt{r_1} K_1 (2\sqrt{r_1}) ,
\]

(C.59)

where \( r_1 = \frac{u_{s,1}(n_1+n_2)}{n_1} \) and \( K_1(\cdot) \) is the modified Bessel function of the second kind of order 1. Further, after substituting \( A_1, A_2, A_{31} \) and \( A_{32} \) into (C.55) and some mathematical manipulations, the term \( A \) can be written in its closed-form as

\[
A = \left( 1 - e^{-C_S} \right) \left( 1 - e^{-C_R} \right) \left( 1 - e^{-\frac{u_{s,1}(n_1+n_2)}{n_1}} 2\sqrt{r_1} K_1 (2\sqrt{r_1}) \right).
\]

(C.60)

Now, the term \( B \) in (C.54) can be further extended as

\[
B = \int_0^{C_R} f_Q(q) dq \int_C^{\infty} f_Z(z) \int_0^{z_{u_{p,1}}} f_Y(y) dy \, dz \biggr|_{B_1} + \int_0^{C_K} f_Q(q) dq \int_C^{\infty} f_Z(z) \int_{z_{u_{p,1}}}^{\infty} f_Y(y) F_X \left( \frac{\Delta_p}{n_1} \right) dy \, dz ,
\]

(C.61)

where \( B_1 \) and \( B_{21} \) can be respectively derived as \( B_1 = \left( 1 - e^{-C_R} \right) \left( e^{-C_S} - \frac{e^{-CS (u_{p,1})}}{u_{p,1} + 1} \right) \) and \( B_{21} = \left( 1 - e^{-C_R} \right) \) while \( B_{22} \) can be further extended as

\[
B_{22} = \int_C^{\infty} e^{-z} \left[ \int_{z_{u_{p,1}}}^{\infty} e^{-y} dy - \int_{z_{u_{p,1}}}^{\infty} e^{-\frac{\Delta_p}{n_1} y} dy \biggr|_{\Omega_{B_1}} \right] \, dz
\]

(C.62)

Then, using \( p = y - z_{u_{p,1}} \) and [121, Eq. (3.324.1)], \( \Omega_{B_1} \) can be formulated as

\[
\Omega_{B_1} = e^{-\frac{n_{u_{p,1}} - z_{u_{p,1}}}{n_1}} 2\sqrt{b_1} K_1 \left( 2\sqrt{b_1} \right),
\]

(C.63)

where \( b_1 = \frac{z_{u_{p,1}}(n_{u_{p,1}}+1)}{n_1} \). Now, we rewrite \( B_{22} \) as follows

\[
B_{22} = \frac{e^{-CS (u_{p,1})}}{u_{p,1} + 1} - \int_C^{\infty} e^{-\frac{n_{u_{p,1}} (u_{p,1}+1) + n_{u_{p,1}}}{n_1}} 2\sqrt{b_1} K_1 \left( 2\sqrt{b_1} \right) \, dz
\]

(C.64)
Then, using $r = z - C_S$, $B_{22}$ can be reformulated as

$$B_{22} = \frac{e^{-C_S(u_{p,1}+1)}}{u_{p,1} + 1} - e^{-C_S n_1 \frac{(u_{p,1}+1)+u_{p,1}}{n_1}} \int_0^\infty e^{-r_{n_1} \frac{(u_{p,1}+1)+u_{p,1}}{n_1}} 2\sqrt{b_2} K_1 \left(2\sqrt{b_2} \right) dr,$$

(C.65)

where $b_2 = \frac{u_{p,1}(r+C_S)(n_p u_{p,1}+1)}{n_1}$. Then, by applying [196, Eqs. (2.24.1.3) and (8.2.2.15), Ch. 3], $\Omega_{B_2}$ can be derived by follows

$$\Omega_{B_2} = \Upsilon_B \sum_{k=0}^{\infty} \left( -\Theta_B \right)^k \frac{C_{3,2}}{k!} \left( \frac{C_S \Lambda_B}{\Theta_B} \right) \left| 0, k - 1, k \right|,$$

(C.66)

where $\Lambda_B = u_{p,1} + 1$, $\Upsilon_B = \frac{C_S}{\Theta_B}$, $\Theta_B = \frac{u_{p,1} C_S (n_p u_{p,1}+1)}{n_1}$. Thus, $B$ can be written as

$$B = \left(1 - e^{-C_R} \right) \left( e^{-C_S} - e^{-C_R \frac{(u_{p,1}+1)+u_{p,1}}{n_1}} \Omega_{B_2} \right).$$

(C.67)

The term $C$ in (C.54) can be further reformulated as

$$C = \int_0^{C_S} f_z(z) dz \int_{C_R}^{\infty} f_Q(q) dq \int_0^{u_{s,1}} f_Y(y) dy \underbrace{C_1}_{C_{21}} + \int_0^{C_S} f_z(z) dz \int_{C_R}^{\infty} f_Q(q) \int_{u_{s,1}}^{\infty} f_Y(y) F_X \left( \frac{Q \Delta_S}{m_1} \right) dy dq,$$

(C.68)

where $C_1 = e^{-C_R} \left(1 - e^{-C_S} \right) \left(1 - e^{-u_{s,1}} \right)$ and $C_{21} = \left(1 - e^{-C_S} \right)$. Then, $C_{22}$ can be further extended as

$$C_{22} = \int_{C_R}^{\infty} e^{-q} \left[ \int_{u_{s,1}}^{\infty} e^{-y} dy - \int_{u_{s,1}}^{\infty} e^{-\frac{\Delta_S}{m_1} y} dy \right] dq,$$

(C.69)

Further, using $t = y - u_{s,1}$ and [121, Eq. (3.324.1)], $\Omega_{C_1}$ can be rewritten as

$$\beta = e^{\frac{Q n u_{s,1}}{m_1} - u_{s,1}} 2\sqrt{c_1} K_1 \left(2\sqrt{c_1} \right),$$

(C.70)
where \( c_1 = \frac{q u_{s,1}(n_s u_{s,1} + 1)}{m_1} \). Then, we reformulate \( C_{22} \) as

\[
C_{22} = \frac{e^{-C_R(u_{s,1} + 1)}}{u_{s,1} + 1} - \int_{C_R}^{\infty} \frac{e^{-q(m_1 + n_s u_{s,1}) + m_1 u_{s,1}}}{(2 \sqrt{C_1 K_1 (2 \sqrt{C_1})})^{-1}} dq \tag{C.71}
\]

Then, applying \( w = q - C_R \), \( C_{22} \) can be written as

\[
C_{22} = \frac{e^{-C_R(u_{s,1} + 1)}}{u_{s,1} + 1} - e^{-C_R(m_1 + n_s u_{s,1}) + m_1 u_{s,1}} \int_{0}^{\infty} \frac{e^{-w(m_1 + n_s u_{s,1}) + m_1 u_{s,1}}}{2 \sqrt{C_2 K_1 (2 \sqrt{C_2})}} dw, \tag{C.72}
\]

where \( c_2 = \frac{(w+C_R)(n_s u_{s,1} + 1)}{m_1} \). Further, \( \Omega_{C_2} \) can be derived by

\[
\Omega_{C_2} = \frac{C_R}{\Upsilon_C} \sum_{k=0}^{\infty} \frac{(-\Upsilon_C)^k}{k!} C_1^{1/2} \left( \begin{array}{c} C_R \Theta_C / \Upsilon_C \end{array} \right)_{k,0}, \tag{C.73}
\]

where \( \Upsilon_C = \frac{C_R(n_s u_{s,1} + 1)}{m_1} \) and \( \Theta_C = \frac{m_1 + n_s u_{s,1}}{m_1} \). Then, \( C \) can be written as

\[
C = (1 - e^{-C_S}) \left( e^{-C_R (1 - e^{-u_{s,1}})} + \frac{e^{-C_R(u_{s,1} + 1)}}{u_{s,1} + 1} - e^{-C_R(m_1 + n_s u_{s,1}) + m_1 u_{s,1}} \Omega_{C_2} \right). \tag{C.74}
\]

Now, the term \( D \) in (C.54) can be further extended as

\[
D = \int_{C_R}^{\infty} f_Q(q) dq \int_{C_S}^{\infty} f_Z(z) \left[ \int_{0}^{z u_{p,1}} f_Y(y) dy \right] dz
\]

\[
+ \int_{C_R}^{\infty} f_Q(q) \int_{C_S}^{\infty} f_Z(z) \left[ \int_{z u_{p,1}}^{\infty} f_Y(y) F_X(\Delta_P) dy \right] dz dq. \tag{C.75}
\]

The term \( D_1 \) can be derived as

\[
D_1 = \int_{C_R}^{\infty} e^{-q} dq \left( \int_{C_S}^{\infty} e^{-z} dz - \int_{c}^{\infty} e^{-z(u_{p,1} + 1)} dz \right) = e^{-C_R} \left( e^{-C_S} - \frac{e^{-C_S(u_{p,1} + 1)}}{u_{p,1} + 1} \right). \tag{C.76}
\]
Further, the term $D_2$ can be rewritten as

$$
D_2 = \int_{C_R}^{\infty} e^{-q} \int_{C_S}^{\infty} e^{-z} \left( \int_{zu_{p,1}}^{\infty} e^{-y} dy - \int_{zu_{p,1}}^{\infty} e^{-nq_{y_{u_{p,1}}}} dy \right) dz dq. \tag{C.77}
$$

Then, by using substitution $t = y - zu_{p,1}$ and [121, Eq. (3.324.1)], the term $D_{21}$ is derived by

$$
D_{21} = e^{-q_{nq_{u_{p,1}}}} \sqrt{sK_1(2\sqrt{s})}, \tag{C.78}
$$

where $s = \frac{q_{zu_{p,1}}}{m_1}$. Now, we rewrite (C.77) using (C.78) as

$$
D_2 = \int_{C_R}^{\infty} e^{-q} \left( \int_{C_S}^{\infty} e^{-u_{p,1}z-z} dz - \int_{C_S}^{\infty} e^{-\frac{zu_{p,1}}{m_1}+q_{nq_{u_{p,1}}}} 2\sqrt{sK_1(2\sqrt{s})} dz \right) dq \tag{C.79}
$$

Then, by using substituting $w = z - C_S$ in the term $\Omega_{D_1}$ and after some algebraic manipulations, $D_2$ can be reformulated as

$$
D_2 = \frac{e^{-C_R-C_S(u_{p,1}+1)}}{u_{p,1}+1} - \int_{C_R}^{\infty} e^{-q} D_{22} dq, \tag{C.80}
$$

where $D_{22} = \int_{0}^{\infty} e^{-\frac{(um_1+C_Sm_1)(u_{p,1}+1)+q_{nq_{u_{p,1}}}}{m_1}} dw$ and $l = \frac{w_{q_{u_{p,1}}}(u_{p,1}+1)+C_Sq_{u_{p,1}}(u_{p,1}+1)}{m_1}$.

Further, after applying [195, Eqs. (12) and (14)] and [196, Eq. (2.24.1.3)], $D_{22}$ can be derived as

$$
D_{22} = \frac{m_1}{u_{p,1}(u_{p,1}+1)} e^{-c_s(u_{p,1}+1)q^{-1}} e^{-nq_{u_{p,1}}^{m_1}} q^{-1}
\times \sum_{k=0}^{\infty} \frac{(-T)^k}{k!} G_{3}^{1} \left( \frac{(u_{p,1}+1)m_1}{(u_{p,1}+1)u_{p,1}} q^{-1}, k - 1, k \right), \tag{C.81}
$$

where $T = \frac{u_{p,1}}{m_1} q$. Then, we write

$$
\Omega_{D_2} = \mu_{D_1} \sum_{k=0}^{\infty} \Psi(k) \int_{C_R}^{\infty} q^{k-1} e^{-qT_{D_2}} G_{3}^{1} \left( \frac{(u_{p,1}+1)m_1}{(u_{p,1}+1)u_{p,1}} q^{-1}, k - 1, k \right) dq, \tag{C.82}
$$
where $\Upsilon_D = m_1 e^{-Rs_{u,p}^{u,p+1}}$, $\Psi(k) = (-u_{p,1} c_3 (n_{p} u_{p} + 1))^{k}$ and $\Upsilon_D = n_{p} u_{p} + 1$. Assuming that $s = q - C_R$, (C.82) can be rewritten as

$$
\Omega_D = \Upsilon_D e^{-C_R \Upsilon_D} \sum_{k=0}^{\infty} \Psi(k) \int_0^{\infty} \left( s + C_R \right)^{k-1} e^{-s \Upsilon_D} \times G_{2,3}^1 \left( \zeta(s + C_R) \left| \begin{array}{c} 1, 1-k \\ 1, 2-k, 1-k \\ \end{array} \right. \right) ds = \Theta_D \times \Psi(0)
$$

$$
\times \int_0^{\infty} (s + C_R)^{-1} e^{-s \Upsilon_D} \sum_{k=1}^{\infty} \Psi(k) \int_0^{\infty} (s + C_R)^{k-1} e^{-s \Upsilon_D} \Theta_D ds, \tag{C.83}
$$

where $\Theta_D = G_{2,3}^1 \left( \zeta(s + C_R) \left| \begin{array}{c} 1, 1-k \\ 1, 2-k, 1-k \\ \end{array} \right. \right)$, $\Theta_D = \Upsilon_D e^{-C_R \Upsilon_D}$, $\Psi(0) = 1$ and $\zeta = \frac{(n_{p} u_{p} + 1) u_{p} + 1}{u_{p} + 1}$. Then, we have

$$
M_1 = \pi C_R \int_0^{\infty} G_{2,2}^1 \left( \left| \begin{array}{c} 1 \\ 0 \\ \end{array} \right. \right) ds
$$

$$
\times \int_0^{\infty} (s + C_R) \left| \begin{array}{c} \zeta s + \zeta C_R \\ 1, 1, 2, 1 \\ \end{array} \right. \left| \begin{array}{c} 0, 1 \frac{1}{2} \\ 0, 1 \frac{1}{2} \\ \end{array} \right. \right) ds, \tag{C.84}
$$

which has a closed-form solution only if $E_1 \to 0$. If it is true, then

$$
M_1 = \frac{\pi}{C_R} MG_{2,2}^1 \left( \left| \begin{array}{c} \Upsilon_D \\ 0 \\ \end{array} \right. \right) \left| \begin{array}{c} 0 - \frac{1}{2} \frac{1}{2} \\ 0 - \frac{1}{2} \frac{1}{2} \\ \end{array} \right. \right) \left( \left| \begin{array}{c} 0 \frac{1}{2} \frac{1}{2} \\ 1, 2, 1 \\ \end{array} \right. \right), \tag{C.85}
$$

Furthermore,

$$
M_2 = C_R^{-k} \int_0^{\infty} \sum_{n=0}^{k-1} \left( \begin{array}{c} k-1 \\ n \\ \end{array} \right) \left( \frac{s}{C_R} \right)^{n} e^{-s \Upsilon_D} \times G_{2,3}^1 \left( \zeta s + \zeta C_R \left| \begin{array}{c} 1, 1-k \\ 1, 2-k, 1-k \\ \end{array} \right. \right) ds
$$

$$
= C_R^{-k} \sum_{n=0}^{k-1} \left( \begin{array}{c} k-1 \\ n \\ \end{array} \right) \left( \frac{1}{C_R} \right)^{n} \int_0^{\infty} s^{n+1-1} e^{-s \Upsilon_D} ds
$$
\[ \times G_{2,3}^{3,1} \left( \zeta s + \zeta C_R \left| \begin{array}{c} 1, 1-k \\ 1, 2-k, 1-k \end{array} \right. \right) \, ds. \]  

(C.86)

Assuming \( a = \zeta s \) and \( ds = \frac{1}{\zeta} \, da \), \( M_2 \) can be derived as

\[
M_2 = C_{R}^{k-1} \sum_{n=0}^{k-1} \binom{k-1}{n} \left( \frac{1}{C_R \zeta} \right)^n \int_0^\infty a^{n+1-1} e^{-a d_{D_2}} \times G_{2,3}^{3,1} \left( a + \zeta C_R \left| \begin{array}{c} 1, 1-k \\ 1, 2-k, 1-k \end{array} \right. \right) \, ds = C_{R}^{k-1} \sum_{n=0}^{k-1} \binom{k-1}{n} \left( \frac{1}{C_R \zeta} \right)^n \int_0^\infty a^{n+1-1} \mathcal{G}_{1,0}^{1,0} \left( \frac{T_{D_2}}{\zeta} \right) \, ds 
\]

\[ \times G_{2,3}^{3,1} \left( a + \zeta C_R \left| \begin{array}{c} 1, 1-k \\ 1, 2-k, 1-k \end{array} \right. \right) \, ds 
\]

\[ = C_{R}^{k-1} \sum_{n=0}^{k-1} \binom{k-1}{n} \left( \frac{1}{C_R \zeta} \right)^n \sum_{m=0}^{\infty} \frac{(-\zeta C_R)^m}{m!} \times G_{4,4}^{2,4} \left( \frac{T_{D_2}}{\zeta} \left| \begin{array}{c} -n, m-n-1, 2m-n-2, 2m-n-1 \\ 0, m-n-1, 2m-n-1, m-n \end{array} \right. \right). \]  

(C.87)

Then, \( D \) can be written in its closed form as

\[ D = e^{-C_R - C_S} - \Omega_{D_2}. \]  

(C.88)

Finally, after substituting (C.60), (C.67), (C.74) and (C.88) into (C.54), the exact OP expression for \( D_1 \) can be written as in (5.75).
Bibliography


[61] Z. Ding, M. Peng, and H. V. Poor, “Cooperative non-orthogonal multiple access in 5G systems,” *IEEE Communications Letters*, vol. 19, pp. 1462–1465, August 2015.


[185] Y. Liu, Z. Ding, M. Elkashlan, and H. V. Poor, “Cooperative non-orthogonal multiple access with simultaneous wireless information and power transfer,” *IEEE Journal on Selected Areas in Communications*, vol. 34, pp. 938–953, April 2016.


[189] W. Han, J. Ge, and J. Men, “Performance analysis for NOMA energy harvesting relaying networks with transmit antenna selection and maximal-ratio combining over Nakagami-\(m\) fading,” *IET Communications*, vol. 10, no. 18, pp. 2687–2693, 2016.


