

**FULL-DUPLEX WIRELESS COMMUNICATION SYSTEMS:
DESIGN AND PERFORMANCE ANALYSES**



M.Sc. THESIS

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Electronics and Communication Engineering Department

Telecommunications Engineering Programme

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Thesis Advisor: Prof. Dr. İbrahim ALTUNBAŞ

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**TAM ÇİFT YÖNLÜ TELSİZ HABERLEŞME SİSTEMLERİ:
SİSTEM TASARIMI VE PERFORMANS ANALİZİ**

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To my family and spouse



FOREWORD

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ABBREVIATIONS

AF	: Amplify-and-Forward
AWGN	: Additive White Gaussian Noise
BER	: Bit Error Rate
CDF	: Cumulative Distribution Function
CIOD	: Coordinate Interleaved Orthogonal Design
CPEP	: Conditional Pairwise Error Probability
CSI	: Channel State Information
DF	: Decode-and-Forward
EGBMGF	: Extended Generalized Bivariate Meijer-G Function
FD	: Full-Duplex
HD	: Half-Duplex
I-CSI	: Imperfect Channel State Information
ICI	: Inter Channel Interference
i.i.d.	: Identically and Independent Distributed
LI	: Loop-Interference
LOS	: Line-of-Sight
MGF	: Moment Generating Function
MIMO	: Multiple-Input Multiple-Output
ML	: Maximum Likelihood
MRC	: Maximal Ratio Combining
n.i.i.d.	: Non-Identically and Independently Distributed
NLOS	: Non-Line-of-Sight
OS	: Optimal-Selection
OW	: One-Way
OWFD-AF	: One-Way Full-Duplex Amplify-and-Forward Relaying
OWFD-AF-RS	: One-Way Full-Duplex Amplify-and-Forward Relay Selection
OWFD-SSK-AF	: One-Way Full-Duplex Space-Shift-Keying with Amplify-and-Forward Relaying
OWHD-SSK-AF	: One-Way Half-Duplex Space-Shift-Keying with Amplify-and-Forward Relaying
P-CSI	: Perfect Channel State Information
PA	: Power Allocation
PDF	: Probability Density Function
PEP	: Pairwise Error Probability
PLNC	: Physical-Layer Network Coding
PS	: Power-Splitting
PSK	: Phase-Shift-Keying
QAM	: Quadrature Amplitude Modulation
RD	: Partial Selection with R-D Link
RF	: Radio Frequency
RLI	: Partial Selection with Residual LI
SINR	: Signal-to-Interference-Plus-Noise Ratio

SISO	: Single-Input Single-Output
SM	: Spatial Modulation
SNR	: Signal-to-Noise Ratio
SR	: Partial Selection with S-R Link
SRLI	: Partial Selection with S-R Link and Residual LI
SSK	: Space-Shift-Keying
SWIPT	: Simultaneous Wireless Information and Power Transfer
TS	: Time-Switching
TW	: Two-Way
TWFD-AF	: Two-Way Full-Duplex Amplify-and-Forward Relaying
TWFD-AF-RS	: Two-Way Full-Duplex Amplify-and-Forward Relay Selection
TWFD-SM	: Two-Way Full-Duplex Spatial Modulation
TWFD-SM-AF-PS	: Two-Way Full-Duplex Spatial Modulation with Amplify-and-Forward Relaying and Power Splitting
TWHD-AF	: Two-Way Half-Duplex Amplify-and-Forward Relaying
TWHD-SM	: Two-Way Half-Duplex Spatial Modulation
TWHD-SM-AF-PS	: Two-Way Half-Duplex Spatial Modulation with Amplify-and-Forward Relaying and Power Splitting
V-BLAST	: Vertical-Bell Laboratories Layered Space-Time

SYMBOLS

$ \cdot $	: Absolute Value
$\ \cdot\ $	: Frobenius Norm
$\binom{\cdot}{\cdot}$	: Binomial Coefficient
$(\cdot)^H$	: Hermitian Transpose
$(\cdot)^T$	: Transpose
$B(\cdot, \cdot)$	: Beta Function
$\mathcal{CN}(0, \sigma^2)$	: Complex Gaussian Distribution with 0 mean and σ^2 variance
D	: Destination
$E\{\cdot\}$	: Expectation Operator
${}_2F_1(\cdot, \cdot; \cdot; \cdot)$	: Gauss Hypergeometric Function
$F_X(\cdot)$	: CDF of X
$f_X(\cdot)$	: PDF of X
φ	: Energy Conversion Efficiency
G	: Amplification Factor
$\Gamma(\cdot)$	: Gamma Function
$\Gamma(\cdot, \cdot)$	: Upper Incomplete Gamma Function
$\gamma(\cdot, \cdot)$	: Lower Incomplete Gamma Function
$\mathbf{H}_i$	: Residual LI Channel Matrix at S_i
$\mathbf{H}_{j,i}$	: Channel Matrix for $S_i - S_j$ Link
$K_v(\cdot)$	: v^{th} Order Modified Bessel Function of Second Kind
M	: Size of Signal Constellation
$M_X(\cdot)$	: MGF of X
$\mathcal{N}(0, \sigma^2)$	: Gaussian Distribution with 0 mean and σ^2 variance
$Q(\cdot)$	: Q -Function
$\Re[\cdot]$	: Real Part Operator
R	: Relay
R_c	: c^{th} Relay
ρ	: Power Fraction for Information Receiver in the PS Protocol
S	: Source
S_i	: i^{th} Source
$U(\cdot, \cdot; \cdot)$	: Confluent Hypergeometric Function
$\text{Var}\{\cdot\}$	: Variance Operator
$W_{\cdot, \cdot}(\cdot)$	: Whittaker Function



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FULL-DUPLEX WIRELESS COMMUNICATION SYSTEMS: DESIGN AND PERFORMANCE ANALYSES

SUMMARY

Full-Duplex (FD) communications allows receiving and transmitting the signal at the same time interval using the same frequency band, hence it theoretically doubles the capacity, improves the spectral efficiency and reduces end-to-end delay with respect to the conventional half-duplex (HD) communications. Although FD transmission causes strong loop-interference (LI) at the receive antenna, the recent developments on antenna technology and signal processing techniques make it feasible by applying LI cancellation techniques. According to the experimental and theoretical studies in the literature, FD transmission can outperform its counterpart HD transmission, even though the LI cannot be removed completely. Hence, FD communications has recently gained great attention as a potential candidate technology for the next generation wireless communication systems.

In this thesis, a variety of wireless communication systems using FD transmission are designed and analyzed. The outage probability or bit error rate (BER) expressions of the proposed systems are analytically derived. The accuracy of the theoretical analyses for each system is verified by several Monte Carlo type computer simulation results.

Firstly, the cooperative communication systems using one-way (OW) FD transmission and amplify-and-forward (AF) relaying are investigated. The proposed scheme is named as OWFD-AF, where only the relay node operates in FD mode. It is assumed that the direct link between the source and destination is strongly attenuated, so they can communicate with each other via the relay node. In the literature, the performance of the variable-gain OWFD-AF systems over Rayleigh fading channels is investigated and the outage probability expression is obtained in a single-integral form. However, there is no outage performance analysis for the fixed-gain OWFD-AF systems. Hence, as the first study within the scope of thesis, the outage performance of the fixed-gain OWFD-AF systems over Rayleigh fading channels is analyzed and the outage probability expression is derived in a single-integral form. In addition to simpler structure advantage of fixed-gain relaying compared to variable-gain relaying, it is demonstrated that fixed-gain OWFD-AF systems provides better outage performance. Afterwards, the variable-gain and fixed-gain OWFD-AF systems over Nakagami- m fading channels are also investigated in terms of outage performance. The exact, lower-bound and asymptotic outage probability expressions are derived for each system. In order to improve the performance of the variable-gain OWFD-AF systems over Nakagami- m fading channels, five different relay selection policies are applied and the proposed system is called as OWFD-AF-RS system. The exact, lower-bound and asymptotic outage probability expressions are also derived for each relay selection policy. The numerical results shows that the outage performance of the OWFD-AF-RS system can be enhanced by increasing the number of available relays for the relay selection policies.

Secondly, we have analyzed the performance of two-way (TW) FD cooperative systems with variable-gain AF relaying, which is called as TWFD-AF. In the proposed system, the source nodes employ physical-layer network coding (PLNC) technique and they are assumed to communicate with each other via the relay node similar to OWFD-AF. Due to the FD transmission at each node, including the sources, two-way transmission can be accomplished in a single time-slot. Therefore, TWFD-AF provides considerable high spectral efficiency compared to the conventional cooperative systems. At the beginning, the outage performance of the TWFD-AF systems over Nakagami- m fading channels is investigated. The exact, lower-bound and asymptotic outage probability expressions are obtained. Afterwards, the relay selection algorithm is adopted in order to improve the performance of TWFD-AF systems over Nakagami- m fading channels, which is called as TWFD-AF-RS system. We also derive the exact, lower-bound and asymptotic outage probability expressions for the TWFD-AF-RS system.

Then, spatial modulation (SM) technique in multiple-input multiple-output (MIMO) architecture is combined with FD communications. In the SM technique, only one of the selected transmit antenna is activated in each symbol period. Therefore, the information bits are conveyed via not only modulated signals (e.g., M -PSK/QAM), but also the transmit antenna indices. Contrary to the conventional MIMO systems, the necessity of inter channel interference (ICI) cancellation process can be avoided and the number of required transmit radio frequency (RF) chain can be decreased to one in the SM technique. In addition to its low-complexity maximum likelihood (ML) detection, SM technique provides better error performance compared to the V-BLAST architecture. In the literature, there are a few studies analyzing SM with FD transmission. In this thesis, the bit error performance of two-way full-duplex spatial modulation (TWFD-SM) systems with two source nodes over Rayleigh fading channels is investigated under the availability of the perfect channel state information (P-CSI) and imperfect channel state information (I-CSI) at the sources. By using pairwise error probability (PEP) of the TWFD-SM system, the upper-bound BER expression is obtained in a closed-form. According to the numerical results, it is illustrated that TWFD-SM provides better error performance compared to the SM systems using two-way HD transmission, namely TWHD-SM systems, especially for the higher values of spectral efficiency. Then, a cooperative communication system using space-shift-keying (SSK) at the source node and fixed-gain AF relaying at the relay node is designed, where the relay uses one-way FD transmission. The proposed scheme is called as OWFD-SSK-AF. The upper-bound BER expression of the OWFD-SSK-AF systems over Rayleigh fading channels is also derived in a closed-form. We observe that the OWFD-SSK-AF systems outperforms the conventional one-way HD transmission with SSK and fixed-gain AF relaying (OWHD-SSK-AF) systems.

By combining FD communications, SM, PLNC and simultaneous wireless information and power transfer (SWIPT), a new cooperative communication system is designed at the last part of the thesis. In the SWIPT technique, there are generally two common protocols used for energy harvesting: time-switching (TS) and power-splitting (PS). The proposed system consists of two source and one relay node, where all nodes operates in FD mode in order to complete two-way transmission in a single time-slot. The source nodes employ SM and PLNC techniques and the relay node uses fixed-gain AF relaying. Additionally, there is no any external power supply at the relay node and

it harvests the required energy via the transmitted signals from the sources by using the PS protocol. Therefore, the proposed scheme is called as TWFD-SM-AF-PS. The upper-bound BER expression for the TWFD-SM-AF-PS systems over Rayleigh fading channels is derived in a closed-form.

The main contribution of this thesis can be summarized as follows. Firstly, the fixed-gain OWFD-AF systems over Rayleigh and Nakagami- m fading channels and the variable-gain OWFD-AF system over Nakagami- m fading channels are investigated in terms of outage probability. It is observed that the fixed-gain OWFD-AF systems provide better performance in addition to its simpler structure. Then, five different relay selection policies are applied to the variable-gain OWFD-AF systems in order to improve the outage performance. Secondly, the outage performance of the TWFD-AF systems over Nakagami- m fading channels is analyzed. Afterwards, the outage performance analyses are generalized for the relay selection algorithm. Thirdly, the SM technique is integrated into FD communications. The bit error performance of the TWFD-SM systems under both P-CSI and I-CSI is investigated in Rayleigh fading channels. Similarly, the bit error performance of OWFD-SSK-AF system is also analyzed. Finally, as the last study in this thesis, TWFD-SM-AF-PS systems using PS protocol at the relay node so as to employ SWIPT technique are analyzed over Rayleigh fading channels and the upper-bound BER expression for the proposed system is derived in a closed-form.



TAM ÇİFT YÖNLÜ TELSİZ HABERLEŞME SİSTEMLERİ: SİSTEM TASARIMI VE PERFORMANS ANALİZİ

ÖZET

Tam çift yönlü (full-duplex, FD) iletim tekniğinde gönderim ve alım işlemleri eş zamanlı olarak aynı frekans bandı üzerinden yapıldığından, FD iletim tekniği geleneksel yarı çift yönlü (half-duplex, HD) iletim tekniğine göre teorik olarak kapasiteyi iki katına çıkartmaktadır. FD iletim sebebi ile alıcı antende oluşan kuvvetli öz girişim (loop-interference, LI), günümüzde gelişen işarete işleme teknikleri ve anten teknolojileri sayesinde yüksek bir performans ile giderilebilmektedir. Literatürde yapılan pratik ve teorik çalışmalar göstermektedir ki, alıcıda meydana gelen LI tam olarak giderilemese de FD iletim tekniği geleneksel HD iletim tekniğine göre daha iyi performans verebilmektedir. Bu neden ile uygulanabilir hale gelen FD iletim tekniği, yüksek kapasite ve yüksek iletim hızlarına ihtiyaç duyulacak gelecek nesil iletişim teknolojilerinde yer alabilecek çok önemli bir aday tekniktir ve araştırmacıların oldukça ilgisini çekmektedir.

Bu çalışmada FD iletim tekniğini kullanan çeşitli telsiz iletişim sistemleri tasarlanmıştır. Önerilen sistemlerin kesinti olasılığı (outage probability) ya da bit hata olasılığı (bit error rate, BER) performansları analitik olarak ayrı ayrı incelenmiştir. Tüm sistemler için yapılan analizlerin doğruluğu Monte Carlo tipi bilgisayar benzetimleri yardımı ile doğrulanmıştır.

İlk olarak tek-yönlü (one-way, OW) aktarımın yapıldığı işbirlikli iletişim sistemlerinde, röle düğümünün FD iletim yaptığı sistemler incelenmiştir. Röle düğümünde kuvvetlendir ve aktar (amplify-and-forward, AF) tekniğini kullanan bu sistemlere OWFD-AF adı verilmiştir. İncelenecek OWFD-AF sisteminde kaynak ve hedef arasındaki doğrudan bağlantının kuvvetli bir biçimde sönmüldüğü varsayılarak, iletişimin sadece röle birimi aracılığı ile gerçekleştirildiği düşünülmektedir. Tek-yönlü aktarımını yapıldığı ve geleneksel HD iletim ile AF tekniğini kullanan işbirlikli iletişim sistemlerde iletim için iki zaman dilimi gerekirken, OWFD-AF sistemlerde iletim tek zaman diliminde gerçekleştirilmektedir. Literatürde değişken kazançlı OWFD-AF sistemlerin Rayleigh sönmülemeli kanallardaki kesinti olasılıkları incelenmiş ve sisteme ait tam kesinti olasılığı ifadesi tek katlı integral biçiminde bulunmuştur. Fakat daha basit bir yapıya sahip olan sabit kazançlı OWFD-AF sistemlerin kesinti performansına ait literatürde herhangi bir çalışma bulunmamaktadır. Bu sebeple tez kapsamında yapılan ilk çalışmada sabit kazançlı OWFD-AF sistemlerin Rayleigh sönmülemeli kanallardaki kesinti performansı incelenmiş ve tam kesinti olasılığı ifadesi tek katlı integral biçiminde bulunmuştur. Elde edilen nümerik sonuçlar ışığında daha basit yapısına ek olarak, sabit kazançlı OWFD-AF sistemlerin değişken kazançlı OWFD-AF sistemlere göre daha iyi performans verdiği gösterilmiştir. Ardından hem sabit, hem de değişken kazançlı OWFD-AF sistemlerin kesinti olasılığı Nakagami- m sönmülemeli kanallarda incelenmiştir. Her iki sistem için de tam kesinti olasılığı ifadelerinin yanı sıra, alt-sınır ve asimptotik kesinti olasılığı ifadeleri elde edilmiştir.

Sabit ve deęişken kazançlı sistemlerin performansları elde edilen kesinti olasılığı ifadeleri yardımı ile karşılaştırılmış olup, Nakagami- m sönümlemeli kanallarda da sabit kazançlı OWFD-AF sisteminin daha iyi performans verdiği gösterilmiştir. OWFD-AF sistemine ait yapılan son çalışmada, ise röle seçimli (relay selection, RS) OWFD-AF sistemlerin Nakagami- m kanallardaki performansı incelenmiştir. OWFD-AF-RS adı verilen bu sistemlerde başarıyı arttırmak için beş farklı röle seçim kriteri kullanılmaktadır. Önerilen tüm kriterler için tam, alt-sınır ve asimptotik kesinti olasılığı ifadeleri ayrı ayrı elde edilmiştir. Seçim kriterlerinin performansları hem kendi içlerinde, hem de röle seçimsiz durum ile karşılaştırarak OWFD-AF-RS sisteminin sağlayacağı performans avantajları gösterilmiştir.

İkinci incelenen sistemde ise çift-yönlü (two-way, TW) aktarım yapan FD iletim ve deęişken kazançlı AF tekniklerini kullanan telsiz iletişim sistemleri incelenmiştir. TWFD-AF adı verilen bu sistemde, kaynaklar fiziksel katman ağ kodlama (physical-layer network coding, PLNC) tekniğini kullanılmaktadır ve OWFD-AF sistemlerde olduğu gibi kaynaklar arasındaki iletişimin sadece röle üzerinden kurulduğu varsayılmaktadır. Tüm düğümlerin FD iletim yaptığı TWFD-AF sistemlerinde, iki-yönlü aktarım tek zaman diliminde gerçekleşmektedir. Bu nedenle TWFD-AF sistemleri, geleneksel işbirlikli sistemlere göre oldukça yüksek bir bant verimlilięi sağlamaktadır. Rayleigh sönümlemeli kanallarda TWFD-AF sisteminin kesinti performansına ait literatürde çeşitli çalışmalar bulunmaktadır. Tez kapsamında TWFD-AF sistemler için ilk olarak, Nakagami- m sönümlemeli kanallardaki kesinti performansı incelenmiştir. Yapılan teorik analizler ile TWFD-AF sistemler için tam, alt-sınır ve asimptotik kesinti olasılığı ifadeleri elde edilmiştir. Ardından incelenen sistemin performansını iyileştirmek amacı ile röle seçimli TWFD-AF sistemlerin Nakagami- m sönümlemeli kanallardaki kesinti performansı incelenmiştir. TWFD-AF-RS adı verilen bu sistem için de tam, alt-sınır ve asimptotik kesinti olasılığı ifadeleri elde edilmiştir ve uygulanan röle seçiminin sistemin performansını iyileştirdięi gösterilmiştir.

Üçüncü kısımda ise çok girişli çok çıkışlı (multiple-input multiple-output, MIMO) yapılar da FD iletim yapan uzaysal modülasyon (spatial modulation, SM) sistemleri incelenmiştir. Klasik SM tekniğinde, her simge süresince sadece bir adet verici anteni aktif olarak çalışmaktadır. Bu nedenle bilgi bitleri hem modülasyonlu işaretler (örneğin M -PSK/QAM), hem de verici anten indisleri ile taşınmaktadır. SM tekniğinde klasik MIMO sistemlerin aksine verici RF zinciri sayısı bire düşürülmekte ve meydana gelen kanallar arası girişimin (inter channel interference, ICI) oluşması önlenmektedir. Ayrıca en büyük olabilirlik (maximum likelihood, ML) karar kuralını kullanan düşük karmaşıklıkla alıcı yapısına sahip SM tekniğinin, yüksek karmaşıklıęa sahip geleneksel V-BLAST mimarisine göre daha iyi BER performansı verdiği literatürde gösterilmiştir. FD iletim yaparken SM tekniğini kullanan sistemlere ait literatürde birkaç çalışma bulunmaktadır. Bu tez kapsamında ilk olarak, birbirleriyle eş zamanlı olarak çift-yönlü iletişim kuran ve FD iletim ile SM tekniğini kullanan iki kaynak düğümünün Rayleigh sönümlemeli kanallardaki BER performansı incelenmiştir. TWFD-SM adı verilen bu sistemde kaynaklarda hem kesin kanal durum bilgisi (perfect channel state information, P-CSI), hem de kusurlu kanal bilgisinin (imperfect channel state information, I-CSI) olması durumu için çiftsel hata olasılığı (pairwise error probability, PEP) ifadeleri yardımıyla, üst-sınır BER ifadeleri kapalı biçimde bulunmuştur. Ardından TWFD-SM sistemlerin BER performansı, geleneksel HD iletim ile SM tekniğini kullanan (kısaca TWHD-SM)

sistemler ile karşılaştırılmıştır. Artan bant verimlilikleri için TWFD-SM sistemlerin geleneksel TWHF-SM sistemlere göre oldukça iyi BER performansı sağladığı gösterilmiştir. Daha sonra, kaynak düğümünde SM tekniğinin özel bir hali olan ve bilgi bitlerinin sadece seçilen verici anten indisleri ile taşındığı uzay kaydırmalı anahtarlama (space-shift-keying, SSK) tekniğini kullanan ve röle düğümünde FD iletim ve tek-yönlü sabit kazançlı AF aktarım yapan sistemlerin BER performansı incelenmiştir. OWFD-SSK-AF adı verilen bu sistemin üst-sınır BER ifadesi kapalı biçimde bulunmuştur. İncelenen sistemin BER performansı geleneksel HD iletim yapan ve SSK ile tek-yönlü AF aktarım tekniklerini kullanan (kısaca OWHD-SSK-AF) sistemler ile karşılaştırılmış ve OWFD-SSK-AF sistemlerin daha iyi performans verdiği gösterilmiştir.

Tez çalışmasının dördüncü ve son bölümünde ise FD iletim tekniğini ile eş zamanlı telsiz bilgi ve güç aktaran (simultaneous wireless information and power transfer, SWIPT) sistemler incelenmiştir. SWIPT tekniğinde enerji hasatlamak için genel olarak iki farklı protokol bulunmaktadır: zaman anahtarlama (time-switching, TS) ve güç ayırma (power-splitting, PS). Önerilen sistemde kaynaklarda SM ile iletim yapılır ve PLNC kullanılırken, rölede iki-yönlü sabit kazançlı AF aktarım yapılmakta ve enerji hasatlamak için PS kullanılmaktadır. TWFD-SM-AF-PS adı verilen bu sistemde röle düğümünün herhangi bir güç kaynağı bulunmamaktadır ve gerekli enerji kaynaklar tarafından gönderilen işaretler yardımıyla elde edilmektedir. TWFD-AF sistemlerde olduğu gibi tüm düğümler FD iletim tekniğini kullanmaktadır. Önerilen TWFD-SM-AF-PS sistemi için üst-sınır BER ifadesi kapalı biçimde bulunmuştur.

Sonuç olarak tez çalışması kapsamında yapılan çalışmalar ile literatüre yapılan başlıca katkılar şu şekilde özetlenebilir. İlk olarak, sabit kazançlı OWFD-AF sistemlerin Rayleigh ve Nakagami- m sönümlemeli kanallardaki ve değişken kazançlı OWFD-AF sistemlerin Nakagami- m sönümlemeli kanallardaki kesinti performansı incelenmiştir. Daha basit yapıya sahip olmasına karşın, ilk kez önerilen sabit kazançlı OWFD-AF sistemlerin daha yüksek performans sağladığı gösterilmiştir. Değişken kazançlı OWFD-AF sistemlerin performansını arttırmak için beş farklı röle seçim kriteri önerilmiş ve her kriter için kesinti olasılığı ifadeleri ayrı ayrı elde edilmiştir. İkinci olarak, TWFD-AF sistemlerin Nakagami- m sönümlemeli kanallardaki kesinti performansı incelenmiştir ve elde edilen sonuçlar röle seçimli durum için genelleştirilmiştir. Röle seçiminin TWFD-AF sistemler için performans artışı sağladığı gösterilmiştir. Ardından SM tekniğini kullanan TWFD-SM sistemlerin P-CSI ve I-CSI koşulları altında üst-sınır BER ifadeleri elde edilmiştir. Nümerik sonuçlar göstermektedir ki, TWFD-SM sistemleri artan bant verimlilikleri için geleneksel TWHF-SM sistemlere göre oldukça iyi performans sağlamaktadır. Benzer şekilde, işbirlikli sistemlerde OWFD-SSK-AF sisteminin BER performansı da incelenmiş ve klasik SSK-OWHD-AF sistemlere göre daha yüksek BER performansı sağladığı gösterilmiştir. Son olarak da röle düğümünde PS protokolü aracılığı ile SWIPT tekniğini kullanan TWFD-SM-AF-PS sistemlerin BER performansı incelenmiş ve üst-sınır BER ifadesi elde edilmiştir.



1. INTRODUCTION

In wireless communications, the number of users and total traffic rates increase very swiftly through the years. However, the available resources like frequency spectrum and power are quite limited. The fundamental aim of next-generation wireless communication systems is to ensure higher spectral efficiency, capacity and data rates under the constraints of limited resources. Therefore, spectral and energy efficient wireless communication systems have to be generated in order to fulfill the rapidly growing demands.

Full-duplex (FD) communications, allowing simultaneous transmission and reception over the same frequency band, have recently gained great attention as a promising technology for the future generation wireless network technologies [1–7]. Contrary to the conventional half-duplex (HD) communications, FD theoretically doubles the capacity and reduces the end-to-end delay [4]. Additionally, FD improves the network secrecy by means of the mixed signals during the simultaneous transmission [3]. Although FD communication systems are severely affected from loop-interference (LI) (or self-interference) at the receive antenna, the recent developments on antenna technology and signal processing techniques make it feasible by applying LI cancellation techniques, e.g., passive suppression, analog cancellation and digital cancellation [4,8–12]. The practical implementations in [8] shows that the power of LI can be decreased 74 dB on average; however, it cannot be completely removed, even with complex cancellation processes. Even though the residual LI limits performance of the FD communication systems, experimental and theoretical studies in the literature demonstrate that FD transmission can outperform the conventional HD transmission as long as the required values of spectral efficiencies increases and/or quality of the LI cancellation increases. Hence, integration of the FD transmission technique into the conventional HD communication systems has been deeply investigated by researchers in the last decade.

1.1 Objective of Thesis

The main objective of this thesis is to design a variety of new wireless communication systems using FD communications and analyze their performance. During the analyses, the outage probability or bit error rate (BER) expressions of the proposed systems are derived. For the outage performance analyses, cumulative distribution function (CDF) approach is applied [13]. On the other hand, the BER expressions are obtained via pairwise error probability (PEP) and union-bound techniques [13]. For each system, accuracy of the theoretical analyses is validated via a great deal of Monte Carlo type computer simulation results.

1.2 Literature Review

In the recent years, cooperative communication has drawn great interests since it provides coverage extension, capacity, outage and error performance enhancement [14–18]. In cooperative networks, the information at source node is transmitted to the destination node with the help of one or more relays. The relay nodes are mainly categorized into two groups as non-regenerative and regenerative relays respectively called as amplify-and-forward (AF) and decode-and-forward (DF) [15]. In the AF protocol, the relay simply amplifies the received signal and forwards it to the destination or next relay. The amplifying process may be based on the full or partial channel state information (CSI), namely variable-gain AF relaying [16] and fixed-gain AF relaying [17, 18], respectively. Despite the performance enhancement by means of the additional relay node, the cooperative communications increase the required transmission time when the relay operates in conventional HD mode. In order to benefit from the advantages of cooperative networks described above without increasing the total transmission time, FD communications should be considered.

When there is a single relay operating in HD mode and the information is conveyed in one-way (OW) from source to destination, two time-slots are required. On the other hand, the total transmission time can be decreased to one time-slot by employing a relay operating in FD mode [19]. Considering the transmission is established via AF relaying, there are several works in the literature about one-way full-duplex AF relaying (OWFD-AF) systems. In [20], optimal power allocation (PA) problem and capacity of the variable-gain OWFD-AF systems are investigated. In [21], outage

probability of the variable-gain OWFD-AF systems with direct link between the source and destination is studied over Rayleigh fading channels and shown that the direct link provides diversity gain. The outage probability of the variable-gain OWFD-AF systems with direct link is derived for Rayleigh fading channels in [22] under processing delay and residual LI. It is worthwhile to note that the source-relay link is assumed fading-free in [22]. An optimal PA analysis based on minimizing the PEP is carried out for the fixed-gain OWFD-AF systems in [23] and shown that unequal PA presents better error performance than those using equal power in the transmitters. The error performance and diversity of uncoded and coded fixed-gain OWFD-AF systems with direct link are investigated in [24]. The outage performance of the variable-gain OWFD-AF systems over Nakagami- m channel is recently analyzed in [25] and the exact outage probability expression is derived in terms of extended generalized bivariate Meijer-G function (EGBMGF). In [25], an approximate outage performance expression is also obtained in closed-form by ignoring additive noise in the source-relay channel. Besides, the asymptotic outage probability, asymptotic ergodic capacity and ergodic capacity lower-bound expressions are presented. OWFD-AF systems with multiple relays are also investigated (e.g., in [19, 26, 27]) and shown that the performance can be enhanced significantly by using either all the relays or the selected relay.

For the two-way (TW) cooperative networks with two sources and one relay, four time-slots are required when all nodes operate in HD mode. In that configuration, the relay node listens to the transmitted information from the sources in the first and the third time-slots and retransmits the received signals in the second and fourth time-slots, respectively [28]. In the literature, there are two well-known solutions in order to decrease the total transmission time to two time-slots: physical-layer network coding (PLNC) technique (e.g., [28–30]) and FD communications. For the further improvement, TW transmission can be accomplished in a single time-slot by combining PLNC technique and FD communications, where all nodes operate in FD mode. There are a few works in the literature about two-way full-duplex AF relaying systems, which is called as TWFD-AF. By taking into consideration of the imperfect channel state information (I-CSI), the expressions for exact outage probability and its approximation are respectively obtained in a single-integral form and closed-form over

Rayleigh fading channels in [31]. The relay selection for the TWFD-AF systems over Rayleigh fading channels is investigated in [32] where all residual LI channels are modeled as additive white Gaussian noise (AWGN). In [33], the hybrid relay mode selection among the conventional two-way half-duplex AF relaying (TWHD-AF), OWFD-AF, PLNC and TWFD-AF is combined with the PA method. The beneficial effects of the relay transmit power control on the residual LI is demonstrated in [34]. For I -networks (three node) and Y -networks (four node), the outage probability and average rate performance of TWFD-AF systems are investigated in [35]. In that paper, it is demonstrated that I -networks and Y -networks accomplish two-way transmission in one and two time-slots, respectively. The spectral and energy efficient TWFD-AF systems with massive multiple-input multiple-output (MIMO) are analyzed in [36]. In [37], the joint relay and receive beamforming design is investigated in TWFD-AF MIMO systems so as to minimize the effect of residual LI.

As mentioned above, FD transmission doubles the spectral efficiency with respect to the conventional HD transmission. In order to enhance the spectral efficiency and reliability of FD systems, MIMO techniques, such as V-BLAST architecture, can be employed. However, when there are more than one active transmit antennas in the MIMO systems, it is also required to cope with the inter channel interference (ICI) at the receiver side. Therefore, the complexity of the maximum likelihood (ML) detection algorithm exponentially increases with the number of transmit antennas under ICI [38]. The necessity of ICI cancellation process can be avoided by using a technique named as spatial modulation (SM) proposed in [38] by activating only one transmit antenna for one symbol period. In SM, the information bits are conveyed via not only the modulated M -ary signals but also the transmit antenna indices. Moreover, as a special case of SM, all the information bits can be conveyed via the transmit antenna indices by using a constant known signal at the receiver (e.g., +1) instead of modulated M -ary signals, which is known as space-shift-keying (SSK) [39–41]. In addition to the low-complexity detection, SM technique with the optimal ML detection algorithm [42] provides better error performance compared to the V-BLAST architecture under the assumption of either perfect channel state information (P-CSI) or I-CSI available at the receiver as shown in [42] and [43], respectively. Furthermore, SM technique decreases number of the required transmit radio frequency (RF) chains, and therefore

it is an energy-efficient transmission scheme. These properties of SM make it a strong candidate for the next-generation wireless systems [44]. In the literature, there are a few works that combine bi-directional FD transmission with SM, which is named as two-way full-duplex spatial modulation (TWFD-SM) systems. In [45], the outage and ergodic maximum mutual information of TWFD-SM systems with a 2×2 antenna configuration are investigated, where the transmitter is determined according to the SM mapper and the remaining antenna is used as the receive antenna by using the RF switches. By using differential SM, [46] represents that the LI cancellation in the TWFD-SM systems can be further enhanced for 2×2 MIMO structure. The performance of two-way FD transmission with coordinate interleaved orthogonal design (CIOD) and SM technique is individually analyzed for the uplink and downlink scenarios in [47]. Moreover, the FD transmission with SM technique in relay aided cooperative networks are considered in [48] and [49], where DF protocol is employed at the relay node.

In order to overcome the limited battery lifetime and necessity of the external power supply, simultaneous wireless information and power transfer (SWIPT) is proposed in [50]. It has recently gained great attention, which increases exceedingly energy efficiency in wireless networks and prolongs the battery lifetime [50–52]. In contrast to the conventional energy harvesting techniques (e.g., wind, solar, vibration, etc.), it is shown that RF signals can be also used for the energy harvesting. In SWIPT, therefore, the information bearing RF signals are not only used to recover the transmitted information but also used for the energy harvesting, which leads to an interesting trade-off about which portion of the RF signals is utilized for information and power transfer [50, 53, 54]. There are two common energy harvesting protocols as time-switching (TS) and power-splitting (PS). In the TS protocol, the receiver first harvests its energy from the transmitted RF signals, then the remaining time is utilized to recover the transmitted information [54]. Whereas the receiver uses a portion of the RF signals for energy harvesting, at the same time, the information detection is performed via the remainder of RF signals in the PS protocol [54]. In the literature, there are a few works analyzing SWIPT for one-way (e.g., in [55–58]) and two-way (e.g., in [59–61]) cooperative networks in the presence of FD transmission.

1.3 Contributions of Thesis

In this thesis, the performance of mainly four different wireless communication systems using FD transmission is investigated:

- Firstly, the outage performance of the fixed-gain OWFD-AF systems over Rayleigh fading channels is investigated and the exact outage probability expression is derived in a single-integral form. It is illustrated that the fixed-gain relaying outperforms the variable-gain relaying for OWFD-AF systems. Then, the outage probability of the variable-gain and fixed-gain OWFD-AF systems is analyzed over Nakagami- m fading channels. The exact outage probability expressions for both schemes are obtained in a single-integral form. Moreover, the lower-bound and asymptotic outage probability expressions are found in a closed-form. Afterwards, five different relay selection policies are integrated into the OWFD-AF systems in order to improve the outage performance when more than one relay are available in the system. For each relay selection algorithm, the exact, lower-bound and asymptotic outage probability expressions are derived.
- Secondly, the exact, lower-bound and asymptotic outage probability expressions are obtained for the variable-gain TWFD-AF systems over Nakagami- m fading channels. Then, the relay selection algorithm is applied to the same system so as to enhance the outage performance and the theoretical expressions are generalized for the relay selection case.
- Thirdly, the error performance of TWFD-SM systems under both P-CSI and I-CSI is investigated over Rayleigh fading channels. By using the closed-form upper-bound BER expression, it is demonstrated that TWFD-SM can provide better error performance in comparison with the conventional two-way half-duplex spatial modulation (TWHD-SM) systems. Furthermore, the error performance of one-way full-duplex space-shift-keying with AF relaying (OWFD-SSK-AF) over Rayleigh fading channels is analyzed and the upper-bound BER expression is derived in a closed-form. The BER performance of the conventional one-way half-duplex space-shift-keying with AF relaying (OWHD-SSK-AF) is compared with the proposed OWFD-SSK-AF systems.

- As the final study of this thesis, a new two-way cooperative system with FD transmission is proposed, which uses SM and PLNC techniques at the sources and AF relaying and PS protocol for SWIPT at the relay. Hence, the proposed system is named as two-way full-duplex spatial modulation with amplify-and-forward and power-splitting (TWFD-SM-AF-PS) system. An upper-bound BER expression for the wireless-powered TWFD-SM-AF-PS system is derived in a closed-form. The obtained BER expression is also valid for two-way half-duplex spatial modulation with amplify-and-forward and power-splitting (TWHD-SM-AF-PS) systems.





2. BASIC CONCEPTS OF WIRELESS COMMUNICATIONS

This chapter introduces a brief overview for the basic concepts of wireless communications, which are essential in order to explain the proposed FD communication systems.

2.1 Statistical Models for Wired and Wireless Channels

Some statistical models are required in order to model and evaluate the performance of wired and wireless communication systems. Now, assume that there is a one transmitted and one receiver available in the system, namely a single-input single-output (SISO) architecture. In the wired communications, it is convenient to model the transmission channel as additive white Gaussian channel (AWGN) because the transmitted signal is collected with the additional noise at the receiver side as shown in Figure 2.1. However, in the wireless communications, the collected signal at the receive antenna is the combination of multiple reflected, scattered and diffracted signals due to the surrounding objects, such as buildings, trees and clouds as illustrated in Figure 2.2. Therefore, there is a multipath propagation in the wireless radio environment including the line-of-sight (LOS) and non-line-of-sight (NLOS) paths. The well-known wireless channel models can be sorted as Rayleigh, Nakagami- m and Rician fading channels. In this thesis, we have used Rayleigh and Nakagami- m fading channels and properties of these models are introduced in the following sections.

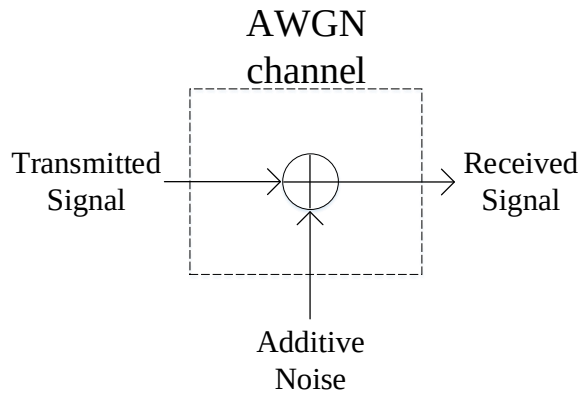


Figure 2.1 : Block diagram of AWGN channels.

2.1.1 AWGN channels

As shown in Figure 2.1, the received signal is exposed to an additional noise term in wired communications. When the noise term is analyzed, it is 'additive' because it is added to the transmitted signal, it is 'white' because it has uniform power distribution across all the frequency domain and it is 'Gaussian' because it is distributed according to Gaussian distribution in time domain. Hence, it is called as AWGN channels. The received signal after filtering can be modeled as

$$y = \sqrt{P}x + n \quad (2.1)$$

where x is the information bearing transmitted signal having unit-average-power, P is the transmission power of the transmitter node and n is the complex Gaussian noise occurred at the receiver due to the thermal noise generated by the electronic components [13]. The noise term n is distributed according to $\mathcal{CN}(0, \sigma^2)$. Therefore, the complex noise term can be generated as

$$n = n_R + jn_I \quad (2.2)$$

where n_R and n_I are the real and imaginary parts of the noise, which are identically and independent distributed (i.i.d.) according to $\mathcal{N}(0, \sigma^2/2)$. Therefore, their PDF expressions can be written as

$$f_{n_R}(x) = f_{n_I}(x) = \frac{1}{\sqrt{\pi\sigma^2}} \exp\left(-\frac{x^2}{4\sigma^2}\right). \quad (2.3)$$

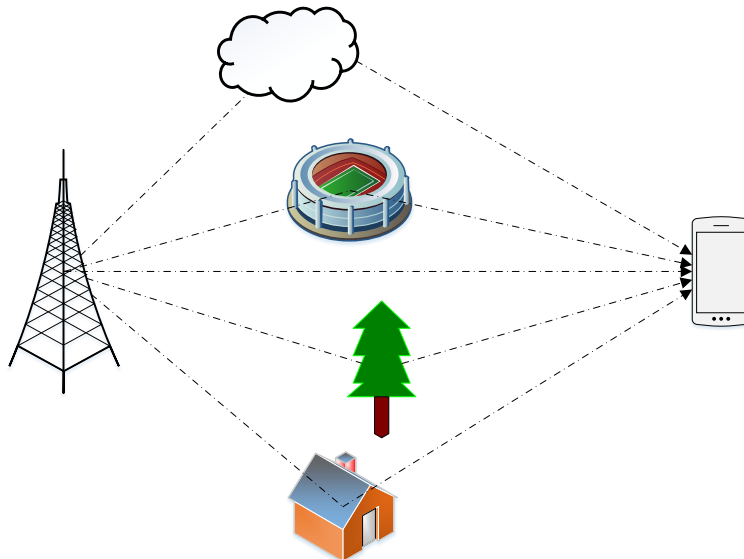


Figure 2.2 : Wireless multipath fading environment.

Then, the deterministic signal-to-noise ratio (SNR) at the receiver is defined as $\gamma = P/\sigma^2$. In Figure 2.3, an illustrative example for the BER performance of M -QAM signals over AWGN channels is plotted versus P/σ^2 . The theoretical curves for 4-QAM, 8-QAM, 16-QAM, 64-QAM and 256-QAM signals are generated by using [62, Eq. (16)]. As seen from the numerical results, slope of the BER curves representing the diversity order increases for the higher values of P/σ^2 as expected because the asymptotic diversity order of AWGN channels diverges to infinity.

2.1.2 Rayleigh fading channels

In wireless communications, when the receiver collects only the reflected NLOS multipath signals and there is no LOS between the transmitter and receiver due to the surrounding objects, these channels are typically classified as Rayleigh fading channels. In general, the received signal in the fading channels is given by

$$y = h\sqrt{P}x + n \quad (2.4)$$

where h represents the complex fading channel coefficient. As seen, h is equal to one for (2.1) because there is no fading effect in the wired channels. For Rayleigh fading channels, h is distributed according to $\mathcal{CN}(0, \Omega)$, where Ω is the average power of fading channel coefficient, i.e., $E\{|h|^2\} = \Omega$. Similar to the noise term n , h can be generated as follows

$$h = h_R + jh_I \quad (2.5)$$

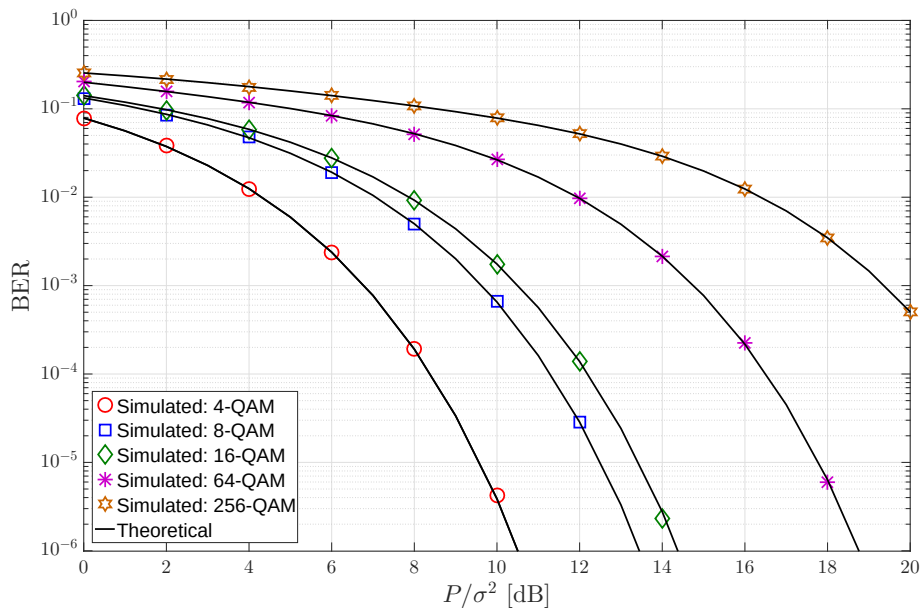


Figure 2.3 : BER performance of M -QAM signals over AWGN channels.

where h_R and h_I are the real and imaginary parts of the channel coefficient, which are i.i.d. according to $\mathcal{N}(0, \Omega/2)$. Therefore, the envelope of h follows Rayleigh distribution, which is given by

$$f_{|h|}(x) = \frac{2x}{\Omega} \exp\left(-\frac{x^2}{\Omega}\right), \quad x \geq 0. \quad (2.6)$$

Then, the instantaneous and average SNR for the fading channels can be written as $\gamma = P|h|^2/\sigma^2$ and $\bar{\gamma} = P\Omega/\sigma^2$, respectively. In Rayleigh fading channels, one can easily show that γ is an exponentially distributed random variable, so its PDF and CDF are respectively given by

$$\begin{aligned} f_{\gamma}(x) &= \frac{1}{\bar{\gamma}} \exp\left(-\frac{x}{\bar{\gamma}}\right), \quad x \geq 0, \\ F_{\gamma}(x) &= 1 - \exp\left(-\frac{x}{\bar{\gamma}}\right), \quad x \geq 0. \end{aligned} \quad (2.7)$$

In Figure 2.4, the BER performance comparison between AWGN and Rayleigh fading channels is represented for BPSK modulation. The theoretical curves for AWGN and Rayleigh fading channels are produced by using [63, Eq. (2.15)] and [63, Eq. (2.18)], respectively. It is shown that the error performance of wireless communications degrades due to the multipath fading environment in comparison to the wired communications. For instance, in order to achieve the same BER value of 10^{-5} in the both wired and wireless networks, the transmission power has to be increased 34.5 dB in Rayleigh fading channels compared to AWGN channels.

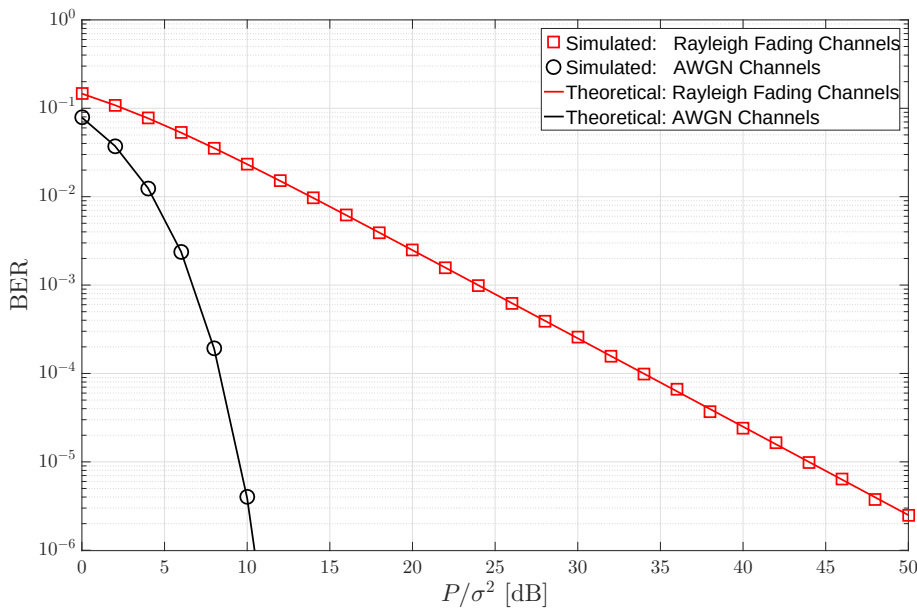


Figure 2.4 : BER performance of BPSK modulation over AWGN and Rayleigh fading channels.

Table 2.1 : The characteristic of Nakagami- m fading channels according to values of m .

The values of m	The characteristics of fading channels.
$0.5 \leq m < 1$	Poor conditions compared to Rayleigh fading channels.
$m = 1$	Equivalent to Rayleigh fading channels.
$m > 1$	Better conditions compared to Rayleigh fading channels. It can be also used for modelling Rician fading channels.

2.1.3 Nakagami- m fading channels

The second wireless channel model used in this thesis is Nakagami- m fading channels, which is a more general model [64]. According to the value of m , Nakagami- m fading channels provide to analyze more flexible wireless channel conditions. For instance, Nakagami- m fading channel model is not only equivalent model Rayleigh fading channels (e.g., $m = 1$), but also equivalent model better (e.g., $m > 1$) and even worse (e.g., $0.5 \leq m < 1$) channel conditions compared to Rayleigh fading channels. Therefore, Nakagami- m fading channels is a more general model with respect to the Rician fading channels [63]. Additionally, the PDF and CDF expressions of Nakagami- m distribution are more tractable compared to Rician distribution. A detailed explanation for Nakagami- m fading channels is given in Table 2.1.

The complex channel coefficient in Nakagami- m fading channels can be also defined as in (2.5), where the distribution of h_R and h_I varies according to value of m . For instance, when m is an integer, h_R and h_I can be generated as

$$h_R = \pm \sqrt{\sum_{i=1}^m \chi_i^2}, \quad h_I = \pm \sqrt{\sum_{i=m+1}^{2m} \chi_i^2} \quad (2.8)$$

where $\chi_i \sim \mathcal{N}(0, \frac{\Omega}{2m})$. From above expression, it is also obvious that Nakagami- m channels are equivalent to Rayleigh channels, when $m = 1$. For the arbitrary values of m , the fading channel coefficients can be also generated by using available built-in functions within common mathematical software packages, e.g., MATLAB, MATHEMATICA or MAPLE.

For Nakagami- m fading channels, the envelope of channel coefficient is distributed according to Nakagami- m distribution and its PDF is given by

$$f_{|h|}(x) = \frac{2m^m x^{2m-1}}{\Gamma(m) \Omega^m} \exp\left(-\frac{mx^2}{\Omega}\right), \quad x \geq 0, \quad m \geq 0.5 \quad (2.9)$$

where $\Gamma(\cdot)$ is the gamma function [65, Eq. (8.310.1)]. Then, γ is a gamma distributed random variable, so its PDF and CDF expressions are respectively given by

$$\begin{aligned} f_{\gamma}(x) &= \frac{m^m x^{m-1}}{\Gamma(m) \bar{\gamma}^m} \exp\left(-\frac{mx}{\bar{\gamma}}\right), \quad x \geq 0, \quad m \geq 0.5, \\ F_{\gamma}(x) &= \frac{\gamma(m, mx/\bar{\gamma})}{\Gamma(m)}, \quad x \geq 0, \quad m \geq 0.5 \end{aligned} \quad (2.10)$$

where $\gamma(\cdot, \cdot)$ is the lower incomplete gamma function [65, Eq. (8.350.1)].

The BER performance of BPSK modulation over AWGN and Nakagami- m fading channels is demonstrated in Figure 2.5, where $m \in \{0.5, 1, 1.5, 2, 2.5, 3, 4, 5\}$. The numerical results obtained via Monte Carlo simulations are perfectly matched with the results given in [13, Figure 13.3–2]. It is shown that the BER performance of Nakagami- m fading channels enhances and converges to AWGN channels as long as the value of m increases just as expected. Moreover, the BER performance of Rayleigh fading channels plotted in Figure 2.4 is exactly the same with $m = 1$ case in Nakagami- m fading channels as expected.

2.2 MIMO Systems

The simplest transmission scheme for the wireless radio environment is SISO channel, where there is only one transmit and receive antenna. The complexity of SISO systems is relatively low, however, it is hard to obtain either the diversity or multiplexing gain.

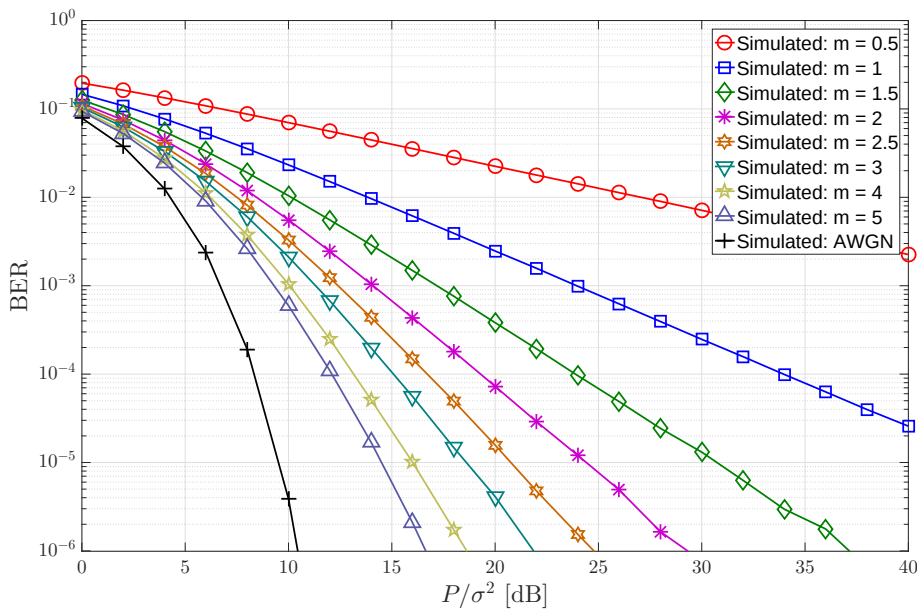


Figure 2.5 : BER performance of BPSK modulation over AWGN and Nakagami- m fading channels for $m \in \{0.5, 1, 1.5, 2, 2.5, 3, 4, 5\}$.

In order to improve the reliability of wireless transmission and cope with destructive effect of fading, it is required to employ some diversity techniques, e.g., time, frequency or antenna diversity [63]. The main idea of diversity techniques is to use two or more independent samples of the information bearing signals in order to recover the transmitted bits, which ensures more correct detection process. Although time and frequency diversity techniques provide better performance, there is a loss in the spectral efficiency due to the additional time-slots and frequency bands, respectively. Moreover, time diversity cannot be obtained in the slow flat-fading channels due to the correlated channel coefficients in the consecutive time intervals. On the other hand, the antenna diversity uses multiple antennas separated physically by a proper distance instead of wasting time and frequency resources. Therefore, the antenna diversity does not induce any loss in spectral efficiency which makes it more popular among others, which can be used at the receiver or transmitter. So, it is categorized as receive antenna diversity and transmit antenna diversity. The receive antenna diversity can be employed by using different receive antennas. In order to obtain uncorrelated signals at each receive antenna, it is required to take into consideration the antenna spacing. In order to detect the transmitted information more precisely, the uncorrelated received signals can be combined by using different algorithms at the receiver, e.g., selection combining, switched combining, equal gain combining and maximal ratio combining (MRC) [63]. In transmit diversity, the transmitter is equipped with multiple transmit antennas and the replicas of information bits are conveyed through multiple antennas [63].

In order to cope with the severe fading effect of the wireless radio environment and increase the total transmission rates, multiple transmit and receive antennas can be deployed at the system, which is called as MIMO systems are proposed in [66]. In the literature, the MIMO systems are well-studied. A basic system model for MIMO channels is given in Figure 2.6, where N_T and N_R are the number of transmitter and receiver, respectively. Then, the channel matrix can be written as

$$\mathbf{H} = \begin{bmatrix} h_{1,1} & h_{1,2} & \cdots & h_{1,N_T} \\ h_{2,1} & h_{2,2} & \cdots & h_{2,N_T} \\ \vdots & \vdots & \ddots & \vdots \\ h_{N_R,1} & h_{N_R,2} & \cdots & h_{N_R,N_T} \end{bmatrix}. \quad (2.11)$$

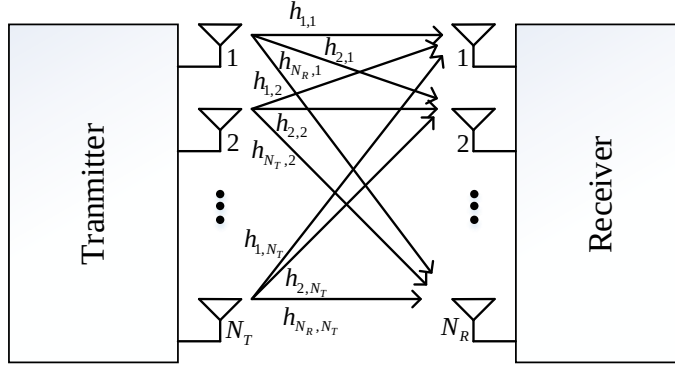


Figure 2.6 : MIMO system model.

In MIMO systems, transmitter side can be configured either to convey different information bits or to obtain transmit diversity. For the higher transmission rates, MIMO systems can be used for multiplexing, e.g., V-BLAST architecture [67]. On the other hand, the replicas of information bits can be transmitted from transmit antennas, when it is required to improve the reliability of wireless transmission, e.g., Alamouti space-time code [63]. Therefore, there is a fundamental trade-off between multiplexing gain and diversity gain in the MIMO systems.

2.3 Cooperative Communications

As introduced in the previous section, MIMO systems are precious in terms of boosting the transmission rates and the quality of wireless communications. However, it may be hard to place multiple antennas in mobile devices owing to both their smaller size and lower processing powers. Therefore, instead of using point-to-point communication techniques, cooperative communications based on cooperation of the users through relay nodes has gained great attention in recent years since it provides coverage extension, capacity, outage and error performance improvement [14–18]. In this context, the source (S) transmits its information to the destination (D) directly and with the help of one or more relays (R) as demonstrated in Figure 2.7. The relay nodes are generally categorized as amplify-and-forward (AF) and decode-and-forward (DF) relays [15].

For the cooperative system given in Figure 2.7, one-way transmission can be accomplished in two time-slots, when the relay operates in the conventional HD mode. Typically, in the first time-slot, the source transmits its information to the relay and destination. In the second time-slot, the source remains silent and the relay retransmits

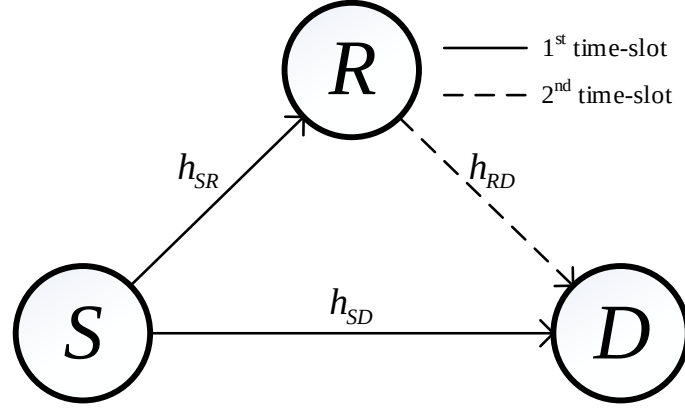


Figure 2.7 : Cooperative networks system model.

either amplified or decoded signal to the destination. Therefore, the destination can recover the transmitted information by using two different signals, which provides a diversity gain as in the MIMO systems.

2.3.1 Amplify-and-forward relaying

In the AF protocol, the received signal at the relay node is simply forwarded to the destination after multiplying with an amplification factor. In the first time-slot, the received signal at the relay and destination are respectively given by

$$r = h_{SR}\sqrt{P_S}x + n_R, \quad (2.12)$$

$$y_{SD} = h_{SD}\sqrt{P_S}x + n_{SD} \quad (2.13)$$

where h_{SR} and h_{SD} are the fading channel coefficients between the source-relay and source-destination links, P_S is the transmission power of source, x is the transmitted signal from source with the unit-average-power, n_R and n_{SD} are the complex Gaussian noise terms occurred at the relay and destination, which are distributed according to $\mathcal{CN}(0, \sigma_R^2)$ and $\mathcal{CN}(0, \sigma_{RD}^2)$, respectively. The average powers of h_{SD} and h_{SR} are denoted as Ω_{SD} and Ω_{SR} . According to the AF protocol, the relay node amplifies the received signal by using an amplification factor, then transmits it to the destination in the second time-slot. The received signal at the destination is written as

$$y_{RD} = h_{RD}s + n_D \quad (2.14)$$

where h_{RD} is the fading channel coefficients between the relay-destination link with the average power of Ω_{RD} , $s = Gr$ is the transmitted signal from the relay node, G is the amplification factor (variable [15] or fixed [18]) and n_D is the complex Gaussian

noise term distributed according to $\mathcal{CN}(0, \sigma_D^2)$. By using (2.14) and $s = Gr$, the received signal at the destination in the second time-slot can be rewritten as

$$y_{RD} = \underbrace{Gh_{SR}h_{RD}\sqrt{P_S}x}_{\text{Information Signal}} + \underbrace{Gh_{RD}n_R + n_D}_{\text{Noise Terms}}. \quad (2.15)$$

By applying the ML detection rule, the received signals respectively given in (2.13) and (2.15) can be combined by using the MRC technique via noise normalization as follows [15, Eq. (3)]

$$y = \frac{h_{SD}^* y_{SD}}{\sigma_{SD}^2} + \frac{(Gh_{SR}h_{RD})^* y_{RD}}{G^2|h_{RD}|^2 + \sigma_D^2}. \quad (2.16)$$

After some mathematical derivations, the end-to-end SNR of the systems is given by

$$\gamma_{end} = \gamma_{SD} + \frac{\gamma_{SR}\gamma_{RD}}{\gamma_{RD} + \frac{P_R}{G^2\sigma_R^2}} \quad (2.17)$$

where $\gamma_{SD} = P_S|h_{SD}|^2/\sigma_{SD}^2$, $\gamma_{SR} = P_S|h_{SR}|^2/\sigma_R^2$, $\gamma_{RD} = P_R|h_{RD}|^2/\sigma_D^2$ and P_R is the transmission power of the relay node located inside the amplification factor, which will be expressed in the following sections for the variable-gain and fixed-gain AF relaying. It can be easily seen from above expression, the end-to-end SNR depends on the value of amplification factor. Therefore, variable-gain and fixed-gain AF relaying should be analyzed individually.

2.3.1.1 Variable-gain AF relaying

In order to normalize the instantaneous transmission power at the relay as P_R , the amplification factor for variable-gain AF relaying is chosen as [14, Eq. (9)]

$$G = \sqrt{\frac{P_R}{P_S|h_{SR}|^2 + \sigma_R^2}}. \quad (2.18)$$

By substituting (2.18) into (2.17), the end-to-end SNR expression becomes

$$\gamma_{end} = \gamma_{SD} + \frac{\gamma_{SR}\gamma_{RD}}{\gamma_{SR} + \gamma_{RD} + 1}. \quad (2.19)$$

In order to represent the performance improvement of the cooperative networks with respect to non-cooperative case, the BER performance of both systems using BPSK over Rayleigh fading channels is plotted in Figure 2.8. In the non-cooperative case, the communication is established via only the direct link between the source and

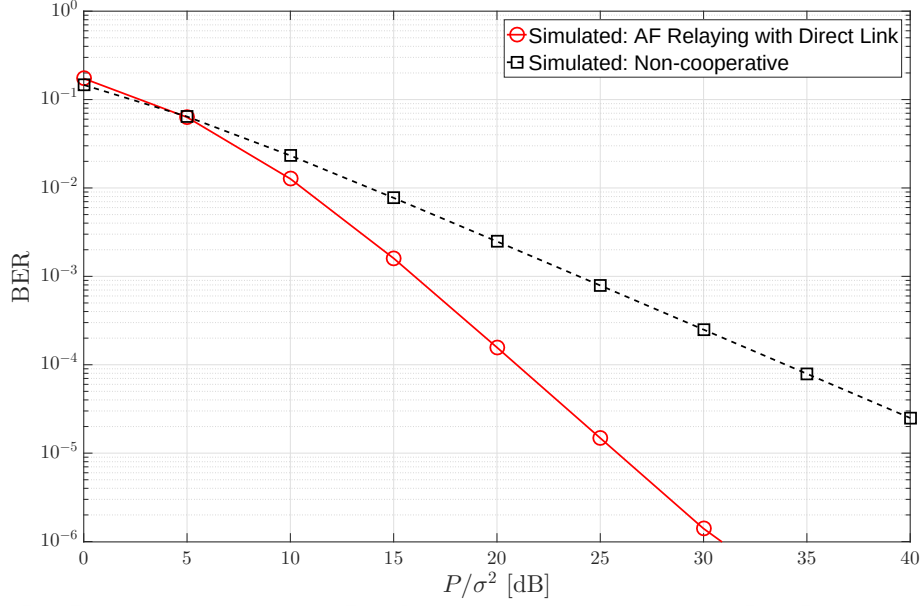


Figure 2.8 : BER performance of BPSK modulation over Rayleigh fading channels for cooperative and non-cooperative transmission scheme.

destination, whereas, both the direct link and relay are participated into transmission in the cooperative networks. During the Monte Carlo simulations, we assume that $P = P_S + P_R$, $P_S = P_R$, $\sigma^2 = \sigma_{SD}^2 = \sigma_R^2 = \sigma_D^2$ and $\Omega_{SD} = \Omega_{SR} = \Omega_{RD} = 1$ as in [15, Figure 2]. It is seen that cooperative transmission scheme provides considerably better error performance and increases the diversity order of system in comparison to the non-cooperative case. To illustrate, the cooperative scheme provides 13 dB performance improvement at a BER value of 10^{-4} .

2.3.1.2 Fixed-gain AF relaying

In contrast to variable-gain AF relaying, the instantaneous CSI of the source-relay link is not required for the fixed-gain AF relaying. Alternatively, the relay needs Ω_{SR} as the average power of the source-relay link for the amplification process. Therefore, the fixed-gain AF relaying is also called as "blind" relaying [17]. So as to normalize the average transmission power at the relay as P_R , the amplification factor for fixed-gain AF relaying is chosen as [18, Eq. (8)]

$$G = \sqrt{\frac{P_R}{P_S \Omega_{SR} + \sigma_R^2}}. \quad (2.20)$$

By combining (2.17) and (2.20), the end-to-end SNR expression for the fixed-gain AF relaying can be easily found as

$$\gamma_{end} = \gamma_{SD} + \frac{\gamma_{SR}\gamma_{RD}}{\gamma_{RD} + C} \quad (2.21)$$

where $C = \bar{\gamma}_{SR} + 1$ and $\bar{\gamma}_{SR} = P_S \Omega_{SR} / \sigma_R^2$. So, the end-to-end SNR expressions for the variable-gain and fixed-gain AF systems are different from each other. The performance comparison between variable-gain and fixed-gain AF relaying systems is carried out in terms of the outage probability, which is evaluated by

$$P_{out} = P(\gamma_{end} \leq \gamma_{th}) \quad (2.22)$$

where γ_{th} is the threshold SNR value. The outage probability curves for the different values of γ_{th} over Rayleigh fading channels are represented in Figure 2.9, where $P_S = P_R$ and $\sigma_{SD}^2 = \sigma_R^2 = \sigma_D^2$ as in [17, Figure 2]. In addition to its simpler structure, the performance of the fixed-gain AF relaying is almost the same as that of the variable-gain AF relaying. Moreover, the fixed-gain relays can interestingly provide better outage performance with respect to the variable-gain relays, especially in the low SNR region.

It is worthwhile to mention that the instantaneous transmission power at the relay for the fixed-gain AF relaying varies according to the instantaneous CSI, which can be considered as a drawback of fixed-gain relays due to the necessity of a power amplifier having a large dynamic range. In [17], the effect of the dynamic range for the power amplifier is analyzed where the authors define a maximum transmission power for the fixed-gain relays. As demonstrated in [17, Figure 4], when the dynamic range of the power amplifier at the relay is limited with a maximum transmission power,

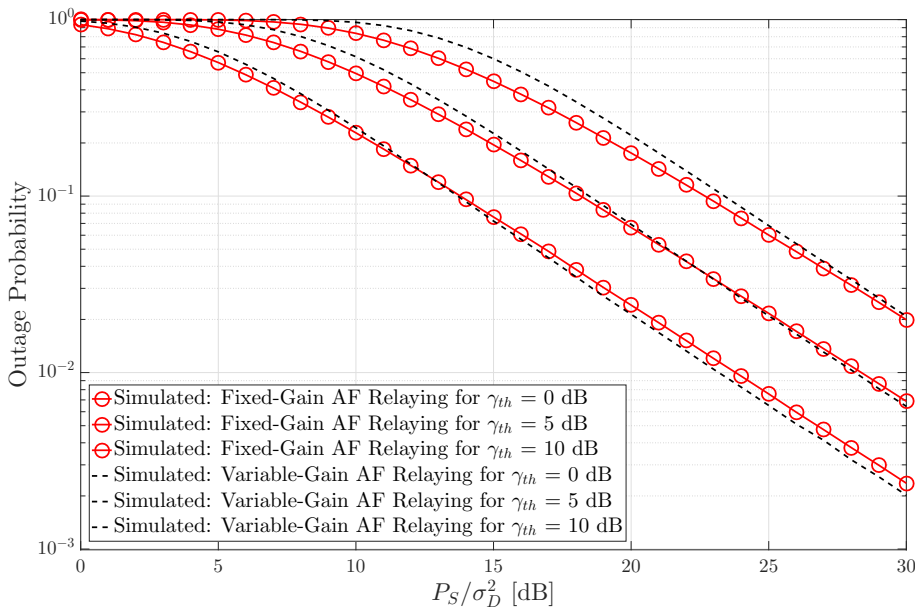


Figure 2.9 : Outage probability of the variable-gain and fixed-gain AF relaying over Rayleigh fading channels for different values of γ_{th} .

the performance of fixed-gain AF relaying is not changed at the medium and high SNR region with respect to the ideal (unrestricted) dynamic range case. Hence, the ideal amplifier assumption for the power amplifier is also used in [18, 68–70] so as to simplify the analytical analyses.

2.3.2 Decode-and-forward relaying

The second strategy as the DF relaying is a more complex technique, where the relay decodes the received information, then the decoded signal is modulated and retransmitted to the destination. In the scope of this thesis, only the AF relaying technique is employed at the relay node of the proposed cooperative systems. For further details of the DF relaying, the interested readers can see [14, 15].

2.4 Physical-Layer Network Coding

As mentioned earlier, when the relay operates in the conventional HD mode, two time-slots are required for one-way relay transmission as shown in Figure 2.7. On the other hand, the required number of time-slots are doubled for the two-way relay transmission as demonstrated in Figure 2.10(a), where there are now two sources ($S_i, i \in \{1, 2\}$) and one relay (R). It is assumed that there is no direct link between the sources and they communicate with each other via the relay node. The significant problem of that transmission scheme is the low spectral efficiency as a result of four time-slots requirement. In the literature, there is a well-known solution to this problem, which is named as physical-layer network coding (PLNC) [28–30].

In the PLNC technique, total transmission time is decreased to two time-slots as illustrated in Figure 2.10(b). Both S_1 and S_2 simultaneously transmits their information signals to the relay in the first time slot. Then, the relay bidirectionally forwards the received signal to the sources by using the AF protocol in the second time-slot. Since each source knows their own transmitted signals, they can remove the interference term easily. In order to express mathematically, the received signal from the relay at the first time-slot is given by

$$r = h_{S_1R}\sqrt{P_S}x_1 + h_{S_2R}\sqrt{P_S}x_2 + n_R \quad (2.23)$$

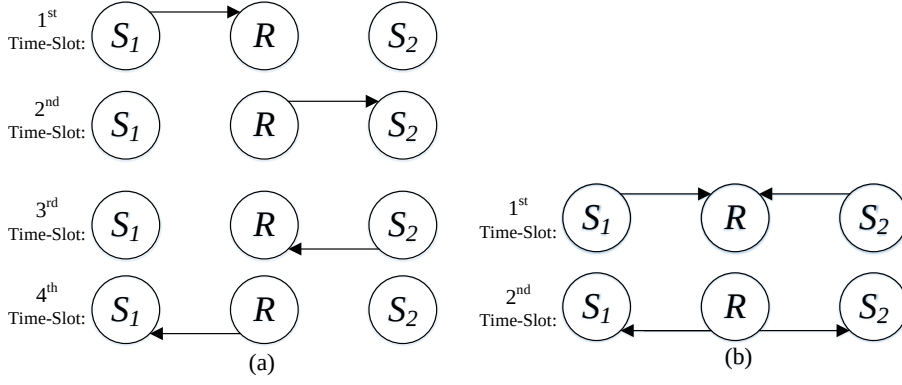


Figure 2.10 : Two-way relay transmission schemes in (a) four and (b) two time-slots.

where h_{S_1R} and h_{S_2R} are the reciprocal channel coefficients for $S_1 - R$ and $S_2 - R$ links, x_1 and x_2 are the unit-average-power signals transmitted from S_1 and S_2 , respectively. According to the AF protocol introduced above, the transmitted signal from the relay is again denoted as $s = Gr$. At the second time-slot, the received signal at S_i can be written as follows

$$\begin{aligned}
 y_i &= h_{S_iR}s + n_{S_i} \\
 &= \underbrace{h_{S_iR}h_{S_jR}G\sqrt{P_S}x_j}_{\text{Desired Signal}} + \underbrace{h_{S_iR}^2G\sqrt{P_S}x_i}_{\text{Removed by PLNC}} + \underbrace{h_{S_iR}Gn_R + n_{S_i}}_{\text{Noise}}
 \end{aligned} \quad (2.24)$$

where $j \in \{1, 2\}$, $j \neq i$ and n_{S_i} is the complex Gaussian noise term at S_i with the distribution of $\mathcal{CN}(0, \sigma_S^2)$. The term containing x_i inside the above expression can be removed by subtracting it [28]. After some mathematical manipulations, one can show that the instantaneous end-to-end SNR is equal to

$$\gamma_{end,i} = \frac{\gamma_{S_iR}\gamma_{S_jR}}{\gamma_{S_iR} + \frac{P_R}{\alpha G^2 \sigma_R^2}} \quad (2.25)$$

where $\gamma_{S_iR} = P_S|h_{S_iR}|^2/\sigma_R^2$ and $\gamma_{S_jR} = P_S|h_{S_jR}|^2/\sigma_R^2$.

When the variable-gain AF relaying is employed at the relay in order to normalize the instantaneous transmission power of the relay to P_R , the amplification factor should be $G = \sqrt{P_R/(P_S|h_{S_1R}|^2 + P_S|h_{S_2R}|^2 + \sigma_R^2)}$. So, the instantaneous end-to-end SNR at S_i can be obtained as [28, Eq. (8)]

$$\gamma_{end,i} = \frac{\alpha \gamma_{S_iR} \gamma_{S_jR}}{\gamma_{S_iR}(\alpha + 1) + \gamma_{S_jR} + 1} \quad (2.26)$$

where $\alpha = (P_R\sigma_R^2)/(P_S\sigma_S^2)$. On the other hand, the fixed-gain amplification factor is set to $G = \sqrt{P_R/(P_S\Omega_{S_1R} + P_S\Omega_{S_2R} + \sigma_R^2)}$ in order to adjust the average transmission power of the relay as P_R , where Ω_{S_1R} and Ω_{S_2R} are the average power

of the fading channel coefficients belongs to $S_1 - R$ and $S_2 - R$ links. So, the instantaneous end-to-end SNR expression at S_i becomes [30, Eq. (5)]

$$\gamma_{end,i} = \frac{\gamma_{S_i R} \gamma_{S_j R}}{\gamma_{S_i R} + C} \quad (2.27)$$

where $C = \bar{\gamma}_{S_i R} + (\bar{\gamma}_{S_j R} + 1) / \alpha$, $\bar{\gamma}_{S_i R} = P_R \Omega_{S_i R} / \sigma_S^2$ and $\bar{\gamma}_{S_j R} = P_S \Omega_{S_j R} / \sigma_R^2$. In terms of performance comparison between the four time-slots and two time-slots transmission schemes, the numerical results in [28] shows that the PLNC technique with two time-slot transmission duration ensures higher capacity.

2.5 Spatial Modulation

The MIMO techniques, such as V-BLAST architecture [67], is a promising solution so as to improve the spectral efficiency and reliability of wireless communication systems. However, when there are more than one active transmit antennas in the MIMO systems, the ICI occurs at the receiver side. Therefore, the complexity of ML detection rule exponentially increases due to the ICI problem. Nevertheless, the ICI problem can be avoided via spatial modulation (SM) proposed in [38], where only one transmit antenna is activated during each symbol period and this antenna is determined according to the input bits. Therefore, the information bits are carried by both the modulated M -ary signals (e.g., M -PSK/QAM) and transmit antenna indices.

The system model of SM is illustrated in Figure 2.11, where there are N_T transmit and N_R receive antennas and the MIMO channels are denoted by $\mathbf{H}$ with the size of $N_R \times N_T$. A random sequence of input bits are divided into $\log_2(N_T M) = \log_2(N_T) + \log_2(M)$ bit length groups. According to the SM mapper rule, the first $\log_2(N_T)$ bits are conveyed via the selected transmit antenna index, while, the remaining $\log_2(M)$ bits are transmitted with M -PSK/QAM modulation signals, where M is the size of signal constellation. In order to clarify the working

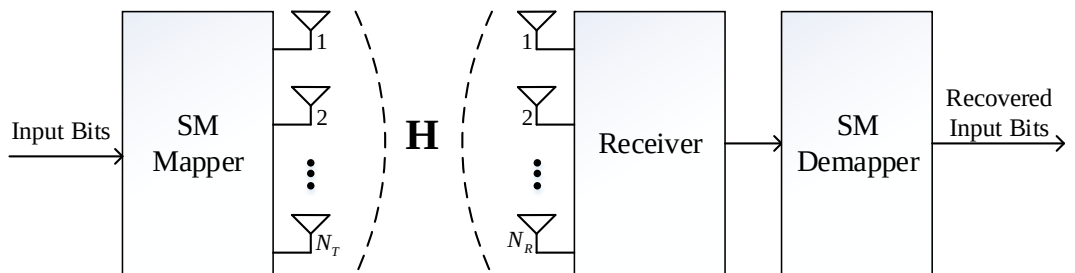


Figure 2.11 : Spatial modulation system model.

Table 2.2 : SM mapping rule for 3 bits transmission.

Input Bits	$N_T = 2, M = 4$		$N_T = 4, M = 2$	
	Antenna Index	QPSK Signal	Antenna Index	BPSK Signal
000	1	$1 + j$	1	1
001	1	$1 - j$	1	-1
010	1	$-1 + j$	2	1
011	1	$-1 - j$	2	-1
100	2	$1 + j$	3	1
101	2	$1 - j$	3	-1
110	2	$-1 + j$	4	1
111	2	$-1 - j$	4	-1

principle of the SM mapper, the selected antenna index and modulated signal are displayed in Table 2.2 for 3 bits transmission. For instance, when the number of transmit antennas is chosen as $N_T = 2$, QPSK signaling should be employed at the transmitter side. In that scenario, the first transmit antenna is utilized to transmit QPSK signals, when the first bit of the input stream is equal to zero. Otherwise, the transmission process is realized over the second transmit antenna. Similarly, the active transmit antenna index is determined via the first two bits, when $N_T = 4$, and the last bit is used for BPSK signals. Therefore, it is required to use RF switches in order to select the active transmit antenna in each symbol period. Furthermore, the number of required transmit RF chains is decreased to one, which makes SM an energy-efficient transmission scheme.

In SM, the vector of received signal is given by

$$\mathbf{y} = \sqrt{P}\mathbf{H}\mathbf{x} + \mathbf{n} \quad (2.28)$$

where P is the transmission power, $\mathbf{x}$ the $N_T \times 1$ unit-average-power SM signal vector and $\mathbf{n}$ is the $N_R \times 1$ complex Gaussian noise vector. The entries of $\mathbf{n}$ are i.i.d. as $\mathcal{CN}(0, \sigma^2)$. As expressed above, only one transmit antenna is active in each symbol period, so the transmitted signal vector can be written as

$$\mathbf{x} = [0 \ \cdots \ 0 \ \underbrace{x^q}_{k^{th} \text{ position}} \ 0 \ \cdots \ 0]^T \quad (2.29)$$

where k is the selected transmit antenna index according to the first $\log_2(N_T)$ bits and x^q denotes the q^{th} symbol from M -ary constellation according to following $\log_2(M)$

bits. By combining (2.28) and (2.29), the received signal can be rewritten as follows

$$\mathbf{y} = \sqrt{P}\mathbf{h}^k x^q + \mathbf{n} \quad (2.30)$$

where $\mathbf{h}^k$ is the k^{th} column of $\mathbf{H}$. For a given $\mathbf{H}$, the received signal is distributed as $\mathbf{y} \sim \mathcal{CN}(\sqrt{P}\mathbf{h}^k x^q, \sigma^2)$. When the optimal ML detector is applied at the receiver, the decision rule is given by [42]

$$[\hat{k}, \hat{q}] = \arg \min_{\tilde{k}, \tilde{q}} \left\| \mathbf{y} - \sqrt{P}\mathbf{h}^{\tilde{k}} x^{\tilde{q}} \right\|^2 \quad (2.31)$$

where $\hat{k}$ and $\hat{q}$ are the recovered transmit antenna and M -ary symbol index, respectively.

In space-shift-keying (SSK) as a special case of SM, all the information bits are conveyed via the transmit antenna indices by using a constant known signal at the receiver (e.g., $x^q = 1$) instead of M -ary signals. Hence, totally $\log_2(N_T)$ bits can be transmitted in the SSK technique. Moreover, SSK has a lower complexity detection algorithm in comparison to SM,

In Figure 2.12, Monte Carlo simulations results for the BER performance of SM, SSK and 8-QAM over Rayleigh fading channels are plotted for 3 bits transmission, where the number of receive antennas are assumed as $N_R = 4$ similar to [39, Figure 4]. In SM, the first two bits are conveyed via activated transmit antenna index, while the last bit is carried via BPSK signals. On the other hand, there are $N_T = 8$ transmit

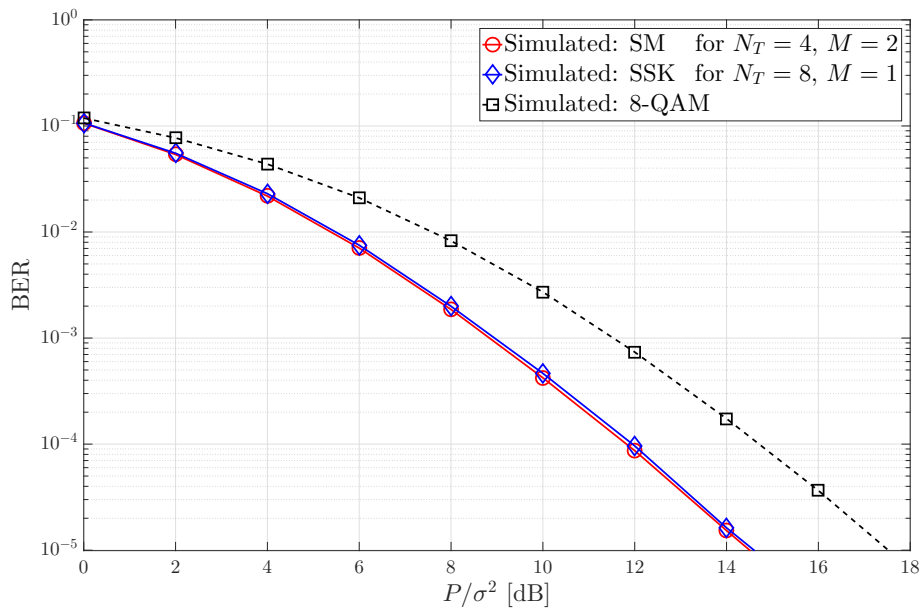


Figure 2.12 : BER performance comparison between SM, SSK and 8-QAM for 3 bits transmission and $N_R = 4$.

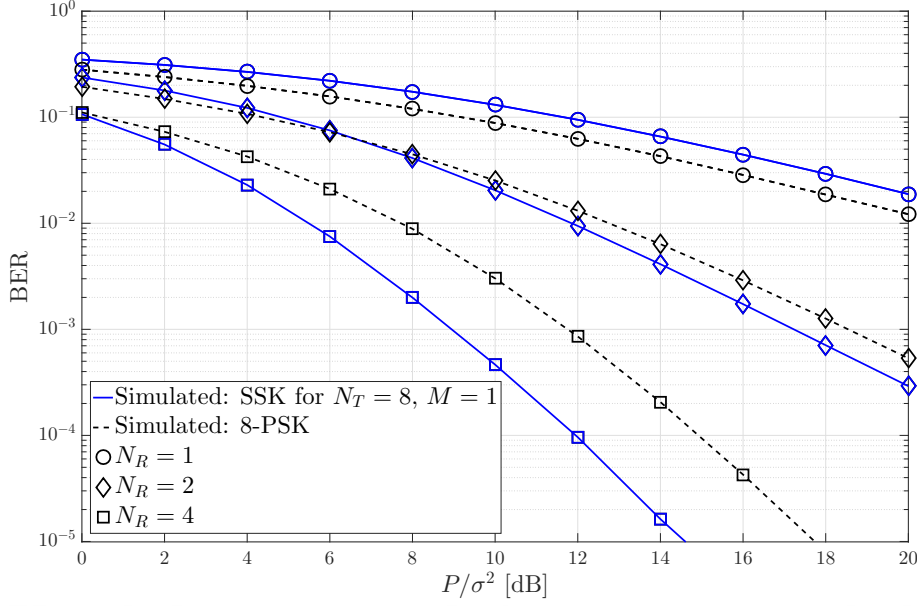


Figure 2.13 : BER performance comparison between SSK and 8-PSK for 3 bits transmission and $N_R \in \{1, 2, 4\}$.

antennas in SSK so as to convey all transmission via transmit antenna indices. As a reference system, all three information bits are transmitted via 8-QAM signals. The numerical results reveal that the SM and SSK provide a significant performance improvement compared to 8-QAM, e.g., there is approximately 2.3 dB performance difference between SM and 8-QAM at a BER value of 10^{-5} . Moreover, SSK technique provides almost identical error performance with SM.

The BER performance comparison of SSK and 8-PSK for different numbers of receive antennas is illustrated for 3 bits transmission in Figure 2.13, where $N_R \in \{1, 2, 4\}$ as in [39, Figure 6]. As in the previous numerical results, SSK provides 2.8 dB and 1.3 dB performance improvement with respect to 8-PSK for $N_R = 4$ and $N_R = 2$, respectively. On the other hand, 8-PSK outperforms SSK when $N_R = 1$. The most significant outcome is that SSK can provide better error performance as long as the number of receive antennas increases.

In a nutshell, it is easily seen that the advantage of SM arises when the number of receive antennas increases. Furthermore, in addition to the low-complexity detection, [38, 39, 42–44] shows that SM with optimal ML detector can also outperforms V-BLAST. These properties of SM make it a strong candidate for next-generation wireless systems [44].

3. FULL-DUPLEX COMMUNICATIONS

After elaborating the basic concepts of wireless communications, full-duplex (FD) communications as the main topic of this thesis is introduced throughout this chapter. The advantages and disadvantages with respect to its conventional counterpart half-duplex (HD) communications are also expressed in detail.

3.1 An Overview: Half-Duplex and Full-Duplex Communications

In general, the traditional wireless communication systems are not able to handle the simultaneous transmission and reception operations over the same frequency band due to the catastrophic loop-interference (LI) occurred at the receive antenna. Therefore, the transmission and reception of the information signals between two sources are divided into two time-slots (or two frequency bands) as shown in Figure 3.1, which is named as HD communications. In terms of the spectral efficiency, the HD systems use only half of the radio resources, so other 50 percent of the valuable resources is wasted just due to the HD transmission. On the other hand, the one-way and two-way relay systems operating in HD mode are illustrated in Figure 2.7 and Figure 2.10, respectively. As mentioned earlier, in spite of the several advantages of those cooperative systems, they are not spectral efficient compared to non-cooperative case. The swiftly increasing user demands on the limited radio resources can be fulfilled by more spectral efficient communication systems. Although FD communications, allowing simultaneous transmission and reception over the same frequency band, could be a possible technique in order to dramatically increase the spectral efficiency, it was thought as unpractical. According to Goldsmith [71, pp. 423]:

It is generally not possible for radios to receive and transmit on the same frequency band due to the interference that results. Thus, bi-directional systems must separate the uplink and downlink channels into orthogonal signaling dimensions, typically using time or frequency dimensions.

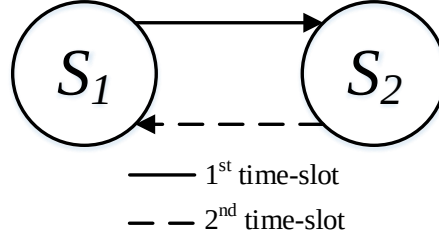


Figure 3.1 : Half-duplex system model.

However, FD communications became feasible thanks to the recent advances on both antenna technology and signal processing techniques for LI cancellation processes e.g., passive suppression, analog cancellation and digital cancellation [4–9]. Even though LI cannot be completely removed and the remaining residual LI limits the performance of FD communications, the experimental [8] and theoretical [4] studies in the literature shows that it can outperform the conventional HD communications.

The transmission schemes for two-way (non-cooperative), one-way relay and two-way relay systems with FD communications are represented in Figure 3.2, where the channels between relay (R) and sources (S_1 and S_2) and residual LI channels are represented via black solid and red dotted lines, respectively. It is worthwhile to note that all three transmission schemes with FD communications can be completed in a single time-slot. Hence, the required time-slots can be decreased via the FD transmission technique. The fundamental advantages of FD communications can be listed as follows:

- **Higher capacity:** Exploiting time and frequency resources enables to double the capacity of wireless communication systems in theory. Furthermore, the practical studies also show that FD transmission can provide higher capacity as long as the quality of LI cancellation increases [4].
- **Lower end-to-end delay:** For the cooperative systems, when the relays operate in FD mode, the end-to-end delay of those systems decreases [19]. For instance, one-way relay transmission can be completed in two time-slots with HD relays as shown in Figure 2.7. However, Figure 3.2(b) illustrates that the transmission can be accomplished in one time-slot with FD relays.
- **Network flexibility:** Thanks to the advance of FD communications, it can be possible to employ both FD and HD transmission in the different parts of

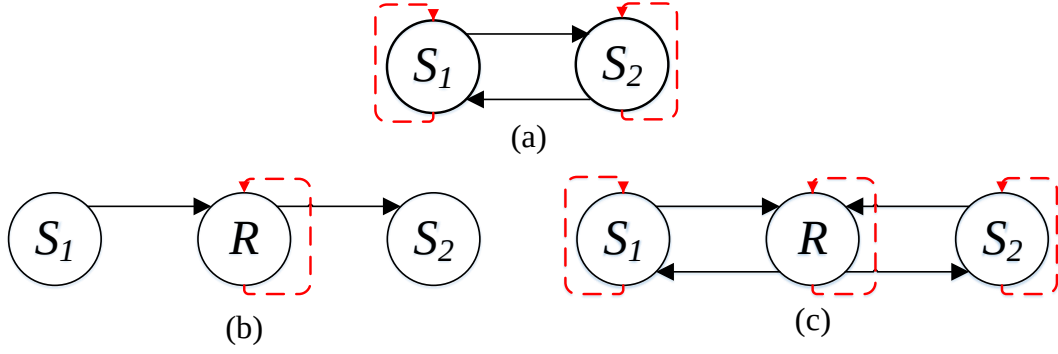


Figure 3.2 : Full-duplex system models: (a) two-way (non-cooperative), (b) one-way relay and (c) two-way relay systems.

heterogeneous networks. For example, the powerful base stations can handle the LI cancellation operations and work in FD mode, although it is hard to deploy FD transmission for all sources [72]. Another example is also given in 3.2(b), where the sources use HD transmission, while the relay operates in FD mode so as to complete the total transmission in one time-slot.

- **Increased secrecy:** One of the crucial benefit of FD transmission is the improved secrecy rate by the mixed radio signals due to the simultaneous transmission and reception operation [3]. Hence, the performance of eavesdropper channels degrades due to FD communications.

On the other hand, there are also some drawbacks of FD transmission, which are summarized as:

- **Imperfect loop-interference (LI) cancellation:** The strong LI component cannot be completely removed, even with multiple stages of cancellation. Therefore, the residual LI limits the performance of FD transmission [10].
- **Higher hardware requirement:** In order to enhance the quality of LI cancellation, which is vital for the FD communications, the hardware requirement of system increases due to various cancellation stages. [3, 4, 8].
- **Increased complexity and power consumption:** The additional hardware for the LI cancellation processes not only consumes more power, but also increases the system complexity.

Hence, there is an attractive trade-off between FD and HD communications and it will be analyzed on different systems throughout this thesis.

3.2 Loop-Interference Cancellation

In order to mitigate the effect of LI over FD transmission, there are two common LI cancellation techniques utilized at the receiver, e.g., passive suppression and active cancellation. Moreover, the active cancellation method is also categorized as analog cancellation and digital cancellation.

3.2.1 Passive suppression

The first LI cancellation method is passive suppression [9], where the basic idea is to attain path loss due to the physical separation/isolation between the transmit and receive antennas. There are several methods presented in the literature so as to impose passive suppression, such as, antenna separation [11] and antenna cancellation [12]. An example scenario for the antenna separation method is given in [11] as indoor-to-outdoor communications, where the base station is located at the outside and the users are inside the building. A relay equipped with one transmitter and one receiver tries to perform simultaneous transmission between the base station and the users. In order to apply the antenna separation, the transmit and receive antennas are placed inside and outside of the building. Additionally, RF absorbers can be also used for antenna separation as illustrated in [9, Figure 2]. For the antenna cancellation method, one receive and two transmit antennas are equipped and the positions of transmit antennas are adjusted according to the receive antenna [12]. When the difference between the distances according to receive antenna is exactly equal to the odd multipliers of a half wavelength, it creates a destructive effect on the LI due to the inverse propagation of radio waves.

3.2.2 Active cancellation

Since passive suppression methods are never perfect, the power of residual LI after passive suppression should be further decreased with the help of analog cancellation and digital cancellation, which are classified as active cancellation techniques. In order to decrease the amount of LI power, the estimated LI signal is subtracted from the

received signal either at the RF band or at the analog baseband stage during the analog cancellation operation. [8] shows that the quality of LI cancellation is highly dependent to the performance of analog cancellation. To achieve further improvement for FD communications, the power of residual LI should be reduced in the digital domain via digital cancellation technique.

After applying both passive and active cancellation techniques, the practical implementations in [8] demonstrates that the LI power can be decreased 74 dB on average. Additionally, it is also shown that FD transmission provides quite better performance under such circumstances with respect to the HD transmission.

3.3 Residual Loop-Interference Channel Models

After several LI cancellation processes, the residual LI channel is still existed between the transmit and receive antenna of a source operating in FD mode. The residual LI channel coefficient is denoted by h_{LI} and it cannot be known by the receiver. So, h_{LI} is also called as the estimation error in [10]. The preliminary works in the FD literature models h_{LI} as Gaussian noise [32, 73], afterwards, it is also modeled as either Rayleigh [19–21] or Nakagami- m fading channels [25, 74], which makes the analyses more accurate for practical purposes. It is worthwhile to note that although h_{LI} is classified as an unknown, the instantaneous and average power of the residual LI channel coefficient (e.g., $|h_{LI}|^2$ and $E\{|h_{LI}|^2\}$) can be obtained by using energy detection techniques as mentioned in [75]. The transmission power of the transmit antenna and the average power of residual LI channel are respectively denoted by P and $E\{|h_{LI}|^2\} = \Omega_{LI}$.

For the power of residual LI component obtained after the cancellation operations, there are generally two different model. In the first model applied in [10, 19, 21, 25, 26], the average power of the residual LI occurred at the receive antenna is supposed to increase linearly with the transmission power P and total amount of LI cancellation is independent from the transmission power. However, the experimental results in [8] reveals that the average amount of LI cancellation enhanced when the transmission power increases. Hence, in the second model utilized in [20, 22–24, 45, 47, 48], total LI power is proportional with the λ^{th} power of P , where $0 \leq \lambda \leq 1$. It includes the optimistic and pessimistic scenarios for FD transmission for $\lambda = 0$ and $\lambda = 1$

cases, respectively. Additionally, the first model is a special case of the second model for $\lambda = 1$. According to the experimental results in [8], it is shown that λ could be decreased to 0.21 after the analog cancellation.



4. ONE-WAY FULL-DUPLEX AF RELAY SYSTEMS

In this chapter, various one-way full-duplex amplify-and-forward (OWFD-AF) relay systems are investigated for variable-gain and fixed-gain cases over Rayleigh and Nakagami- m fading channels. Firstly, the outage probability of the fixed-gain OWFD-AF systems over Rayleigh fading channels is analyzed and it is demonstrated that the fixed-gain relaying provide better outage performance with respect to the variable-gain relaying. Then, the outage probability expressions of variable-gain and fixed-gain OWFD-AF systems are generalized for Nakagami- m fading channels. Finally, five different relay selection policies are utilized for the variable-gain OWFD-AF systems so as to enhance the outage performance.

4.1 OWFD Fixed-Gain AF Relay Systems over Rayleigh Fading Channels

There are several works for the variable-gain OWFD-AF systems, e.g., [19, 21, 26, 75]. However, to the best of our knowledge, there is no outage performance investigation for the fixed-gain OWFD-AF systems in the literature. In this section, outage performance of fixed-gain OWFD-AF systems is analyzed over Rayleigh fading channels. The exact outage probability expression is derived in a single-integral form by using CDF of the end-to-end signal-to-interference-plus-noise ratio (SINR). The theoretical results are validated by Monte Carlo type computer simulations.

4.1.1 System model

We consider a dual-hop transmission scheme as in Figure 4.1, where both the source (S) and the destination (D) have one antenna. The source and destination communicate with each other through a FD relay (R) and there is no direct link between them. The relay is equipped with one receive and one transmit antenna and employs fixed-gain AF method for the signal transmission from the source to the destination. As seen from Figure 4.1, h_{SR} and h_{RD} are the complex channel coefficients between $S - R$ and $R - D$ links, respectively. Similarly, h_R is the complex residual LI channel coefficient.

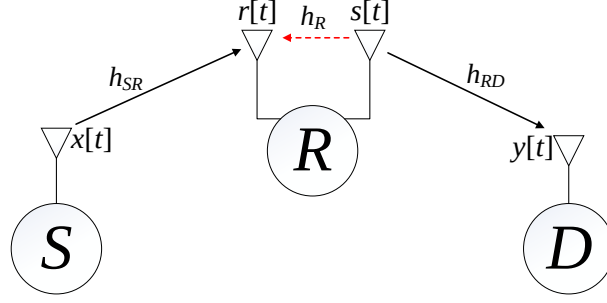


Figure 4.1 : OWFD-AF system model.

Their envelopes (i.e., $|h_{SR}|$, $|h_{RD}|$, $|h_R|$) are assumed to be i.i.d. Rayleigh distributed random variables with average powers of Ω_{SR} , Ω_{RD} and Ω_R , respectively. The PDF of envelopes can be found by using (2.6). When the unit-power signal $x[t]$ is transmitted from the source, the received signal at the relay becomes

$$r[t] = h_{SR}\sqrt{P_S}x[t] + h_{RS}[t] + n_R[t] \quad (4.1)$$

where P_S is the transmission powers of the source, $s[t] = Gr[t-1]$ is the signal transmitted by the relay, G is the amplification factor, $n_R[t]$ is the complex Gaussian noise term distributed according to $\mathcal{CN}(0, \sigma_R^2)$ and t is the time index. By recursively using (4.1), $s[t]$ can be written as follows

$$s[t] = G \sum_{u=0}^{\infty} (Gh_R)^u q[t-u] \quad (4.2)$$

where $q[t] = h_{SR}\sqrt{P_S}x[t-1] + n_R[t-1]$. Finally, the received signal at the destination can be given by

$$\begin{aligned} y[t] &= h_{RD}s[t] + n_D[t] \\ &= \underbrace{Gh_{SR}h_{RD}\sqrt{P_S}x[t-1]}_{\text{Desired Signal}} + \underbrace{Gh_{RD}h_{RS}[t-1]}_{\text{Residual Loop-Interference}} \\ &\quad + \underbrace{Gh_{RD}n_R[t-1] + n_D[t]}_{\text{Noise}} \end{aligned} \quad (4.3)$$

where $n_D[t]$ is the complex Gaussian noise term distributed according to $\mathcal{CN}(0, \sigma_D^2)$.

By analyzing (4.2), one can easily show that stability of the relay is guaranteed and oscillations are prevented, when the $G^2 < 1/|h_R|^2$ condition is satisfied [10]. By taking time average over one symbol duration and assuming the condition is satisfied, the instantaneous transmission power of the relay for a given h_{SR} and h_R and the instantaneous end-to-end SINR at the destination are respectively given as [10]

$$E\{|s[t]|^2 | h_{SR}, h_R\} = P_R \frac{\gamma_{SR} + 1}{\frac{P_R}{G^2 \sigma_R^2} - \gamma_R} \quad (4.4)$$

$$\gamma_{end} = \frac{\gamma_{SR}\gamma_{RD}}{\frac{\gamma_{RD}\gamma_R(\gamma_{SR}+1)}{\frac{P_R}{G^2\sigma_R^2} - \gamma_R} + \gamma_{RD} + \frac{P_R}{G^2\sigma_R^2}} \quad (4.5)$$

where P_R is the average transmission power of the relay, $\gamma_{SR} = P_S|h_{SR}|^2/\sigma_R^2$, $\gamma_{RD} = P_R|h_{RD}|^2/\sigma_D^2$ and $\gamma_R = P_R|h_R|^2/\sigma_R^2$. By defining $\bar{\gamma}_{SR} = P_S\Omega_{SR}/\sigma_R^2$, $\bar{\gamma}_{RD} = P_R\Omega_{RD}/\sigma_D^2$, $\bar{\gamma}_R = P_R\Omega_R/\sigma_R^2$ and using (2.7), the PDF and CDF expressions of exponentially distributed γ_A with $A \in \{SR, RD, R\}$ can be respectively found as $f_{\gamma_A}(x) = \frac{1}{\bar{\gamma}_A} \exp(-x/\bar{\gamma}_A)$ and $F_{\gamma_A}(x) = 1 - \exp(-x/\bar{\gamma}_A)$.

4.1.1.1 Variable-gain AF relaying

In order to assure the instantaneous transmission power as P_R , the amplification factor is chosen as

$$G = \sqrt{\frac{P_R}{P_S|h_{SR}|^2 + P_R|h_R|^2 + \sigma_R^2}}. \quad (4.6)$$

Please note that the same amplification factor is commonly used in the FD literature such as in [19,21,25,27,75]. As mentioned before, although LI channel coefficient is an unknown for the relay, $|h_R|^2$ can be obtained by using the energy detection techniques [75]. Then, the instantaneous end-to-end SINR becomes

$$\gamma_{end} = \frac{X_1\gamma_{RD}}{X_1 + \gamma_{RD} + 1} \quad (4.7)$$

where $X_1 = \gamma_{SR}/(\gamma_R + 1)$. The $G^2 < 1/|h_R|^2$ condition is always satisfied for the variable-gain relaying. By substituting (4.6) into (4.4), the instantaneous transmission power for the variable-gain OWFD-AF relaying for any given h_{SR} and h_R can be easily found as [10]

$$E\{|s[t]|^2 | h_{SR}, h_R\} = P_R. \quad (4.8)$$

4.1.1.2 Fixed-gain AF relaying

In order to adjust the average transmission power to P_R for the fixed-gain OWFD-AF systems, the amplification factor is set to

$$G = \sqrt{\frac{P_R}{P_S\Omega_{SR} + P_R\Omega_R + \sigma_R^2}}. \quad (4.9)$$

Then, the instantaneous end-to-end SINR is obtained as

$$\gamma_{end} = \begin{cases} \frac{\gamma_{SR}\gamma_{RD}(C-\gamma_R)}{\gamma_{SR}\gamma_R\gamma_{RD} + C(C+\gamma_{RD}-\gamma_R)}, & \gamma_R < C \\ 0, & \gamma_R \geq C \end{cases} \quad (4.10)$$

with $C = \bar{\gamma}_{SR} + \bar{\gamma}_R + 1$. Here, $\gamma_R < C$ is obtained from the $G^2 < 1/|h_R|^2$ condition expressed before. It is worthwhile to mention that when the $G^2 < 1/|h_R|^2$ condition is not satisfied, i.e., when $\gamma_R \geq C$, then $\gamma_{end} = 0$ since the power of the interference goes to infinity.

In order to make a fair comparison between the variable-gain and fixed-gain OWFD-AF systems, the average power consumption of fixed-gain OWFD-AF systems should be taken into account. By using (4.4) and (4.9), the average transmission power at the relay is given by

$$E \{|s(t)|^2\} = P_R E \left\{ \frac{\gamma_{SR} + 1}{C - \gamma_R} \right\}. \quad (4.11)$$

By using the independence property of γ_{SR} and γ_R , the average power consumption can be obtained as

$$E \{|s(t)|^2\} = P_R \frac{E \{\gamma_{SR} + 1\}}{E \{C - \gamma_R\}} = P_R \frac{\bar{\gamma}_{SR} + 1}{C - \bar{\gamma}_R} = P_R. \quad (4.12)$$

The above expression proves that although the instantaneous transmission power at the relay varies in the fixed-gain OWFD-AF systems, the average value is set to P_R by means of the selected G . It is important to note that we assume the dynamic range of the power amplifier at the relay is unrestricted for the fixed-gain systems similar to [18, 68–70].

Additionally, Monte Carlo simulation results for 500 different time realizations are plotted in Figure 4.2 so as to ensure the total transmission power equality, where $P_S = 20$, $P_R = 10$, $\sigma_R^2 = \sigma_D^2 = 1$, $\Omega_{SR} = \Omega_{RD} = 1$ and $\Omega_R = 0.01$. As stated above, the instantaneous transmission power of the fixed-gain OWFD-AF systems varies with respect to time, whereas it is constant and equal to $P_R = 10$ for the variable-gain OWFD-AF systems. However, the average transmission power for the fixed-gain relaying is also equal to $P_R = 10$. These results validate fairness of the comparison between variable-gain and fixed-gain OWFD-AF systems.

4.1.2 Outage probability analysis

The exact outage probability of the fixed-gain OWFD-AF system over Rayleigh fading channels is analyzed by using the CDF of γ_{end} given in (4.10). The exact outage probability equals $P_{out}(\gamma_{th}) = F_{\gamma_{end}}(\gamma_{th})$, where $F_{\gamma_{end}}(\cdot)$ is CDF of γ_{end} and γ_{th} is a

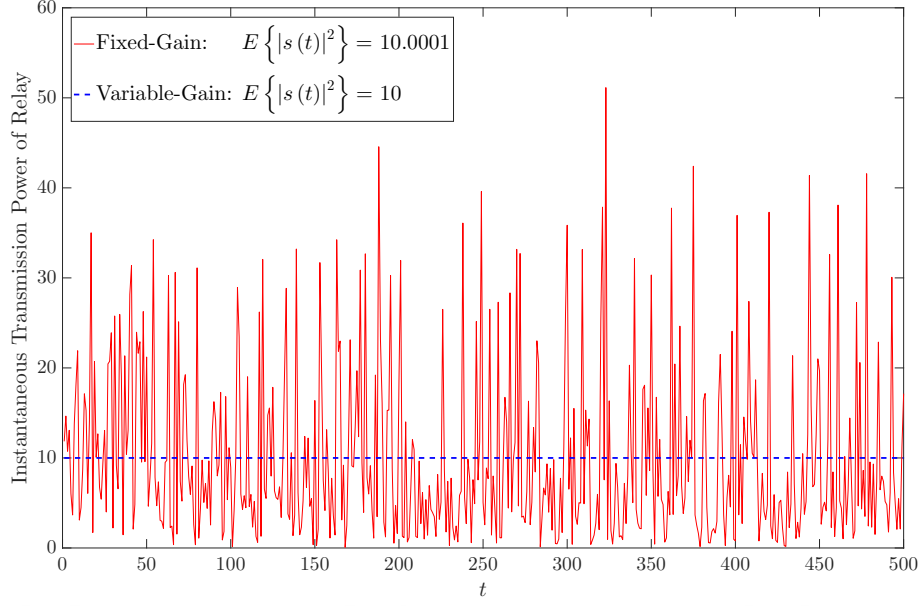


Figure 4.2 : The instantaneous transmission power of relay for the variable-gain and fixed-gain OWFD-AF systems.

threshold SINR. The CDF expression of γ_{end} can be expressed as

$$F_{\gamma_{end}}(x) = P(\gamma_R < C)F_{\gamma_{end}}(x | \gamma_R < C) + P(\gamma_R \geq C)F_{\gamma_{end}}(x | \gamma_R \geq C) \quad (4.13)$$

where $P(\gamma_R < C) = F_{\gamma_R}(C) = 1 - \exp(-C/\bar{\gamma}_R)$ and therefore $P(\gamma_R \geq C) = \exp(-C/\bar{\gamma}_R)$. In (4.13), the conditional CDF equals $F_{\gamma_{end}}(x | \gamma_R \geq C) = P(\gamma_{end} < x | \gamma_R \geq C) = 1$ since $\gamma_{end} = 0$. Hence, we obtain

$$F_{\gamma_{end}}(x) = 1 - [1 - \exp(-C/\bar{\gamma}_R)][1 - F_{\gamma_{end}}(x | \gamma_R < C)]. \quad (4.14)$$

The conditional CDF $F_{\gamma_{end}}(x | \gamma_R < C)$ can be written as

$$\begin{aligned} F_{\gamma_{end}}(x | \gamma_R < C) &= P(\gamma_{end} < x | \gamma_R < C) \\ &= P\left(\frac{\gamma_{SR}\gamma_{RD}(C - \gamma_R)}{\gamma_{SR}\gamma_R\gamma_{RD} + C(C + \gamma_{RD} - \gamma_R)} < x | \gamma_R < C\right) \\ &= \int_{z=0}^C \int_{y=0}^{\infty} P\left[\frac{\gamma_{SR}y(C - z)}{\gamma_{SR}yz + C(C + y - z)} < x\right] f_{\gamma_{RD}}(y) f_{\gamma_R}(z) dy dz. \end{aligned} \quad (4.15)$$

In the above integral, the probability term can be written as

$$\begin{aligned} P\left[\frac{\gamma_{SR}y(C - z)}{\gamma_{SR}yz + C(C + y - z)} < x\right] &= P[\gamma_{SR}y(C - z(x + 1)) < xC(C + y - z)] \\ &= \begin{cases} F_{\gamma_{SR}}\left[\frac{xC(C + y - z)}{y(C - z(x + 1))}\right], & z < \frac{C}{x + 1} \\ 1, & C \geq z \geq \frac{C}{x + 1}. \end{cases} \end{aligned} \quad (4.16)$$

By substituting (4.16) into (4.15), $F_{\gamma_{end}}(x | \gamma_R < C)$ can be given by

$$F_{\gamma_{end}}(x | \gamma_R < C) = \underbrace{\int_{z=0}^C \int_{y=0}^{\infty} f_{\gamma_{RD}}(y) f_{\gamma_R}(z) dy dz}_{I_1} - \underbrace{\int_{z=0}^{\frac{C}{x+1}} \int_{y=0}^{\infty} \bar{F}_{\gamma_{SR}} \left[\frac{xC(C+y-z)}{y(C-z(x+1))} \right] f_{\gamma_{RD}}(y) f_{\gamma_R}(z) dy dz}_{I_2} \quad (4.17)$$

where $\bar{F}_{\gamma_{SR}}(\cdot) = 1 - F_{\gamma_{SR}}(\cdot)$. In the above equation, the first integral term I_1 can be easily solved as

$$I_1 = \int_{z=0}^C f_{\gamma_R}(z) dz = F_{\gamma_R}(C) = 1 - e^{-C/\bar{\gamma}_R}. \quad (4.18)$$

Then, the second integral term I_2 can be rewritten as follows

$$I_2 = \frac{1}{\bar{\gamma}_{RD} \bar{\gamma}_R} \int_{z=0}^{\frac{C}{x+1}} e^{-\frac{xC}{\bar{\gamma}_{SR}(C-z(x+1))} - \frac{z}{\bar{\gamma}_R}} \left[\int_{y=0}^{\infty} e^{-\frac{xC(C-z)}{\bar{\gamma}_{SR}y(C-z(x+1))} - \frac{y}{\bar{\gamma}_{RD}}} dy \right] dz. \quad (4.19)$$

By using [65, Eq. (3.324.1)], I_2 can be derived as

$$I_2 = \frac{2\sqrt{xC}}{\bar{\gamma}_R \sqrt{\bar{\gamma}_{SR} \bar{\gamma}_{RD}}} \int_0^{\frac{C}{x+1}} e^{-\frac{xC}{\bar{\gamma}_{SR}(C-z(x+1))} - \frac{z}{\bar{\gamma}_R}} \times \sqrt{\frac{C-z}{C-z(x+1)}} K_1 \left[2\sqrt{\frac{xC(C-z)}{\bar{\gamma}_{SR} \bar{\gamma}_{RD}(C-z(x+1))}} \right] dz \quad (4.20)$$

where $K_v(\cdot)$ is the v^{th} order modified Bessel function of second kind [65, Eq. (8.407.1)]. Unfortunately, there is no closed-form solution for I_2 , so the conditional CDF $F_{\gamma_{end}}(x | \gamma_R < C) = I_1 - I_2$ can be obtained in a single-integral. Finally, by substituting (4.17), (4.18) and (4.20) into (4.14), the outage probability of fixed-gain OWFD-AF systems is obtained as

$$P_{out}(\gamma_{th}) = 1 - (1 - e^{-C/\bar{\gamma}_R}) \left\{ e^{-C/\bar{\gamma}_R} + \frac{2\sqrt{\gamma_{th}C}}{\bar{\gamma}_R \sqrt{\bar{\gamma}_{SR} \bar{\gamma}_{RD}}} \times \int_0^{\frac{C}{\gamma_{th}+1}} e^{-\frac{\gamma_{th}C}{\bar{\gamma}_{SR}(C-z(\gamma_{th}+1))} - \frac{z}{\bar{\gamma}_R}} \sqrt{\frac{C-z}{C-z(\gamma_{th}+1)}} \times K_1 \left[2\sqrt{\frac{\gamma_{th}C(C-z)}{\bar{\gamma}_{SR} \bar{\gamma}_{RD}(C-z(\gamma_{th}+1))}} \right] dz \right\}. \quad (4.21)$$

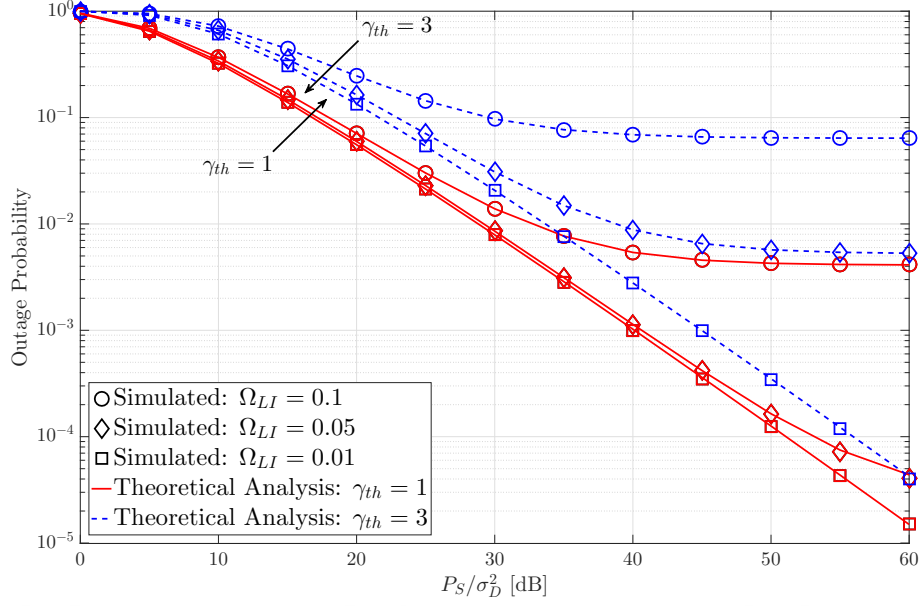


Figure 4.3 : Outage probability of the fixed-gain OWFD-AF systems when $\Omega_R \in \{0.1, 0.05, 0.01\}$, $\Omega_{SR} = \Omega_{RD} = 1$, $\gamma_{th} \in \{1, 3\}$.

The outage probability expression is in a single-integral form. However, it can be easily and efficiently evaluated with common mathematical software packages such as MATLAB, MATHEMATICA or MAPLE.

4.1.3 Numerical results

The theoretical and Monte Carlo type computer simulation results are represented for the outage performance of fixed-gain OWFD-AF system over Rayleigh fading channels in this section. The accuracy of exact outage probability expression is verified via various scenarios. During analyses, all curves are plotted versus P_S/σ_D^2 and we assume that $P_S = P_R$, $\Omega_{SR} = \Omega_{RD} = 1$ and $\sigma_D^2 = \sigma_R^2$.

Figure 4.3 plots the outage probability curves for different LI powers ($\Omega_R \in \{0.1, 0.05, 0.01\}$) and for different thresholds ($\gamma_{th} \in \{1, 3\}$). It is shown that the simulation and theoretical results are perfectly matched with each other for every scenario. As seen from the curves, outage performance improves as long as γ_{th} and also as Ω_R decreases, i.e., efficiency of the LI cancellation process increases. We also observe that the error floor caused by residual LI occurs for the larger values of Ω_R at even moderate values of P_S/σ_D^2 . For example, the outage probability converges to 6×10^{-2} when $\Omega_R = 0.1$ and $\gamma_{th} = 3$ after P_S/σ_D^2 value of 40 dB. However, no error floor is observed for both values of the thresholds when $\Omega_R = 0.01$.

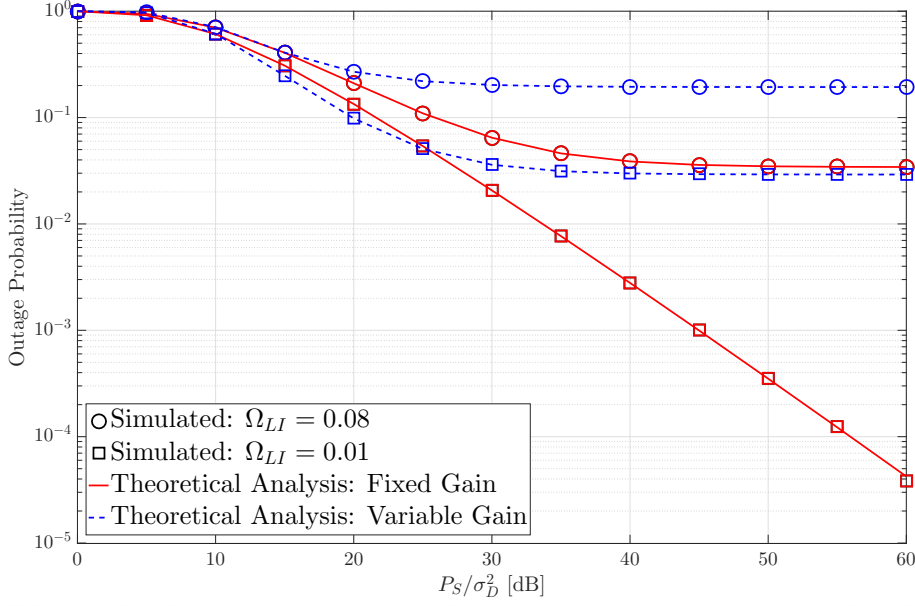


Figure 4.4 : Outage probability of the variable-gain and fixed-gain OWFD-AF systems when $\Omega_R \in \{0.08, 0.01\}$, $\Omega_{SR} = \Omega_{RD} = 1$, $\gamma_{th} = 3$.

Figure 4.4 compares the fixed-gain results obtained in this thesis with the variable-gain results obtained in [19]. Additionally, [19, Eq. (15)] is utilized in order to generate the theoretical curves for the variable-gain case. It is observed that the fixed-gain OWFD-AF systems generally provide quite better outage performance and lower error floors compared to the variable-gain OWFD-AF systems. Moreover, the fixed-gain relaying dramatically outperforms the variable-gain relaying at moderate and high values P_S/σ_D^2 when the effect of LI decreases, e.g., the outage probabilities of the variable-gain and fixed-gain relaying at 60 dB for $\Omega_R = 0.01$ are approximately 4×10^{-5} and 3×10^{-2} , respectively. The performance improvement in the fixed-gain relaying are directly related to the different SINR expressions given in (4.7) and (4.10). On the other hand, the outage performance of the variable-gain relaying is better than the fixed-gain relaying for $\Omega_R = 0.01$, when P_S/σ_D^2 is smaller than 25 dB.

4.2 OWFD Variable-Gain and Fixed-Gain AF Relay Systems over Nakagami- m Fading Channels

After analyzing the outage performance of the fixed-gain OWFD-AF systems over Rayleigh fading channels, the outage probability of both variable-gain and fixed-gain OWFD-AF systems over non-identical Nakagami- m fading channels is also investigated. The system model is demonstrated in Figure 4.1. The exact outage probability expressions are obtained in a single-integral form by again using the CDF

approach. To the best of our knowledge, there is no outage performance investigation in the literature for fixed-gain OWFD-AF systems over non-identical Nakagami- m fading channels. Also, our exact outage performance result for the variable-gain OWFD-AF systems can be easily calculated numerically in comparison to the exact outage probability expression in terms of EGBMGF found in [25], which is based on double Mellin-Barnes type integrals. Furthermore, there is no restriction on the m parameter of the relay-destination link in our analyses. For both of the considered systems, the exact outage performance analyses are supported with the lower-bound and asymptotic outage probability expressions obtained in the closed-forms. The theoretical results are validated by computer simulations. The advantages of fixed-gain relaying with respect to the variable-gain relaying in terms of outage performance are represented. We demonstrate that the system has lower error floor when the fixed-gain relaying is deployed. It is shown that m parameter of the residual LI channel has dominant effect for the performance limitation in the fixed-gain OWFD-AF systems. On the other hand, the dominant m parameter of the variable-gain OWFD-AF systems belongs to the source-relay link channel for the error floor.

4.2.1 Outage probability analyses

In this section, exact, lower-bound and asymptotic outage probability expressions of the considered FD systems are derived by using γ_{end} given in (4.7) and (4.10). Here, the envelopes of h_{SR} , h_{RD} and h_R (i.e., $|h_{SR}|$, $|h_{RD}|$, $|h_R|$) are assumed to be non-identically and independently (n.i.i.d.) Nakagami- m distributed with average powers of Ω_{SR} , Ω_{RD} and Ω_R , respectively. Therefore, the PDF and CDF of γ_A with $A \in \{SR, RD, R\}$ can be respectively given by $f_{\gamma_A}(x) = \frac{\mu_A^{m_A} x^{m_A-1}}{\Gamma(m_A)} e^{-x\mu_A}$ and $F_{\gamma_A}(x) = \frac{\gamma(m_A, x\mu_A)}{\Gamma(m_A)}$, where $\mu_A = m_A/\bar{\gamma}_A$.

4.2.1.1 Variable-gain AF relaying

The outage probability expressions for the variable-gain OWFD-AF systems are derived throughout this part. In order to find the exact outage probability, the CDF expression for the end-to-end SINR given in (4.7) is written as

$$\begin{aligned} F_{\gamma_{end}}(x) &= P(\gamma_{end} < x) = P\left(\frac{X_1 \gamma_{RD}}{X_1 + \gamma_{RD} + 1} < x\right) \\ &= \int_{y=0}^{\infty} P\left(\frac{X_1 y}{X_1 + y + 1} < x\right) f_{\gamma_{RD}}(y) dy. \end{aligned} \quad (4.22)$$

By using [19, Eq. (13)], $F_{\gamma_{end}}(x)$ can be rewritten as

$$\begin{aligned} F_{\gamma_{end}}(x) &= F_{\gamma_{RD}}(x) + \int_{y=x}^{\infty} F_{X_1}\left(\frac{x(y+1)}{y-x}\right) f_{\gamma_{RD}}(y) dy \\ &= 1 - \int_{u=0}^{\infty} \bar{F}_{X_1}\left(\frac{x(u+x+1)}{u}\right) f_{\gamma_{RD}}(u+x) du \end{aligned} \quad (4.23)$$

where $\bar{F}_{X_1}(\cdot) = 1 - F_{X_1}(\cdot)$. Again by using [19, Eq. (14)],

$$\begin{aligned} F_{X_1}(z) &= \int_{v=0}^{\infty} F_{\gamma_{SR}}(z(v+1)) f_{\gamma_R}(v) dv \\ &= \frac{\mu_R^{m_R}}{\Gamma(m_{SR})\Gamma(m_R)} \int_{v=0}^{\infty} \gamma(m_{SR}, \mu_{SR}z(v+1)) v^{m_R-1} e^{-\mu_R v} dv \end{aligned} \quad (4.24)$$

can be obtained. For integer values of m_{SR} , by using [65, Eq. (8.352.6)], [76, Eq. (8)] and also the binomial expansion,

$$\bar{F}_{X_1}(z) = \frac{\mu_R^{m_R} e^{-\mu_{SR}z}}{\Gamma(m_R)} \sum_{k=0}^{m_{SR}-1} \sum_{b=0}^k \frac{\Gamma(b+m_R) (\mu_{SR}z)^k}{b! (k-b)! (\mu_R + \mu_{SR}z)^{b+m_R}} \quad (4.25)$$

can be obtained. Now substituting (4.25) into (4.23), $P_{out}(\gamma_{th})$ for the variable-gain OWFD-AF systems is

$$\begin{aligned} P_{out}(\gamma_{th}) &= 1 - \frac{e^{-(\mu_{SR} + \mu_{RD})\gamma_{th}} \mu_{RD}^{m_{RD}} \mu_R^{m_R}}{\Gamma(m_{RD}) \Gamma(m_R)} \sum_{k=0}^{m_{SR}-1} \sum_{b=0}^k \frac{\Gamma(b+m_R)}{b! (k-b)!} \\ &\quad \times (\mu_{SR}\gamma_{th})^k \int_{u=0}^{\infty} e^{-\mu_{RD}u - \mu_{SR} \frac{(\gamma_{th}^2 + \gamma_{th})}{u}} \left[\frac{u + \gamma_{th} + 1}{u} \right]^k \\ &\quad \times (u + \gamma_{th})^{m_{RD}-1} \left[\mu_R + \mu_{SR} \frac{\gamma_{th}(u + \gamma_{th} + 1)}{u} \right]^{-b-m_R} du. \end{aligned} \quad (4.26)$$

The above expression is in a single-integral form. Hence, it can be easily and efficiently evaluated with common mathematical software packages such as MATLAB, MATHEMATICA or MAPLE. As mentioned before, outage performance of the variable-gain OWFD-AF systems over Nakagami- m fading channels is also independently evaluated in [25]. On the other hand, the exact outage probability expression given in [25, Eq. (16)] is derived in terms of EGBMGF which contains Mellin-Barnes type double-contour-integration [77, Table I]. Although, MATHEMATICA implementation of the EGBMGF function has been provided in [77, Table II], the built-in function is not available for EGBMGF to the best of

our knowledge. Also, different from [25], our result is valid for both integer and non-integer m_{RD} values.

In order to obtain a lower-bound expression for the variable-gain OWFD-AF systems, we use the harmonic mean property of two random variables as $\gamma_{end} < X_1\gamma_{RD}/(X_1 + \gamma_{RD}) < \min(X_1, \gamma_{RD})$, where the instantaneous end-to-end SINR is given in (4.7). So, the lower-bound can be expressed as

$$\begin{aligned} P_{out,low}(\gamma_{th}) &= P(\min\{X_1, \gamma_{RD}\} \leq \gamma_{th}) = 1 - P(\min\{X_1, \gamma_{RD}\} > \gamma_{th}) \\ &= 1 - \bar{F}_{X_1}(\gamma_{th}) \bar{F}_{\gamma_{RD}}(\gamma_{th}). \end{aligned} \quad (4.27)$$

where $\bar{F}_{X_1}(\cdot)$ was already found in (4.25). Therefore, we can easily derive the lower-bound expression in a closed-form as

$$\begin{aligned} P_{out,low}(\gamma_{th}) &= 1 - \frac{\Gamma(m_{RD}, \gamma_{th}\mu_{RD}) \mu_R^{m_R}}{\Gamma(m_{RD}) \Gamma(m_R) e^{\mu_{SR}\gamma_{th}}} \\ &\quad \times \sum_{k=0}^{m_{SR}-1} \sum_{b=0}^k \frac{\Gamma(b + m_R) (\mu_{SR}\gamma_{th})^k}{b! (k-b)! (\mu_R + \mu_{SR}\gamma_{th})^{b+m_R}} \end{aligned} \quad (4.28)$$

where $\Gamma(\cdot, \cdot)$ is the upper incomplete gamma function [65, Eq. (8.350.2)].

When SNR goes to infinity, i.e., when $\{P_S, P_R\} \rightarrow \infty$ for the finite σ_R^2 and σ_D^2 , the instantaneous SINR given in (4.7) becomes

$$\lim_{\substack{P_S \rightarrow \infty \\ P_R \rightarrow \infty}} \gamma_{end} = \lim_{\substack{P_S \rightarrow \infty \\ P_R \rightarrow \infty}} \frac{X_1\gamma_{RD}}{X_1 + \gamma_{RD} + 1} = \frac{\gamma_{SR}}{\gamma_R}. \quad (4.29)$$

In that case, the corresponding asymptotic CDF can be given by $F_{\gamma_{end}}^\infty(x) = P(\gamma_{SR}/\gamma_R < x) = \int_0^\infty F_{\gamma_{SR}}(xy) f_{\gamma_R}(y) dy$. For integer values of m_{SR} by using [65, eq. (8.352.6)], the asymptotic CDF expression is rewritten as

$$\begin{aligned} F_{\gamma_{end}}^\infty(x) &= \int_{y=0}^\infty \left[1 - e^{-xy\mu_{SR}} \sum_{i=0}^{m_{SR}-1} \frac{(xy\mu_{SR})^i}{i!} \right] \frac{\mu_R^{m_R} y^{m_R-1} e^{-y\mu_R}}{\Gamma(m_R)} dy \\ &= 1 - \frac{\mu_R^{m_R}}{\Gamma(m_R)} \sum_{i=0}^{m_{SR}-1} \frac{(x\mu_{SR})^i}{i!} \int_{y=0}^\infty y^{m_R+i-1} e^{-y(\mu_R+x\mu_{SR})} dy. \end{aligned} \quad (4.30)$$

The above integral can be easily solved by using [65, eq. (3.351.3)]. At the end, we obtain the asymptotic outage probability in a closed-form as

$$P_{out}^\infty(\gamma_{th}) = 1 - \frac{1}{\Gamma(m_R)} \left(1 + \frac{\gamma_{th}\mu_{SR}}{\mu_R} \right)^{-m_R} \sum_{i=0}^{m_{SR}-1} \frac{\Gamma(m_R + i)}{i!} \left(1 + \frac{\mu_R}{\gamma_{th}\mu_{SR}} \right)^{-i}. \quad (4.31)$$

As seen, $P_{out}^\infty(\cdot)$ for the variable-gain case is a constant, hence leads to zero diversity order at high SNR. In addition, it does not depend on m_{RD} and Ω_{RD} parameters. Note that similar expression was also obtained independently in [25, Eq. (22)]. Here, we provide (4.31) for the sake of consistency.

4.2.1.2 Fixed-gain AF relaying

The outage probability of the fixed-gain OWFD-AF systems are analyzed in this part. First of all, the CDF expression for the end-to-end SINR given in (4.10) can be expressed in (4.13). According to Nakagami- m fading channel model, $P(\gamma_R < C) = F_{\gamma_R}(C) = \frac{\gamma(m_R, \mu_R C)}{\Gamma(m_R)}$ and $P(\gamma_R \geq C) = 1 - F_{\gamma_R}(C)$. It is worthwhile to remind that $F_{\gamma_{end}}(x | \gamma_R \geq C) = P(\gamma_{end} < x | \gamma_R \geq C) = 1$ since $\gamma_{end} = 0$ when $\gamma_R \geq C$. Therefore, the CDF expression can be written as

$$F_{\gamma_{end}}(x) = 1 - \frac{\gamma(m_R, \mu_R C)}{\Gamma(m_R)} \left[1 - F_{\gamma_{end}}(x | \gamma_R < C) \right]. \quad (4.32)$$

By using (4.17), the above conditional CDF expression can be given by

$$F_{\gamma_{end}}(x | \gamma_R < C) = \underbrace{\int_{z=0}^C \int_{y=0}^{\infty} f_{\gamma_{RD}}(y) f_{\gamma_R}(z) dy dz}_{I_3} - \underbrace{\int_{z=0}^{\frac{C}{x+1}} \int_{y=0}^{\infty} \bar{F}_{\gamma_{SR}} \left[\frac{x C (C + y - z)}{y (C - z (x + 1))} \right] f_{\gamma_{RD}}(y) f_{\gamma_R}(z) dy dz}_{I_4} \quad (4.33)$$

In the above equation, the initial integral term I_3 can be simply solved as

$$I_3 = \int_{z=0}^C \int_{y=0}^{\infty} f_{\gamma_{RD}}(y) f_{\gamma_R}(z) dy dz = F_{\gamma_R}(C) = \frac{\gamma(m_R, \mu_R C)}{\Gamma(m_R)}. \quad (4.34)$$

The other integral term I_4 can be calculated as

$$I_4 = \frac{\mu_{RD}^{m_{RD}} \mu_R^{m_R}}{\Gamma(m_{SR}) \Gamma(m_{RD}) \Gamma(m_R)} \int_{z=0}^{\frac{C}{x+1}} z^{m_R-1} e^{-\mu_R z} \times \int_{y=0}^{\infty} \left[\Gamma(m_{SR}) - \gamma \left(m_{SR}, \frac{x \mu_{SR} C (C + y - z)}{y (C - z (x + 1))} \right) \right] y^{m_{RD}-1} e^{-\mu_{RD} y} dy dz. \quad (4.35)$$

For integer values of m_{SR} , by using [65, eq. (8.352.6)] and applying the binomial expansion, I_4 can be obtained as

$$I_4 = \frac{\mu_{RD}^{m_{RD}} \mu_R^{m_R}}{\Gamma(m_{RD})\Gamma(m_R)} \sum_{k=0}^{m_{SR}-1} \sum_{i=0}^k \frac{(x\mu_{SR}C)^k}{i!(k-i)!} \int_{z=0}^{\frac{C}{x+1}} \frac{(C-z)^i}{(C-z(x+1))^k} \times z^{m_R-1} e^{-\mu_R z - \frac{x\mu_{SR}C}{C-z(x+1)}} \int_{y=0}^{\infty} e^{-\frac{x\mu_{SR}C(C-z)}{y(C-z(x+1))}} y^{m_{RD}-1-i} e^{-\mu_{RD}y} dy dz. \quad (4.36)$$

By using [65, eq. (3.471.9)], the inner integration inside the above expression can be solved. Afterwards, I_4 can be rewritten as

$$I_4 = \frac{2(x\mu_{SR}\mu_{RD}C)^{\frac{m_{RD}}{2}} \mu_R^{m_R}}{\Gamma(m_{RD})\Gamma(m_R)} \sum_{k=0}^{m_{SR}-1} \sum_{i=0}^k \frac{(x\mu_{SR}C)^{\frac{2k-i}{2}}}{i!(k-i)! \mu_{RD}^{\frac{-i}{2}}} \times \int_{z=0}^{\frac{C}{x+1}} \frac{(C-z)^{\frac{m_{RD}+i}{2}} z^{m_R-1}}{(C-z(x+1))^{\frac{m_{RD}-i+2k}{2}}} e^{-\mu_R z - \frac{x\mu_{SR}C}{C-z(x+1)}} \times K_{m_{RD}-i} \left[2\sqrt{\frac{x\mu_{SR}\mu_{RD}C(C-z)}{C-z(x+1)}} \right] dz. \quad (4.37)$$

To the best of our knowledge, there is no closed-form solution for the above expression. By substituting (4.33), (4.34) and (4.37) into (4.32), $P_{out}(\cdot)$ for the fixed-gain OWFD-AF systems can be evaluated via numerical integration and it is written as

$$P_{out}(\gamma_{th}) = 1 - \frac{\gamma(m_R, \mu_R C)}{\Gamma(m_R)} \left\{ 1 - \frac{\gamma(m_R, \mu_R C)}{\Gamma(m_R)} + \frac{2(\gamma_{th}\mu_{SR}\mu_{RD}C)^{\frac{m_{RD}}{2}}}{\mu_R^{-m_R}\Gamma(m_{RD})\Gamma(m_R)} \times \sum_{k=0}^{m_{SR}-1} \sum_{i=0}^k \frac{(\gamma_{th}\mu_{SR}C)^{\frac{2k-i}{2}}}{i!(k-i)! \mu_{RD}^{\frac{-i}{2}}} \int_{z=0}^{\frac{C}{\gamma_{th}+1}} \frac{(C-z)^{\frac{m_{RD}+i}{2}} z^{m_R-1}}{(C-z(\gamma_{th}+1))^{\frac{m_{RD}-i+2k}{2}}} \times e^{-\mu_R z - \frac{\gamma_{th}\mu_{SR}C}{C-z(\gamma_{th}+1)}} K_{m_{RD}-i} \left[2\sqrt{\frac{\gamma_{th}\mu_{SR}\mu_{RD}C(C-z)}{C-z(\gamma_{th}+1)}} \right] dz \right\}. \quad (4.38)$$

The above expression is in a single-integral form as well. Therefore, it can be easily and efficiently evaluated with common mathematical software packages.

One can show that the single-integral term inside the exact outage probability of the fixed-gain OWFD-AF systems given in (4.38) can be lower-bounded when $C - z(\gamma_{th} + 1)$ term within the integration is replaced with $C - z$ term. Then, the

lower-bound outage probability expression can be given by

$$\begin{aligned}
P_{out,low}(\gamma_{th}) = & 1 - \frac{\gamma(m_R, \mu_R C)}{\Gamma(m_R)} \left[1 - \frac{\gamma(m_R, \mu_R C)}{\Gamma(m_R)} + \frac{2(\gamma_{th} \mu_{SR} \mu_{RD} C)^{\frac{m_{RD}}{2}}}{\mu_R^{-m_R} \Gamma(m_{RD}) \Gamma(m_R)} \right. \\
& \times \sum_{k=0}^{m_{SR}-1} \sum_{i=0}^k \frac{(\gamma_{th} \mu_{SR} C)^{\frac{2k-i}{2}}}{i!(k-i)! \mu_{RD}^{\frac{i}{2}}} K_{m_{RD}-i} \left(2\sqrt{\gamma_{th} \mu_{SR} \mu_{RD} C} \right) \\
& \times \underbrace{\int_{z=0}^{\frac{C}{\gamma_{th}+1}} (C-z)^{i-k} z^{m_R-1} e^{-\mu_R z - \frac{\gamma_{th} \mu_{SR} C}{C-z}} dz}_{I_5} \left. \right]. \quad (4.39)
\end{aligned}$$

By applying the variable transformation as $y = \frac{C-z}{C}$ and using the binomial expansion, the above integral I_5 is expressed as

$$I_5 = \sum_{b=0}^{m_R-1} \binom{m_R-1}{b} \frac{C^{m_R+i-k}}{e^{\mu_R C} (-1)^{-b}} \int_{y=\frac{\gamma_{th}}{\gamma_{th}+1}}^1 \frac{y^{i+b-k}}{e^{-\mu_R C y + \frac{\gamma_{th} \mu_{SR}}{y}}} dy. \quad (4.40)$$

Then, the power series and the binomial expansions are used respectively as follows

$$\begin{aligned}
I_5 = & \frac{C^{m_R+i-k}}{e^{\mu_R C}} \sum_{b=0}^{m_R-1} \sum_{n=0}^{\infty} \binom{m_R-1}{b} \frac{(-1)^b}{n!} \int_{y=\frac{\gamma_{th}}{\gamma_{th}+1}}^1 y^{i+b-k} \left(\frac{\gamma_{th} \mu_{SR}}{y} - \mu_R C y \right)^n dy \\
= & \frac{C^{m_R+i-k}}{e^{\mu_R C}} \sum_{b=0}^{m_R-1} \sum_{n=0}^{\infty} \sum_{u=0}^n \binom{m_R-1}{b} \frac{(\gamma_{th} \mu_{SR})^u (-1)^{b+u}}{(n-u)! u! (\mu_R C)^{u-n}} \underbrace{\int_{y=\frac{\gamma_{th}}{\gamma_{th}+1}}^1 y^{i+b+n-k-2u} dy}_{I_6} \quad (4.41)
\end{aligned}$$

where I_6 can be easily solved as

$$I_6 = \begin{cases} \ln \left(\frac{\gamma_{th}+1}{\gamma_{th}} \right), & \text{if } n+b+i-k-2u = -1 \\ \frac{1}{\gamma_{th}+1}, & \text{if } n+b+i-k-2u = 0 \\ \frac{1 - [\gamma_{th}/(\gamma_{th}+1)]^{n+b+i-k-2u+1}}{n+b+i-k-2u+1}, & \text{otherwise.} \end{cases} \quad (4.42)$$

By combining (4.39), (4.41) and (4.42), we obtain the lower-bound expression for the fixed-gain OWFD-AF systems in a closed-form as

$$\begin{aligned}
P_{out,low}(\gamma_{th}) = & 1 - \frac{\gamma(m_R, \mu_R C)}{\Gamma(m_R)} \left[1 - \frac{\gamma(m_R, \mu_R C)}{\Gamma(m_R)} + \frac{2e^{-\mu_R C}}{\Gamma(m_{RD}) \Gamma(m_R)} \right. \\
& \times \sum_{k=0}^{m_{SR}-1} \sum_{i=0}^k \sum_{b=0}^{m_R-1} \sum_{n=0}^{\infty} \sum_{u=0}^n \binom{m_R-1}{b} \frac{(\gamma_{th} \mu_{SR})^{\frac{m_{RD}-i}{2}+k+u}}{i!(k-i)!(n-u)! u!} \\
& \times \frac{\mu_{RD}^{\frac{m_{RD}+i}{2}} C^{m_R+\frac{m_{RD}+i}{2}+n-u}}{(-1)^{-b-u} \mu_R^{u-m_R-n}} K_{m_{RD}-i} \left(2\sqrt{\gamma_{th} \mu_{SR} \mu_{RD} C} \right) I_6 \left. \right]. \quad (4.43)
\end{aligned}$$

When $\{P_S, P_R\} \rightarrow \infty$, the instantaneous SINR given in (4.10) becomes

$$\begin{aligned} \lim_{\substack{P_S \rightarrow \infty \\ P_R \rightarrow \infty}} \gamma_{end} &= \lim_{\substack{P_S \rightarrow \infty \\ P_R \rightarrow \infty}} \frac{\gamma_{SR}\gamma_{RD}(C - \gamma_R)}{\gamma_{SR}\gamma_R\gamma_{RD} + C(C + \gamma_{RD} - \gamma_R)} \\ &= \lim_{\substack{P_S \rightarrow \infty \\ P_R \rightarrow \infty}} \frac{C - \gamma_R}{\gamma_R} = \frac{\bar{\gamma}_{SR} + \bar{\gamma}_R}{\gamma_R} - 1. \end{aligned} \quad (4.44)$$

The asymptotic expression for the outage probability of the fixed-gain OWFD-AF systems can be obtained from (4.32). One can easily show that $\gamma(m_R, \mu_R C)$ goes to $\gamma(m_R, m_R + m_{SR}/\zeta)$ for $\{P_S, P_R\} \rightarrow \infty$ where $\zeta = \mu_{SR}/\mu_R$. Then, the conditional CDF term in (4.32) is obtained for the asymptotic SINR as

$$\begin{aligned} F_{\gamma_{end}}^\infty(x | \gamma_R < C) &= P\left(\frac{\bar{\gamma}_{SR} + \bar{\gamma}_R}{\gamma_R} - 1 < x \mid \gamma_R < C\right) \\ &= \frac{P\left(\frac{\bar{\gamma}_{SR} + \bar{\gamma}_R}{x+1} < \gamma_R < C\right)}{P(\gamma_R < C)} \\ &= 1 - \frac{F_{\gamma_R}\left(\frac{\bar{\gamma}_{SR} + \bar{\gamma}_R}{x+1}\right)}{F_{\gamma_R}(C)} \\ &= 1 - \frac{\gamma[m_R, (m_R + m_{SR}/\zeta) / (x+1)]}{\gamma(m_R, m_R + m_{SR}/\zeta)}. \end{aligned} \quad (4.45)$$

By substituting (4.45) in (4.32), the asymptotic outage probability is obtained in a closed-form as

$$P_{out}^\infty(\gamma_{th}) = 1 - \frac{\gamma[m_R, (m_R + m_{SR}/\zeta) / (x+1)]}{\Gamma(m_R)}. \quad (4.46)$$

As seen, $P_{out}^\infty(\cdot)$ for the fixed-gain case is also a constant and therefore leads to zero diversity order at high SNR region. Moreover, it does not depend on m_{SR} , m_{RD} and Ω_{RD} parameters.

4.2.2 Numerical results

The theoretical and Monte Carlo simulation results for the outage performance of the variable-gain and fixed-gain OWFD-AF systems are shown in this section. During the outage performance analyses, we assume that $\sigma_D^2 = \sigma_R^2$ for all the scenarios and $\Omega_{SR} = \Omega_{RD} = 1$, $P_S = P_R$ unless otherwise stated. Moreover, the residual LI power, apart from Figure 4.9, is chosen in the range of 0.01 and 0.1 (−20 dB and −10 dB) similar to [19, 21, 25, 27, 75].

The outage probability results of the variable-gain OWFD-AF systems for $\Omega_R \in \{0.01, 0.05\}$, $\gamma_{th} = 1$ and various m values are demonstrated by Figure 4.5. As

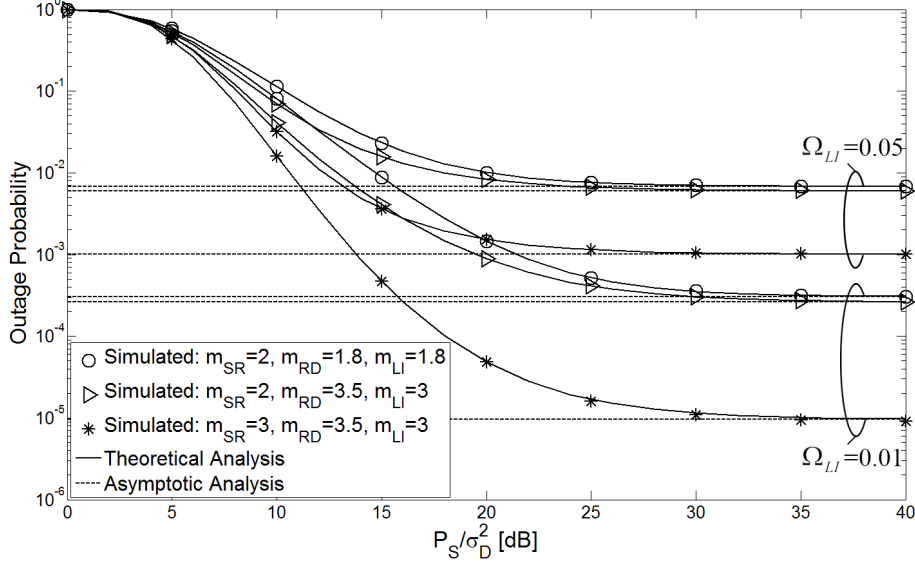


Figure 4.5 : Outage probability of the variable-gain OWFD-AF systems when $\Omega_R \in \{0.01, 0.05\}$, $\gamma_{th} = 1$ and various m values.

seen from the curves, the outage probability decreases as Ω_R decreases, i.e., efficiency of the LI cancellation process improves. The simulation results and the error floors at the high SNR region are perfectly matched with both the analytical exact and the asymptotic outage probability results obtained from (4.26) and (4.31), respectively. It is observed that the parameter m_{SR} has more dominant effect than the other m parameters on the asymptotic performance in the case of variable-gain relaying.

Figure 4.6 plots the outage probability results of the fixed-gain OWFD-AF systems versus P_S/σ_D^2 for two different LI powers, $\Omega_R \in \{0.05, 0.1\}$, and various m values. The threshold SINR was set to $\gamma_{th} = 3$ for these scenarios. As seen, the exact analytical results generated from (4.38) perfectly overlap with the simulation results for all P_S/σ_D^2 values, verifying the accuracy of our analyses. Moreover, the error floors observed in simulations are consistent with the asymptotic expression for the fixed-gain OWFD-AF relaying given in (4.46). Different from the variable-gain OWFD-AF systems, we see that the dominant m parameter on the asymptotic performance is m_R and other m parameters have no effect on the error floor for the fixed-gain OWFD-AF systems.

Figure 4.7 compares the outage performance for the variable-gain and fixed-gain OWFD-AF systems for $\Omega_R = 0.06$, $\gamma_{th} = 2$ and various m values. As seen, the fixed-gain relaying provides better outage performance than the variable-gain relaying at moderate and high SINR regions since the LI effect decreases due to the fixed

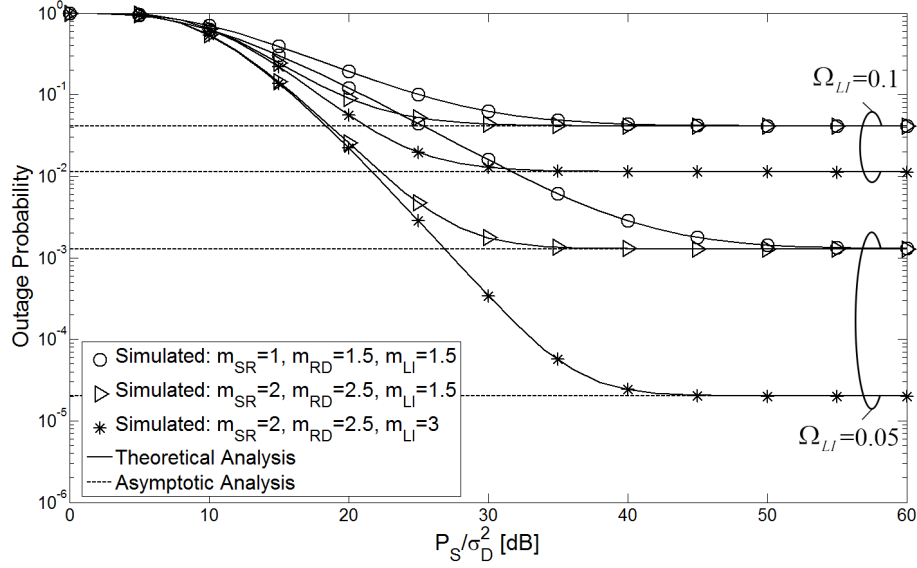


Figure 4.6 : Outage probability of the fixed-gain OWFD-AF systems when $\Omega_R \in \{0.05, 0.1\}$, $\gamma_{th} = 3$ and various m values.

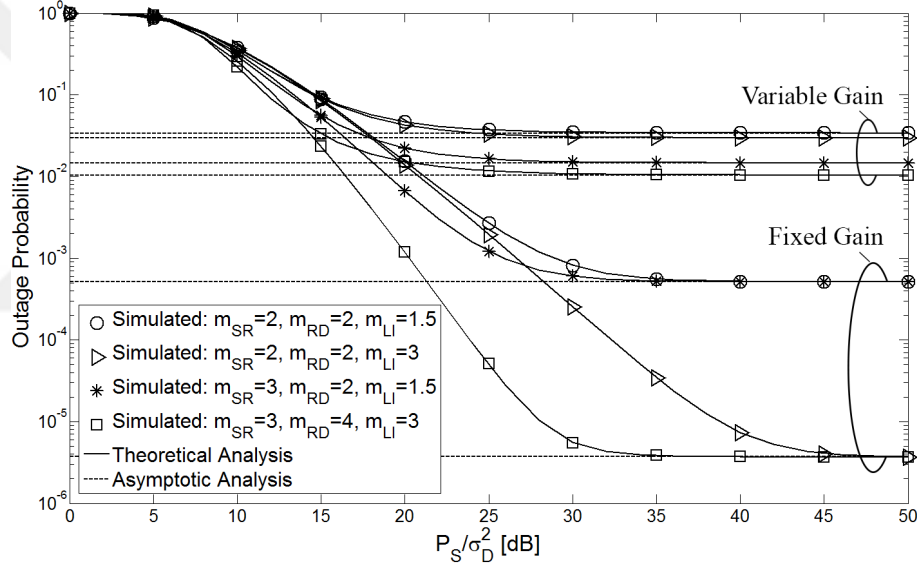


Figure 4.7 : Outage probability of the variable-gain and fixed-gain OWFD-AF systems when $\Omega_R = 0.06$, $\gamma_{th} = 2$ and various m values.

amplification factor. We also present the asymptotic outage results and observe that the error floor occurs for both systems even at moderate values of P_S/σ_D^2 due to the residual LI at the relay. The variable-gain OWFD-AF systems have much higher error floor and reach to the error floor at lower P_S/σ_D^2 values compared to the fixed-gain OWFD-AF systems.

In Figure 4.8, the consistency of the lower-bound analyses for both the variable-gain and fixed-gain OWFD-AF systems are demonstrated where the system parameters are chosen as $2P_S = 3P_R$, $\Omega_{SR} = 1$, $\Omega_{RD} = 2$, $\Omega_R = 0.025$, $m_{SR} = 4$, $m_{RD} = 2.5$, $m_R = 1$ and $\gamma_{th} = 4$. It is observed that the exact outage probability of the fixed-gain

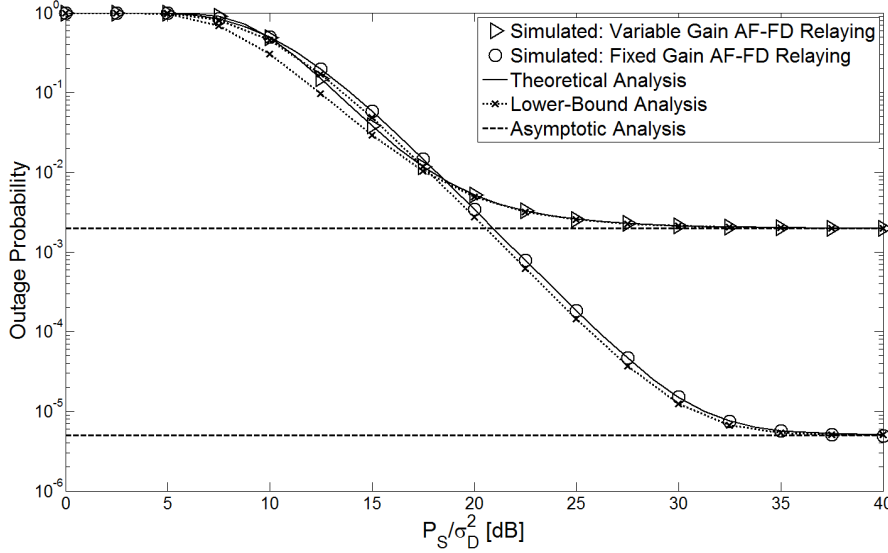


Figure 4.8 : Outage probability of the variable-gain and fixed-gain OWFD-AF systems when $2P_S = 3P_R$, $\Omega_{SR} = 1$, $\Omega_{RD} = 2$, $\Omega_R = 0.025$, $m_{SR} = 4$, $m_{RD} = 2.5$, $m_R = 1$ and $\gamma_{th} = 4$.

OWFD-AF systems is tightly lower-bounded by the expression given in (4.43). For instance, there is only 0.38 dB difference between the lower-bound and the exact outage probability curves when the outage probability equals to 10^{-4} . In addition, the lower-bound overlaps with the exact outage probability at the error floor. For the variable-gain OWFD-AF systems, the lower-bound results obtained from (4.28) are close to the exact outage performance, e.g., the gap between the lower-bound and exact outage probability curves is approximately 1.4 dB at 2×10^{-2} . As in the fixed-gain OWFD-AF systems, the lower-bound curves reach to the exact performance at the error floor when variable-gain relays are employed.

Figure 4.9 indicates the outage probability versus Ω_R for P_S/σ_D^2 values of 10 dB, 20 dB and 30 dB with $m_{SR} = 3$, $m_{RD} = m_R = 2.5$ and $\gamma_{th} = 2$. It is seen that the fixed-gain relaying achieves better outage performance as long as Ω_R and P_S/σ_D^2 increases. On the other hand, the variable gain relaying has lower outage probability for the small values of Ω_R , e.g. when Ω_R is smaller than approximately -2 dB, -16 dB and -24 dB for P_S/σ_D^2 equals to 10 dB, 20 dB and 30 dB, respectively. Moreover, it is observed that the outage probability values for both the variable-gain and fixed-gain OWFD-AF systems are higher than 5×10^{-1} , even for the high SNR values, e.g., $P_S/\sigma_D^2 = 30$ dB, when Ω_R is set to $\Omega_{SR} = \Omega_{RD} = 1$. This example clearly presents the importance of the LI cancellation in the FD transmission.

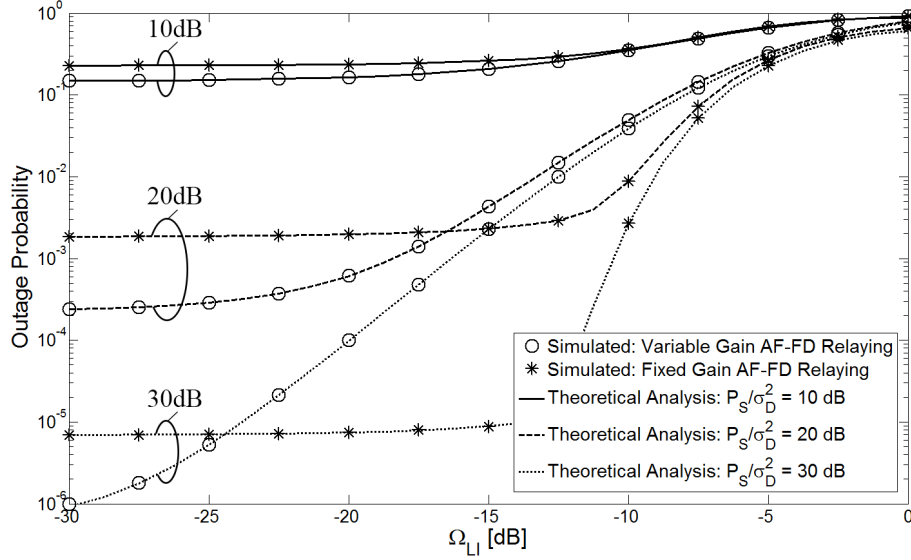


Figure 4.9 : Outage probability of the variable-gain and fixed-gain OWFD-AF systems when $m_{SR} = 3$, $m_{RD} = m_R = 2.5$ and $\gamma_{th} = 2$.

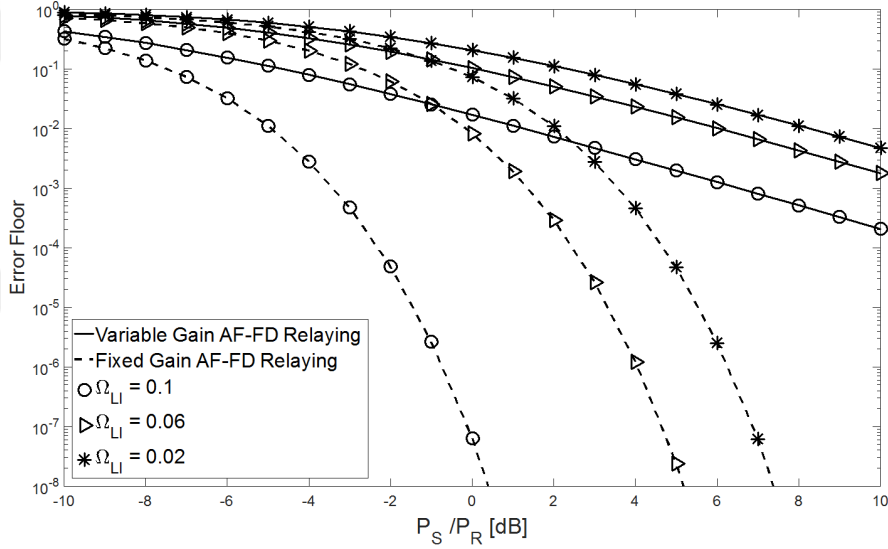


Figure 4.10 : Error floor of the variable-gain and fixed-gain OWFD-AF systems when $\Omega_{SR} = 1.2$, $\Omega_{RD} = 0.8$, $\Omega_R \in \{0.02, 0.06, 0.1\}$, $m_{SR} = 2$, $m_{RD} = 1.4$, $m_R = 1.9$ and $\gamma_{th} = 5$.

In Figure 4.10, by using the consistency of asymptotic outage probability expressions given in (4.31) and (4.46) with the error floors for the variable-gain and fixed-gain OWFD-AF systems, respectively, the error floors are represented versus P_S/P_R for $\Omega_{SR} = 1.2$, $\Omega_{RD} = 0.8$, $\Omega_R \in \{0.02, 0.06, 0.1\}$, $m_{SR} = 2$, $m_{RD} = 1.4$, $m_R = 1.9$ and $\gamma_{th} = 5$. We observe that the error floors decrease as long as the P_S/P_R ratio increases for both systems. Actually, this is a predictable outcome because both the end-to-end SINR expressions given in (4.44) and (4.29) for $P_S, P_R \rightarrow \infty$ are directly proportional to P_S and inversely proportional to P_R . Similar to the previous observations, it is

also shown that the fixed-gain relaying has considerably lower error floors. Also, the error floor for the fixed-gain relays decreases more sharply in comparison with the variable-gain relays as long as P_S/P_R increases.

4.3 Relay Selection for OWFD Variable-Gain AF Relay Systems over Nakagami- m Fading Channels

In the literature, the outage probability of variable-gain OWFD-AF systems adopting several relay selection algorithms over Rayleigh fading channels is analyzed in [19]. Here, we first investigate the outage performance of variable-gain OWFD-AF systems over Nakagami- m fading channels, and then apply five different relay selection policies in order to improve the performance of proposed system, which is called as one-way full-duplex amplify-and-forward relay selection (OWFD-AF-RS) system. By using the CDF approach, the exact outage probability expressions are obtained in a single-integral form for each relay selection policies. The theoretical analyses are verified with several Monte Carlo simulation scenarios. The numerical results reveal that the best relay selection policy so as to increase the performance is the optimal relay selection policy which requires full CSI. Additionally, lower-bound and asymptotic outage probability expressions are obtained in the closed-forms for each relay selection policies. The error floors due to the residual LI can be determined easily by using the derived asymptotic expressions.

4.3.1 System model

We again consider a dual-hop transmission scheme for the variable-gain OWFD-AF systems as illustrated in 4.11, where there are one source node (S), one destination node (D) and N relay nodes ($R_c, 1 \leq c \leq N$). For one-way transmission, each relay node has one transmit and one receive antenna, while the source is equipped with a single transmit antenna and the destination is equipped with a single receive antenna. It is assumed that there is no direct link between source and destination. In the proposed system, the transmission is accomplished via the selected relay among N relays according to one of the selection policies explained in the following section. Therefore, only one relay node is active in order to convey information from source to destination, while $N - 1$ relays remain silent. As seen from Figure 4.11, the complex channel

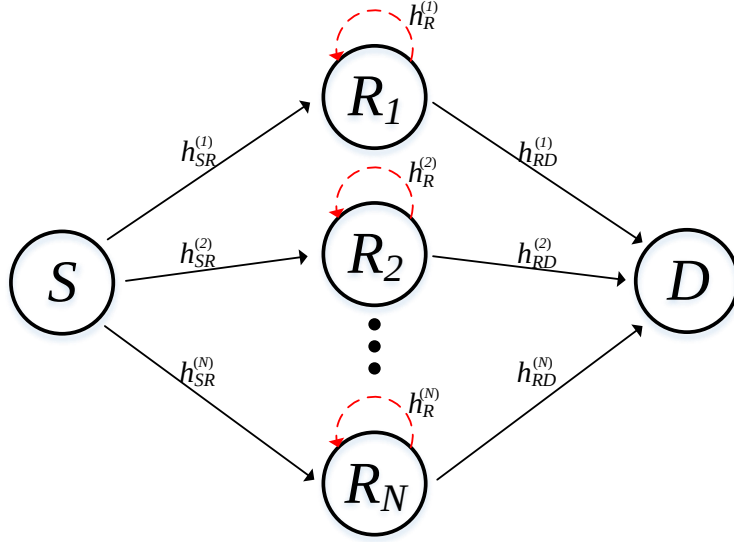


Figure 4.11 : OWFD-AF-RS system model.

coefficients between $S - R_c$ and $R_c - D$ are respectively denoted as $h_{SR}^{(c)}$ and $h_{RD}^{(c)}$. Moreover, the residual LI channel coefficients at R_c is denoted as $h_R^{(c)}$. The envelope of all channel coefficients, including the residual LI channels, (i.e., $|h_{SR}^{(c)}|, |h_{RD}^{(c)}|, |h_R^{(c)}|$) are assumed to be n.i.i.d. Nakagami- m distributed. However, $|h_{SR}^{(c)}|, |h_{RD}^{(c)}|$ and $|h_R^{(c)}|$ are modeled as i.i.d. among themselves with average powers of Ω_{SR} , Ω_{RD} and Ω_R , respectively. Note that the complex Gaussian noise terms occurred at each relay are assumed to be i.i.d. according to $\mathcal{CN}(0, \sigma_R^2)$. Similar to (4.7), when R_c is utilized for the transmission, the end-to-end SINR of variable-gain OWFD-AF-RS system becomes

$$\gamma_{end}^{(c)} = \frac{X_1^{(c)} \gamma_{RD}^{(c)}}{X_1^{(c)} + \gamma_{RD}^{(c)} + 1} \quad (4.47)$$

where $X_1^{(c)} = \gamma_{SR}^{(c)} / (\gamma_R^{(c)} + 1)$, $\gamma_{SR}^{(c)} = P_S |h_{SR}^{(c)}|^2 / \sigma_R^2$, $\gamma_{RD}^{(c)} = P_R |h_{RD}^{(c)}|^2 / \sigma_D^2$ and $\gamma_R^{(c)} = P_R |h_R^{(c)}|^2 / \sigma_R^2$. By means of identically distribution in themselves, the PDF and CDF expressions of $\gamma_A^{(c)}$ with $A \in \{SR, RD, R\}$ can be respectively written as $f_{\gamma_A^{(c)}}(x) = \frac{\mu_A^{m_A} x^{m_A-1}}{\Gamma(m_A)} e^{-x\mu_A}$ and $F_{\gamma_A^{(c)}}(x) = \frac{\gamma(m_A, x\mu_A)}{\Gamma(m_A)}$, respectively.

4.3.2 Relay selection policies

In this section, we propose five different relay selection policies, which can be employed in the variable-gain OWFD-AF-RS systems having N relays. In each algorithm, the relay selection can be applied in accordance with either one or more of $\gamma_{SR}^{(c)}$, $\gamma_{RD}^{(c)}$ and $\gamma_R^{(c)}$.

4.3.2.1 Optimal selection

The first selection policy is optimal selection (OS), where the basic idea is to choose the relay providing the highest SINR. In the OS policy, the index of selected relay is

$$b_{OS} = \arg \max_{1 \leq c \leq N} \left\{ \gamma_{end}^{(c)} \right\} \quad (4.48)$$

where full CSI is required for the selection. Therefore, OS has the highest complexity with respect to the other following policies.

4.3.2.2 Partial selection with S-R link and residual LI

By using the relation between $\gamma_{SR}^{(c)}$ and $\gamma_R^{(c)}$ in (4.47), the second relay selection policy called as SRLI can be performed as follows

$$b_{SRLI} = \arg \max_{1 \leq c \leq N} \left\{ \frac{\gamma_{SR}^{(c)}}{\gamma_R^{(c)} + 1} \right\}. \quad (4.49)$$

Different from OS, $\gamma_{RD}^{(c)}$ is not required so as to select the relay when SRLI is employed.

4.3.2.3 Partial selection with S-R link

For the following relay selection policy, it is assumed that the selection is applied according to the first link based on only the values of $\gamma_{SR}^{(c)}$. For the partial selection with $S - R$ link (SR) policy, the index of selected relay is given by

$$b_{SR} = \arg \max_{1 \leq c \leq N} \left\{ \gamma_{SR}^{(c)} \right\}. \quad (4.50)$$

4.3.2.4 Partial selection with residual LI

When only the residual LI components are taken into account for the relay selection, the relay node having the minimum residual LI has to be chosen because the performance of OWFD-AF-RS systems deteriorates with the increase of residual LI component. Hence, the partial selection policy called as RLI is written as follows

$$b_{RLI} = \arg \min_{1 \leq c \leq N} \left\{ \gamma_R^{(c)} \right\}. \quad (4.51)$$

4.3.2.5 Partial selection with R-D link

In the last relay selection policy, the partial selection is performed according to the second link by using the values of $\gamma_{RD}^{(c)}$. Therefore, the index of the chosen relay node for the partial selection with $R - D$ link (RD) policy is written as

$$b_{RD} = \arg \max_{1 \leq c \leq N} \left\{ \gamma_{RD}^{(c)} \right\}. \quad (4.52)$$

4.3.3 Outage probability analyses

The exact, lower-bound and asymptotic outage probability expressions of the OWFD-AF-RS system for each relay selection policy are derived throughout this section. By using the end-to-end SINR expression given in (4.47), the exact outage probability of the variable-gain OWFD-AF-RS systems is given by

$$P_*(\gamma_{th}) = P\left(\frac{X_1^{(b_*)}\gamma_{RD}^{(b_*)}}{X_1^{(b_*)} + \gamma_{RD}^{(b_*)} + 1} < \gamma_{th}\right) \quad (4.53)$$

where $* \in \{OS, SRLI, SR, RLI, RD\}$ represents the utilized relay selection policy for the transmission. Furthermore, an upper-bound expression for the end-to-end SINR can be defined as $\gamma_{end}^{(c)} \leq \min\{X_1^{(b_*)}, \gamma_{RD}^{(b_*)}\}$ as in Section 4.2.1.1. The upper-bound for the end-to-end SINR implies a lower-bound for the outage probability, which can be written as follows

$$\begin{aligned} P_{*,low}(\gamma_{th}) &= P\left(\min\{X_1^{(b_*)}, \gamma_{RD}^{(b_*)}\} \leq \gamma_{th}\right) \\ &= 1 - P\left(X_1^{(b_*)} \geq \gamma_{th}\right) P\left(\gamma_{RD}^{(b_*)} \geq \gamma_{th}\right) \\ &= 1 - \bar{F}_{X_1^{(b_*)}}(\gamma_{th}) \bar{F}_{\gamma_{RD}^{(b_*)}}(\gamma_{th}) \end{aligned} \quad (4.54)$$

where $\bar{F}_{X_1^{(b_*)}}(\cdot) = 1 - F_{X_1^{(b_*)}}(\cdot)$ and $\bar{F}_{\gamma_{RD}^{(b_*)}}(\cdot) = 1 - F_{\gamma_{RD}^{(b_*)}}(\cdot)$. Another significant performance criteria is the asymptotic outage probability, which presents the performance of OWFD-AF-FD systems in the high SNR region, i.e., when $\{P_S, P_R\} \rightarrow \infty$ for the finite σ_R^2 and σ_D^2 . Then, the asymptotic outage probability in terms of the lower-bound expression can be obtained as

$$\begin{aligned} P_*^\infty(\gamma_{th}) &= \lim_{\substack{P_S \rightarrow \infty \\ P_R \rightarrow \infty}} P\left(\frac{X_1^{(b_*)}\gamma_{RD}^{(b_*)}}{X_1^{(b_*)} + \gamma_{RD}^{(b_*)} + 1} < \gamma_{th}\right) \\ &\approx \lim_{\substack{P_S \rightarrow \infty \\ P_R \rightarrow \infty}} P\left(\min\{X_1^{(b_*)}, \gamma_{RD}^{(b_*)}\} \leq \gamma_{th}\right) \\ &= \lim_{\substack{P_S \rightarrow \infty \\ P_R \rightarrow \infty}} P_{*,low}(\gamma_{th}). \end{aligned} \quad (4.55)$$

4.3.3.1 Optimal selection

In the first relay selection policy as OS, by using (4.53) and [19, Eq. (12)], the exact outage probability is given by

$$P_{OS}(\gamma_{th}) = P\left(\frac{X_1^{(b_{OS})}\gamma_{RD}^{(b_{OS})}}{X_1^{(b_{OS})} + \gamma_{RD}^{(b_{OS})} + 1} \leq \gamma_{th}\right) = \left[F_{\gamma_{end}^{(c)}}(\gamma_{th})\right]^N \quad (4.56)$$

where $F_{\gamma_{end}^{(c)}}(\cdot)$ is the CDF of $\gamma_{end}^{(c)}$. Then, the CDF of end-to-end SINR given in (4.47) can be written as

$$\begin{aligned} F_{\gamma_{end}^{(c)}}(x) &= P\left(\frac{X_1^{(c)}\gamma_{RD}^{(c)}}{X_1^{(c)} + \gamma_{RD}^{(c)} + 1} \leq x\right) \\ &= 1 - \int_{y=x}^{\infty} f_{\gamma_{RD}^{(c)}}(y) \bar{F}_{X_1^{(c)}}\left[\frac{x(y+1)}{y-x}\right] dy. \end{aligned} \quad (4.57)$$

Then, the CDF of $X_1^{(c)}$ can be given by

$$\begin{aligned} F_{X_1^{(c)}}(x) &= P\left(\frac{\gamma_{SR}^{(c)}}{\gamma_R^{(c)} + 1} \leq x\right) = \int_{y=0}^{\infty} f_{\gamma_R^{(c)}}(y) F_{\gamma_{SR}^{(c)}}[x(y+1)] dy \\ &= \frac{\mu_R^{m_R}}{\Gamma(m_R)\Gamma(m_{SR})} \int_{y=0}^{\infty} \frac{\gamma[m_{SR}, x(y+1)\mu_{SR}]}{y^{1-m_R}e^{y\mu_R}} dy. \end{aligned} \quad (4.58)$$

By using [65, Eq. (8.352.6)], the above expression for the integer values of m_{SR} can be rewritten as follows

$$F_{X_1^{(c)}}(x) = 1 - \frac{\mu_R^{m_R}}{\Gamma(m_R)} \sum_{q=0}^{m_{SR}-1} \frac{(x\mu_{SR})^q}{q!e^{x\mu_{SR}}} \int_{y=0}^{\infty} \frac{y^{m_R-1}(y+1)^q}{e^{y(\mu_R+x\mu_{SR})}} dy. \quad (4.59)$$

Afterwards, the CDF of $X_1^{(c)}$ can be found in a closed-form and it is given by

$$F_{X_1^{(c)}}(x) = 1 - \frac{\mu_R^{m_R}}{e^{x\mu_{SR}}} \sum_{q=0}^{m_{SR}-1} \frac{(x\mu_{SR})^q}{q!} U(m_R, m_R + q + 1; \mu_R + x\mu_{SR}) \quad (4.60)$$

where $U(\cdot, \cdot; \cdot)$ confluent hypergeometric function [65, Eq. (9.211.4)]. By combining (4.56), (4.57) and (4.60), the exact outage probability expression for the OS policy can be obtained as

$$\begin{aligned} P_{OS}(\gamma_{th}) &= \left[1 - \frac{\mu_{RD}^{m_{RD}}\mu_R^{m_R}}{\Gamma(m_{RD})} \sum_{q=0}^{m_{SR}-1} \frac{(\gamma_{th}\mu_{SR})^q}{q!} \right. \\ &\quad \times \int_{y=\gamma_{th}}^{\infty} \frac{y^{m_{RD}-1}(y+1)^q}{e^{y\mu_{RD}+\gamma_{th}\mu_{SR}(y+1)/(y-\gamma_{th})}} \\ &\quad \left. \times \frac{U\left(m_R, m_R + q + 1; \mu_R + \frac{\gamma_{th}\mu_{SR}(y+1)}{y-\gamma_{th}}\right)}{(y-\gamma_{th})^q} dy \right]^N. \end{aligned} \quad (4.61)$$

Unfortunately, there is no closed-form solution for the above expression. In order to find a closed-form solution, the lower-bound of the OS policy is investigated. By

substituting (4.56) and (4.60) into (4.54), the closed-form lower-bound expression of OS policy can be easily found as

$$P_{OS,low}(\gamma_{th}) = \left[1 - \frac{\Gamma(m_{RD}, \gamma_{th}\mu_{RD}) \mu_R^{m_R}}{\Gamma(m_{RD}) e^{\gamma_{th}\mu_{SR}}} \sum_{q=0}^{m_{SR}-1} (\gamma_{th}\mu_{SR})^q \times \frac{U(m_R, m_R + q + 1; \mu_R + \gamma_{th}\mu_{SR})}{q!} \right]^N. \quad (4.62)$$

In order to find the asymptotic outage probability expression for OS, we need to use both (4.55) and the above lower-bound expression as follows

$$P_{OS}^\infty(\gamma_{th}) = \lim_{\substack{P_S \rightarrow \infty \\ P_R \rightarrow \infty}} \left[1 - \frac{\Gamma(m_{RD}, \gamma_{th}\mu_{RD})}{\Gamma(m_{RD}) e^{\gamma_{th}\mu_{SR}}} \sum_{q=0}^{m_{SR}-1} \mu_R^{m_R+q} \times \frac{U(m_R, m_R + q + 1; \mu_R + \gamma_{th}\mu_{SR})}{(\zeta\gamma_{th})^{-q} q!} \right]^N. \quad (4.63)$$

In the limit case, $\{P_S, P_R\} \rightarrow \infty$ means $\{\mu_{SR}, \mu_R\} \rightarrow 0$. By using the definition of incomplete gamma function, one can easily show that $\lim_{x \rightarrow 0} \Gamma(\alpha, x)/e^x = \Gamma(\alpha)$. Hence, the above asymptotic expression becomes

$$P_{OS}^\infty(\gamma_{th}) = \left[1 - \sum_{q=0}^{m_{SR}-1} \frac{(\zeta\gamma_{th})^q}{q!} \left(\lim_{\substack{P_S \rightarrow \infty \\ P_R \rightarrow \infty}} \frac{U(m_R, m_R + q + 1; \mu_R + \gamma_{th}\mu_{SR})}{\mu_R^{-m_R-q}} \right) \right]^N. \quad (4.64)$$

In order to derive a closed-form solution for the asymptotic outage probability, the confluent hypergeometric function should be expanded as a single-integral by using [65, Eq. (9.211.4)]. Then, the limit expression is rewritten as

$$\lim_{\substack{P_S \rightarrow \infty \\ P_R \rightarrow \infty}} \frac{U(m_R, m_R + q + 1; \mu_R + \gamma_{th}\mu_{SR})}{\mu_R^{-m_R-q}} = \lim_{\substack{P_S \rightarrow \infty \\ P_R \rightarrow \infty}} \frac{1}{\Gamma(m_R)} \int_{y=0}^{\infty} \frac{y^{m_R-1} (y + \mu_R)^q}{e^{y(1+\zeta\gamma_{th})}} dy. \quad (4.65)$$

The limit operation can be eliminated by simply using $\lim_{x \rightarrow 0} (y + x)^q = y^q$. Then, by applying [65, Eq. (3.351.3)], the required limit expression can be found as

$$\lim_{\substack{P_S \rightarrow \infty \\ P_R \rightarrow \infty}} \frac{U(m_R, m_R + q + 1; \mu_R + \gamma_{th}\mu_{SR})}{\mu_R^{-m_R-q}} = \frac{\Gamma(m_R + q)}{\Gamma(m_R) (1 + \zeta\gamma_{th})^{m_R+q}}. \quad (4.66)$$

By substituting (4.66) into (4.64), the closed-form asymptotic outage probability expression for the OS policy is derived as

$$P_{OS}^\infty(\gamma_{th}) = \left[1 - \frac{1}{\Gamma(m_R)} \sum_{q=0}^{m_{SR}-1} \frac{(\zeta\gamma_{th})^q \Gamma(m_R + q)}{q! (1 + \zeta\gamma_{th})^{m_R+q}} \right]^N. \quad (4.67)$$

It is important to note that the above expression does not depend on the values of m_{RD} and Ω_{RD} .

4.3.3.2 Partial selection with S-R link and residual LI

When the SRLI policy is deployed, the distribution of $\gamma_{RD}^{(b_{SRLI})}$ remain the same as expressed in Section 4.3.1. However, the distribution of $\gamma_{SR}^{(b_{SRLI})}$ and $\gamma_R^{(b_{SRLI})}$ are not gamma distributed random variables because the relay selection applied by using both $\gamma_{SR}^{(c)}$ and $\gamma_R^{(c)}$. Therefore, the exact outage probability for the SRLI policy is given by

$$P_{SRLI}(\gamma_{th}) = 1 - \int_{y=\gamma_{th}}^{\infty} f_{\gamma_{RD}^{(c)}}(y) \bar{F}_{X_1^{(b_{SRLI})}} \left[\frac{\gamma_{th}(y+1)}{y - \gamma_{th}} \right] dy. \quad (4.68)$$

The required CDF expression belonging to $X_1^{(b_{SRLI})}$ can be written as

$$F_{X_1^{(b_{SRLI})}}(x) = P \left(\max_{1 \leq c \leq N} \{X_1^{(c)}\} \leq x \right) = \left[F_{X_1^{(c)}}(x) \right]^N. \quad (4.69)$$

By substituting (4.60) and (4.69) into (4.68), the exact outage probability for the SRLI policy is derived as

$$P_{SRLI}(\gamma_{th}) = 1 - \frac{\mu_{RD}^{m_{RD}}}{\Gamma(m_{RD})} \int_{y=\gamma_{th}}^{\infty} \frac{y^{m_{RD}-1}}{e^{y\mu_{RD}}} \left\{ 1 - \left[1 - \frac{\mu_R^{m_R}}{e^{\gamma_{th}\mu_{SR}(y+1)/(y-\gamma_{th})}} \right. \right. \\ \left. \left. \times \sum_{q=0}^{m_{SR}-1} \frac{1}{q!} \frac{U\left(m_R, m_R + q + 1; \mu_R + \frac{\gamma_{th}\mu_{SR}(y+1)}{y-\gamma_{th}}\right)}{\left(\frac{y-\gamma_{th}}{\gamma_{th}\mu_{SR}(y+1)}\right)^q} \right]^N \right\} dy. \quad (4.70)$$

To the best of our knowledge, there is no closed-form solution for $P_{SRLI}(\cdot)$. However, by combining (4.54), (4.60) and (4.69), the lower-bound outage probability expression can be derived in a closed-form and it is given by

$$P_{SRLI,low}(\gamma_{th}) = \frac{\gamma(m_{RD}, \gamma_{th}\mu_{RD})}{\Gamma(m_{RD})} + \frac{\Gamma(m_{RD}, \gamma_{th}\mu_{RD})}{\Gamma(m_{RD})} \left[1 - \frac{\mu_R^{m_R}}{e^{\gamma_{th}\mu_{SR}}} \right. \\ \left. \times \sum_{q=0}^{m_{SR}-1} (\gamma_{th}\mu_{SR})^q \frac{U(m_R, m_R + q + 1; \mu_R + \gamma_{th}\mu_{SR})}{q!} \right]^N. \quad (4.71)$$

When we substitute the lower-bound expression into (4.54) and use both $\lim_{x \rightarrow 0} \gamma(\alpha, x) = 0$ and $\lim_{x \rightarrow 0} \Gamma(\alpha, x)/e^x = \Gamma(\alpha)$, the asymptotic outage probability for the SRLI policy can be written as

$$P_{SRLI}^{\infty}(\gamma_{th}) = \left[1 - \sum_{q=0}^{m_{SR}-1} \frac{(\zeta\gamma_{th})^q}{q!} \left(\lim_{\substack{P_S \rightarrow \infty \\ P_R \rightarrow \infty}} \frac{U(m_R, m_R + q + 1; \mu_R + \gamma_{th}\mu_{SR})}{\mu_R^{-m_R-q}} \right) \right]^N. \quad (4.72)$$

One can show that the asymptotic outage probabilities of both the OS and SRLI policies are identical, which are respectively given in (4.64) and (4.72). Hence, by

using (4.67), the the asymptotic outage probability expression of the SRLI policy can be easily written as follows

$$P_{SRLI}^{\infty}(\gamma_{th}) = \left[1 - \frac{1}{\Gamma(m_R)} \cdot \sum_{q=0}^{m_{SR}-1} \frac{(\zeta \gamma_{th})^q \Gamma(m_R + q)}{q! (1 + \zeta \gamma_{th})^{m_R+q}} \right]^N. \quad (4.73)$$

4.3.3.3 Partial selection with S-R link

In the SR policy defined as the third relay selection policy, $\gamma_{RD}^{(b_{SR})}$ and $\gamma_R^{(b_{SR})}$ are distributed according to the gamma distribution because the relay selection is utilized with the values of $\gamma_{SR}^{(c)}$. So, the exact outage probability of the SR policy can be easily given by

$$P_{SR}(\gamma_{th}) = 1 - \int_{y=\gamma_{th}}^{\infty} f_{\gamma_{RD}^{(c)}}(y) \bar{F}_{X_1^{(b_{SR})}} \left[\frac{\gamma_{th}(y+1)}{y - \gamma_{th}} \right] dy. \quad (4.74)$$

In order to find the CDF of $X_1^{(b_{SR})}$, we need the CDF of $\gamma_{SR}^{(b_{SR})}$, which is given by

$$F_{\gamma_{SR}^{(b_{SR})}}(x) = P \left(\max_{1 \leq c \leq N} \{ \gamma_{SR}^{(c)} \} \leq x \right) = \left[\frac{\gamma(m_{SR}, x \mu_{SR})}{\Gamma(m_{SR})} \right]^N. \quad (4.75)$$

Afterwards, the CDF of $X_1^{(b_{SR})}$ with the help of [65, Eq. (8.352.6)] is given by

$$\begin{aligned} F_{X_1^{(b_{SR})}}(x) &= \int_{y=0}^{\infty} f_{\gamma_R^{(c)}}(y) F_{\gamma_{SR}^{(b_{SR})}}[x(y+1)] dy \\ &= \frac{\mu_R^{m_R}}{\Gamma(m_R)} \int_{y=0}^{\infty} \frac{y^{m_R-1}}{e^{y\mu_R}} \left[1 - \sum_{u=0}^{m_{SR}-1} \frac{[x(y+1)\mu_{SR}]^u}{e^{x(y+1)\mu_{SR}u!}} \right]^N dy. \end{aligned} \quad (4.76)$$

In order to solve above integration, it is required to apply the binomial expansion and multinomial coefficients as in [78, Eq. (7)]. Then, the CDF is rewritten as

$$\begin{aligned} F_{X_1^{(b_{SR})}}(x) &= \frac{\mu_R^{m_R}}{\Gamma(m_R)} \sum_{k=0}^N \binom{N}{k} \frac{(-1)^k}{e^{kx\mu_{SR}}} \int_{y=0}^{\infty} \frac{y^{m_R-1}}{e^{y(\mu_R+kx\mu_{SR})}} \left[\sum_{u=0}^{m_{SR}-1} \frac{[x(y+1)\mu_{SR}]^u}{u!} \right]^k dy \\ &= \frac{\mu_R^{m_R}}{\Gamma(m_R)} \sum_{k=0}^N \sum_{q=0}^{\theta_1} \binom{N}{k} \frac{\Phi_1^q(x\mu_{SR})^q}{e^{kx\mu_{SR}}} \int_{y=0}^{\infty} \frac{y^{m_R-1}(y+1)^q}{e^{y(\mu_R+kx\mu_{SR})}} dy \end{aligned} \quad (4.77)$$

where $\theta_1 = k(m_{SR}-1)$ and Φ_1^q is the coefficient of $[x(y+1)\mu_{SR}]^q$ in the expansion of

$$\left[- \sum_{t=0}^{m_{SR}-1} \frac{[x(y+1)\mu_{SR}]^t}{t!} \right]^k. \quad (4.78)$$

By applying [65, Eq. (9.211.4)], the CDF in a closed-form can be found as

$$F_{X_1^{(b_{SR})}}(x) = \mu_R^{m_R} \sum_{k=0}^N \sum_{q=0}^{\theta_1} \binom{N}{k} \frac{\Phi_1^q(x\mu_{SR})^q}{e^{kx\mu_{SR}}} U(m_R, m_R+q+1; \mu_R+kx\mu_{SR}). \quad (4.79)$$

Finally, by substituting (4.79) into (4.74), the exact outage probability expression is obtained as

$$P_{SR}(\gamma_{th}) = \frac{\Gamma(m_{RD}, \gamma_{th}\mu_{RD})}{\Gamma(m_{RD})} + \frac{\mu_{RD}^{m_{RD}} \mu_R^{m_R}}{\Gamma(m_{RD})} \sum_{k=0}^N \sum_{q=0}^{\theta_1} \binom{N}{k} \frac{\Phi_1^q}{(\gamma_{th}\mu_{SR})^{-q}} \\ \times \int_{y=\gamma_{th}}^{\infty} \frac{y^{m_{RD}-1} \left(\frac{y+1}{y-\gamma_{th}}\right)^q U\left(m_R, m_R+q+1; \mu_R + \frac{k\gamma_{th}\mu_{SR}(y+1)}{y-\gamma_{th}}\right)}{e^{y\mu_{RD}+k\gamma_{th}\mu_{SR}(y+1)/(y-\gamma_{th})}} dy. \quad (4.80)$$

There is again no closed-form solution for the exact outage probability. By using (4.54) and (4.79), the closed-form lower-bound expression can be derived as

$$P_{SR,low}(\gamma_{th}) = 1 - \frac{\Gamma(m_{RD}, \gamma_{th}\mu_{RD})}{\Gamma(m_{RD})} \left[1 - \mu_R^{m_R} \sum_{k=0}^N \sum_{q=0}^{\theta_1} \binom{N}{k} \right. \\ \left. \times \frac{\Phi_1^q U(m_R, m_R+q+1; \mu_R + k\gamma_{th}\mu_{SR})}{(\gamma_{th}\mu_{SR})^{-q} e^{k\gamma_{th}\mu_{SR}}} \right]. \quad (4.81)$$

The asymptotic outage probability of the SR policy can also be written by using both the lower-bound expression and (4.55). Then, by applying the limit case given in (4.66), the closed-form asymptotic expression is obtained as follows

$$P_{SR}^{\infty}(\gamma_{th}) = \frac{1}{\Gamma(m_R)} \sum_{k=0}^N \sum_{q=0}^{\theta_1} \binom{N}{k} \frac{\Phi_1(\zeta\gamma_{th})^q \Gamma(m_R+q)}{(1+k\zeta\gamma_{th})^{m_R+q}}. \quad (4.82)$$

Similar to OS and SRLI, the asymptotic outage probability of the SR policy is independent from m_{RD} and μ_{RD} .

4.3.3.4 Partial selection with residual LI

According to the RLI policy for reducing the effect of residual LI caused by FD transmission, the exact outage probability of the OWFD-AF-RS system can be easily written as

$$P_{RLI}(\gamma_{th}) = 1 - \int_{y=\gamma_{th}}^{\infty} f_{\gamma_{RD}^{(e)}}(y) \bar{F}_{X_1^{(b_{RLI})}} \left[\frac{\gamma_{th}(y+1)}{y-\gamma_{th}} \right] dy. \quad (4.83)$$

As in the SR policy, it is required to find CDF of $X_1^{(b_{RLI})}$. The desired CDF expression can be found via the PDF of $\gamma_R^{(b_{RLI})}$. By using the property of

$f_{\gamma_R^{(b_{RLI})}}(\cdot) = \frac{\partial}{\partial x} F_{\gamma_R^{(b_{RLI})}}(\cdot)$, the PDF can be given by

$$\begin{aligned} f_{\gamma_R^{(b_{RLI})}}(x) &= \frac{\partial}{\partial x} P\left(\min_{1 \leq c \leq N} \{\gamma_R^{(c)}\} \leq x\right) = \frac{\partial}{\partial x} \left[1 - \prod_{c=1}^N \bar{F}_{\gamma_R^{(b_{RLI})}}(x)\right] \\ &= \frac{\partial}{\partial x} \left[\frac{\Gamma(m_R, x\mu_R)}{\Gamma(m_R)}\right]^N \\ &= \frac{N\mu_R^{m_R} x^{m_R-1} [\Gamma(m_R, x\mu_R)]^{N-1}}{e^{x\mu_R} [\Gamma(m_R)]^N}. \end{aligned} \quad (4.84)$$

Then, by applying [65, Eq. (8.352.6)], the CDF of $X_1^{(b_{RLI})}$ for the integer values of m_{SR} can be written as follows

$$\begin{aligned} F_{X_1^{(b_{RLI})}}(x) &= \int_{y=0}^{\infty} f_{\gamma_R^{(b_{RLI})}}(y) F_{\gamma_{SR}^{(c)}}[x(y+1)] dy \\ &= 1 - \frac{N\mu_R^{m_R}}{[\Gamma(m_R)]^N} \sum_{u=0}^{m_{SR}-1} \frac{(x\mu_{SR})^u}{u! e^{x\mu_{SR}}} \int_{y=0}^{\infty} \frac{(y+1)^u [\Gamma(m_R, y\mu_R)]^{N-1}}{y^{1-m_R} e^{y(\mu_R+x\mu_{SR})}} dy. \end{aligned} \quad (4.85)$$

Before starting to the solution of the integration, the above expression should be expanded for the integer values of m_R . Then, the CDF of $X_1^{(b_{RLI})}$ is rewritten as

$$\begin{aligned} F_{X_1^{(b_{RLI})}}(x) &= 1 - \frac{N\mu_R^{m_R}}{\Gamma(m_R)} \sum_{u=0}^{m_{SR}-1} \frac{(x\mu_{SR})^u}{u! e^{x\mu_{SR}}} \\ &\quad \times \int_{y=0}^{\infty} \frac{y^{m_R-1} (y+1)^u}{e^{y(N\mu_R+x\mu_{SR})}} \left[\sum_{k=0}^{m_R-1} \frac{(y\mu_R)^k}{k!} \right]^{N-1} dy. \end{aligned} \quad (4.86)$$

Similar to (4.77), the above expression can be derived as

$$\begin{aligned} F_{X_1^{(b_{RLI})}}(x) &= \frac{N}{\Gamma(m_R)} \sum_{u=0}^{m_{SR}-1} \sum_{q=0}^{\theta_2} \frac{\Phi_2^q (x\mu_{SR})^u \Gamma(m_R+q)}{\mu_R^{-m_R-q} u! e^{x\mu_{SR}}} \\ &\quad \times U(m_R+q, m_R+q+u+1; N\mu_R+x\mu_{SR}) \end{aligned} \quad (4.87)$$

where $\theta_2 = (N-1)(m_R-1)$ and Φ_2^q is the coefficient $(y\mu_R)^q$ in the expansion of

$$\left[\sum_{k=0}^{m_R-1} \frac{(y\mu_R)^k}{k!} \right]^{N-1}. \quad (4.88)$$

By substituting (4.87) into (4.83), the exact outage probability is obtained as

$$\begin{aligned} P_{RLI}(\gamma_{th}) &= 1 - \frac{N\mu_{RD}^{m_{RD}}}{\Gamma(m_R)\Gamma(m_{RD})} \sum_{u=0}^{m_{SR}-1} \sum_{q=0}^{\theta_2} \frac{\Phi_2^q \mu_R^{m_R+q}}{u!} \\ &\quad \times \frac{\Gamma(m_R+q)}{(\gamma_{th}\mu_{SR})^{-u}} \int_{y=\gamma_{th}}^{\infty} \frac{y^{m_{RD}-1} \left(\frac{y+1}{y-\gamma_{th}}\right)^u}{e^{y\mu_{RD}+\gamma_{th}\mu_{SR}(y+1)/(y-\gamma_{th})}} \\ &\quad \times U\left(m_R+q, m_R+q+u+1; N\mu_R + \frac{\gamma_{th}\mu_{SR}(y+1)}{y-\gamma_{th}}\right) dy. \end{aligned} \quad (4.89)$$

As seen, the exact outage probability belonging to the RLI policy is also derived in a single-integral form. Then, by using (4.54) and (4.87), the closed-form lower-bound expression can be given by

$$P_{RLI,low}(\gamma_{th}) = 1 - \frac{N \Gamma(m_{RD}, \gamma_{th} \mu_{RD})}{\Gamma(m_R) \Gamma(m_{RD})} \sum_{u=0}^{m_{SR}-1} \sum_{q=0}^{\theta_2} \frac{\Phi_2^q(\gamma_{th} \mu_{SR})^u \mu_R^{m_R+q}}{u! e^{\gamma_{th} \mu_{SR}}} \times \Gamma(m_R + q) U(m_R + q, m_R + q + u + 1; N \mu_R + \gamma_{th} \mu_{SR}). \quad (4.90)$$

The asymptotic outage probability expression for the RLI policy can be expressed in a closed-form by using the lower-bound, (4.55), the limit cases given for the SRLI policy and (4.66), respectively. Finally, the asymptotic outage probability is derived as

$$P_{RLI}^{\infty}(\gamma_{th}) = 1 - \frac{N}{\Gamma(m_R)} \sum_{u=0}^{m_{SR}-1} \sum_{q=0}^{\theta_2} \frac{\Phi_2^q(\zeta \gamma_{th})^u \Gamma(m_R + q + u)}{u! (N + \zeta \gamma_{th})^{m_R+q+u}}. \quad (4.91)$$

4.3.3.5 Partial selection with R-D link

In the last relay selection policy as RD, the values of $\gamma_{RD}^{(c)}$ are used for the relay selection. Hence, the $\gamma_{SR}^{(b_{RD})}$ and $\gamma_R^{(b_{RD})}$ are gamma distributed random variables and the exact outage probability can be given by

$$P_{RD}(\gamma_{th}) = 1 - \int_{y=\gamma_{th}}^{\infty} f_{\gamma_{RD}^{(b_{RD})}}(y) \bar{F}_{X_1^{(c)}} \left[\frac{\gamma_{th}(y+1)}{y - \gamma_{th}} \right] dy \quad (4.92)$$

At the beginning, the CDF of $\gamma_{RD}^{(b_{RD})}$ can be easily written as follows

$$F_{\gamma_{RD}^{(b_{RD})}}(x) = P \left(\max_{1 \leq c \leq N} \{ \gamma_{RD}^{(c)} \} \leq x \right) = \left[\frac{\gamma(m_{RD}, x \mu_{RD})}{\Gamma(m_{RD})} \right]^N. \quad (4.93)$$

By taking the derivative of above CDF expression, the PDF of $\gamma_{RD}^{(b_{RD})}$ can be easily written as follows

$$f_{\gamma_{RD}^{(b_{RD})}}(x) = \frac{N \mu_{RD}^{m_{RD}} x^{m_{RD}-1} [\gamma(m_{RD}, x \mu_{RD})]^{N-1}}{e^{x \mu_{RD}} [\Gamma(m_{RD})]^N}. \quad (4.94)$$

Then, by substituting (4.60) and (4.94) into (4.92), the exact outage probability in a single-integral form is obtained as

$$P_{RD}(\gamma_{th}) = 1 - \frac{N \mu_{RD}^{m_{RD}} \mu_R^{m_R}}{[\Gamma(m_{RD})]^N} \sum_{t=0}^{m_{SR}-1} \frac{(\gamma_{th} \mu_{SR})^t}{t!} \times \int_{y=\gamma_{th}}^{\infty} \frac{y^{m_{RD}-1} (y+1)^t [\gamma(m_{RD}, y \mu_{RD})]^{N-1}}{e^{y \mu_{RD} + \gamma_{th} \mu_{SR}(y+1)/(y-\gamma_{th})}} \times \frac{U \left(m_R, m_R + t + 1; \mu_R + \frac{\gamma_{th} \mu_{SR}(y+1)}{y-\gamma_{th}} \right)}{(y - \gamma_{th})^t} dy. \quad (4.95)$$

To the best of our knowledge, there is again no closed-form solution of the above expression. By using (4.54), (4.60) and (4.93), the closed-form lower-bound expression can be written as

$$P_{RD,low}(\gamma_{th}) = 1 - \left\{ 1 - \left[\frac{\gamma(m_{RD}, \gamma_{th}\mu_{RD})}{\Gamma(m_{RD})} \right]^N \right\} \frac{\mu_R^{m_R}}{e^{\gamma_{th}\mu_{SR}}} \times \sum_{t=0}^{m_{SR}-1} \frac{(\gamma_{th}\mu_{SR})^t}{t!} U(m_R, m_R + t + 1; \mu_R + \gamma_{th}\mu_{SR}). \quad (4.96)$$

The asymptotic outage probability of the RD policy can be written by using (4.54) and (4.96). By applying $\lim_{x \rightarrow 0} \gamma(\alpha, x) = 0$ and $\lim_{x \rightarrow 0} e^{-x} = 1$, respectively, the asymptotic outage probability is obtained as

$$P_{RD}^{\infty}(\gamma_{th}) = 1 - \frac{1}{\Gamma(m_R)} \sum_{t=0}^{m_{SR}-1} \frac{(\zeta\gamma_{th})^t \Gamma(m_R + q)}{t!(1 + \zeta\gamma_{th})^{m_R+q}} \quad (4.97)$$

The above expression implies that the asymptotic outage probability of the RD policy is independent from the number of relays N . Moreover, the asymptotic outage probability of the RD policy is equal to that of the OS and SRLI policies when there is only one relay ($N = 1$) in the OWFD-AF-RS system.

In a nutshell, all the exact outage probabilities for each relay selection policy are derived in a single-integral form, which are easily numerically evaluated by using common mathematical software packages such as MATLAB, MATHEMATICA and MAPLE. Moreover, both the lower-bound and asymptotic outage probability expressions are provided in the closed-forms.

4.3.4 Numerical results

Theoretical and Monte Carlo type computer simulation results for the outage performance of the variable-gain OWFD-AF-RS systems over Nakagami- m fading channels are represented throughout this section. The accuracy of all theoretical analyses for each relay selection policy are validated via comprehensive simulation scenarios. The total transmission power of the OWFD-AF-RS systems is defined as $P_T = P_S + P_R$. During the numerical analyses, we also assume that $\Omega_{SR} = \Omega_{RD} = 1$ and $\sigma_R^2 = \sigma_D^2 = 1$.

Figure 4.12 demonstrates the outage probability of the OWFD-AF-RS systems using the proposed relay selection policies versus different values of P_T , where $N = 4$,

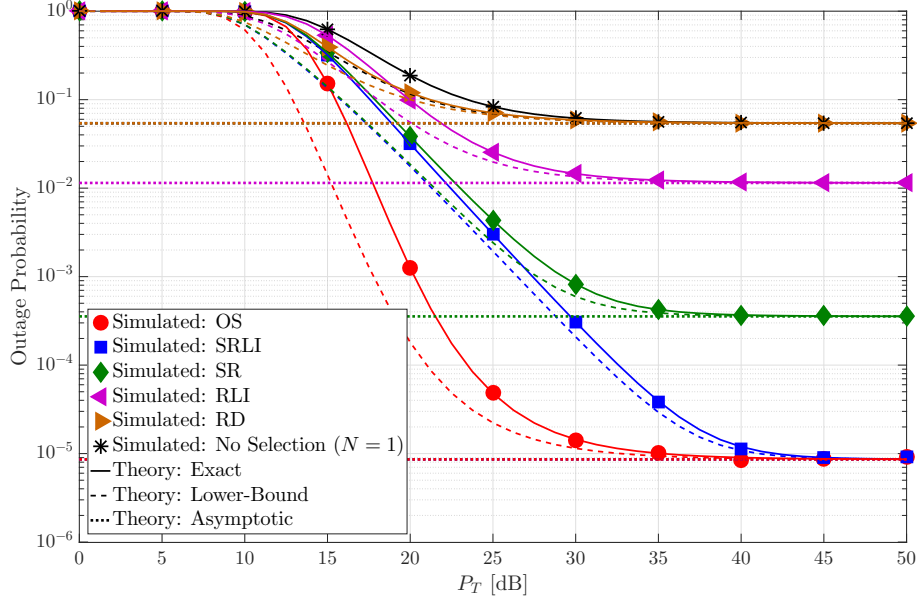


Figure 4.12 : Outage probability of OWFD-AF-RS systems versus P_T when $N = 4$, $\gamma_{th} = 5$, $\Omega_R = 0.05$, $P_S = P_R$, $m_{SR} = 3$, $m_{RD} = 2$ and $m_R = 4$.

$\gamma_{th} = 5$, $\Omega_R = 0.05$, $P_S = P_R$, $m_{SR} = 3$, $m_{RD} = 2$ and $m_R = 4$. Here, the OWFD-AF-RS systems are also compared with the no relay selection case, i.e., when there is a single relay ($N = 1$) in the system. It is shown that the computer simulation results are perfectly matched with the exact outage probability curves for the OS, SRLI, SR, RLI and RD policies, which are produced by using (4.61), (4.70), (4.80), (4.89) and (4.95). Hence, it proves the accuracy of analyses. In the light of obtained numerical results, OS is the best relay selection policy as expected. Although the performance of the SR and SRLI policies are close to each other in the low and medium SNR region, the performance of the SRLI policy converges to the OS policy in the high SNR region. Furthermore, the performance of the RD policy is better than the RLI policy in the low SNR region, while RLI outperforms RD when the values of P_T exceeds 18.9 dB in this scenario. In addition, the outage performance of OWFD-AF-RS system is tightly lower-bounded with (4.62), (4.71), (4.81), (4.90) and (4.96) especially in the medium and high SNR region. For instance, the difference between the lower-bound and exact outage probability curves for the OS, SRLI and SR is only 2.4 dB, 0.9 dB and 1.5 dB, respectively, at the outage probability value of 10^{-3} . It is again observed that the residual LI terms due the FD transmission limit performance of the OWFD-AF-RS system and cause the error floor at the medium and high SNR region for each relay selection policy. By using the asymptotic outage probability expressions given in (4.67), (4.73), (4.82), (4.91) and (4.97), the error floors can be easily and accurately

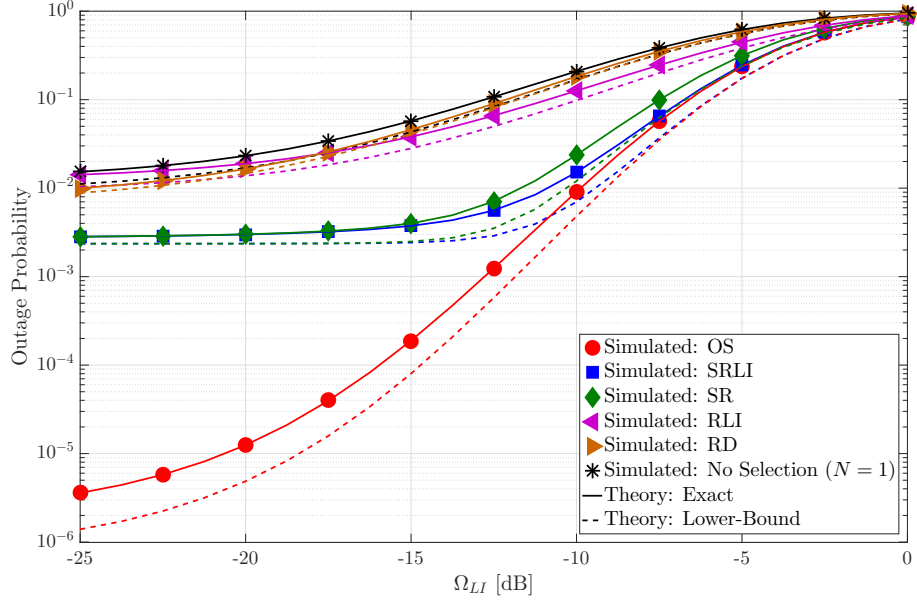


Figure 4.13 : Outage probability of OWFD-AF-RS systems versus Ω_R when $N = 3$, $\gamma_{th} = 3$, $P_T = 20$ dB, $P_S = P_R$, $m_{SR} = 2$, $m_{RD} = 2.5$ and $m_R = 5$.

predicted. When the relay selection policies are compared in terms of error floors in this scenario, the best policies are OS and SRLI with the same error floors at 8.6×10^{-6} and they are followed by the SR, RLI and RD policies with the error floors at 3.6×10^{-4} , 1.2×10^{-2} and 5.4×10^{-2} , respectively.

In Figure 4.13, the outage probability of the OWFD-AF-RS systems is plotted versus different values of the residual LI power Ω_R , where $N = 3$, $\gamma_{th} = 3$, $P_T = 20$ dB, $P_S = P_R$, $m_{SR} = 2$, $m_{RD} = 2.5$ and $m_R = 5$. In comparison to the no relay selection case ($N = 1$), the highest performance improvement can be achieved via the OS policy among all relay selection policies. It is observed the performance of SR and SRLI are converged as long as the residual LI decreases when the transmission power P_T is constant. Moreover, the advantage of the RLI policy vanishes when the efficiency of LI cancellation increases.

Figure 4.14 represents the outage probability of the OWFD-AF-RS systems for different values of γ_{th} , where $N = 2$, $\Omega_R = 0.02$, $P_T = 25$ dB, $P_S = P_R$, $m_{SR} = 2$, $m_{RD} = 2.5$ and $m_R = 5$. As expected, the outage performance degrades when γ_{th} increases for all relay selection policies. It is shown that the performance of the SRLI and SR policies converges to each other when the threshold end-to-end SINR is low. Additionally, we observe that the lower-bound curves become tighter for the lower values of γ_{th} .

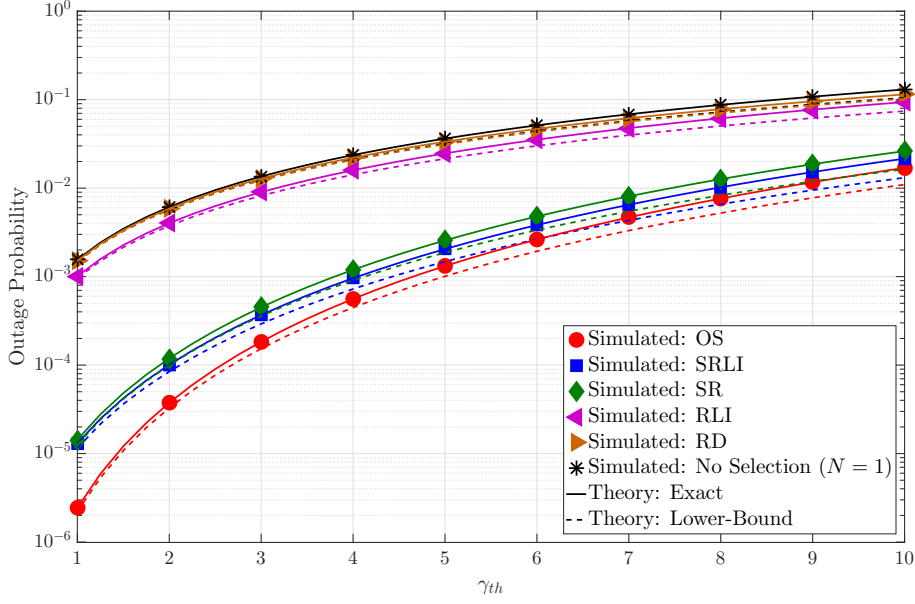


Figure 4.14 : Outage probability of OWFD-AF-RS systems versus γ_{th} when $N = 2$, $\Omega_R = 0.02$, $P_T = 25$ dB, $P_S = P_R$, $m_{SR} = 2$, $m_{RD} = 2.5$ and $m_R = 5$.

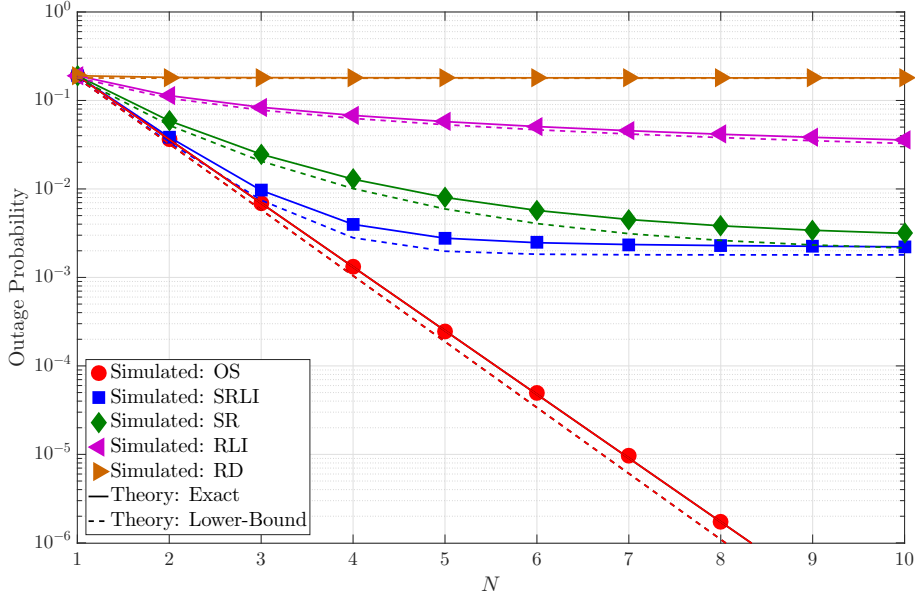


Figure 4.15 : Outage probability of OWFD-AF-RS systems versus N when $\gamma_{th} = 6$, $\Omega_R = 0.06$, $P_T = 30$ dB, $P_S = P_R$, $m_{SR} = 2$, $m_{RD} = 1.5$ and $m_R = 3$.

The outage probability of the OWFD-AF-RS system versus different values of N is illustrated in Figure 4.15, where $\gamma_{th} = 6$, $\Omega_R = 0.06$, $P_T = 30$ dB, $P_S = P_R$, $m_{SR} = 2$, $m_{RD} = 1.5$ and $m_R = 3$. The first and the most important outcome of this analysis is that the performance of the OS policy linearly improves as long as the number of available relay in the OWFD-AF-RS system increases. On the other hand, the outage probabilities of other four selection policies do not change after a certain number of relays utilized in the system. For example, the increase of N does not provide any improvement for RD and the performance of SRLI remains constant for $N \geq 5$.

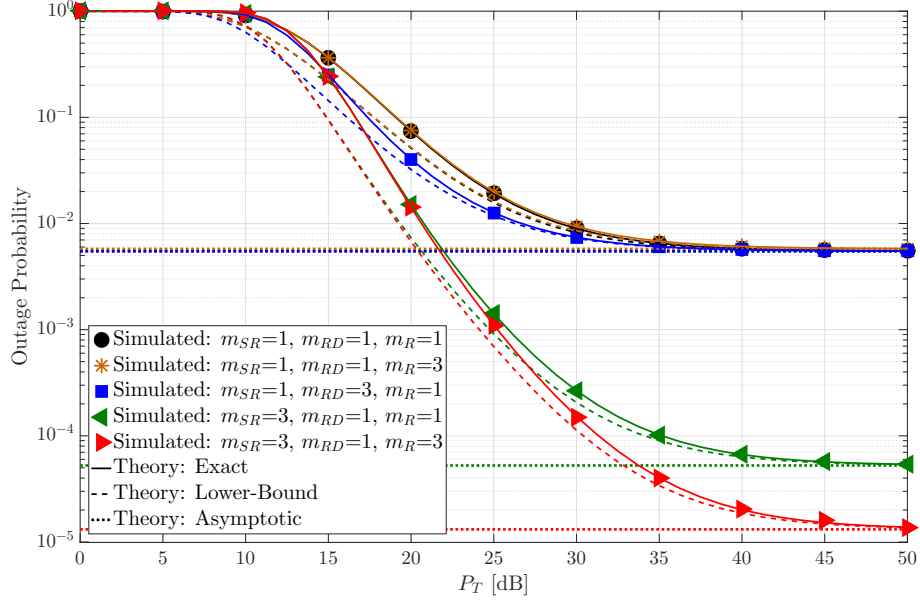


Figure 4.16 : Outage probability of OWFD-AF-RS systems with the OS policy versus P_T when $N = 2$, $\gamma_{th} = 4$, $\Omega_R = 0.01$ and $2P_S = P_R$.

Figure 4.16 shows the outage performance of the OWFD-AF-RS systems with the OS policy providing the best performance improvement for the OWFD-AF systems. Here, five different outage probability results over Nakagami- m fading channels for the various m_{SR} , m_{RD} and m_R values are plotted versus P_T , where $N = 2$, $\gamma_{th} = 4$, $\Omega_R = 0.01$ and $2P_S = P_R$. Firstly, the outage performance of the OWFD-AF-RS systems over Rayleigh fading channels is demonstrated by taking all m values as one. Then, the value of m_R is increased from 1 to 3 when $m_{SR} = m_{RD} = 1$. It is observed that there is no performance improvement for the higher values of m_R , moreover, we can see that the performance is even worsened when it is explored in more detail. To illustrate, the error floor in Rayleigh fading channel is occurred at 5.5×10^{-3} , while it increases to 5.8×10^{-3} when $m_{SR} = m_{RD} = 1$ and $m_R = 3$. When we increase the value of m_{RD} , i.e, $m_{SR} = m_R = 1$ and $m_{RD} = 3$, the performance advantage with respect to the Rayleigh fading channels in the low SNR region vanishes at the high SNR region and they converges to the same error floor. This result can be explained by using the asymptotic outage probability expression given in (4.67), which is independent from m_{RD} . However, the quite lower error floors can be achieved with the higher values of m_{SR} . For example, when m_{SR} is increased to 3, the error floor decreases to 5.2×10^{-5} , which is approximately 100 times lower than the error floor in Rayleigh fading channels. Furthermore, the slope of outage probability curve also increases in this scenario. Hence, m_{SR} is the dominant m parameter for the outage

performance of the variable-gain OWFD-AF-RS systems. Finally, we also observe that the effect of m_R in terms of the error floor depends on the value of m_{SR} .

Consequently, the outage probability of the OWFD-AF systems has been investigated in detail throughout this chapter^{1,2}. At the beginning, the fixed-gain OWFD-AF systems has been proposed and the outage performance over Rayleigh fading channels has been analyzed. By using the derived exact outage probability of the fixed-gain OWFD-AF systems, we have shown that it can provide quite better performance compared to the variable-gain case. Afterwards, the outage probability of the variable-gain and fixed-gain OWFD-AF systems has been investigated over Nakagami- m fading channels. The exact, lower-bound and asymptotic outage probability expressions have been obtained for both scheme. Finally, five different relay selection policies have been proposed in order to improve the outage performance of the variable-gain OWFD-AF systems. The exact, lower-bound and asymptotic outage probability expressions have been derived for each relay selection policy.

¹A part of this chapter is presented in *2016 Wireless Days Conf.* [79].

²Separate parts of this chapter are accepted for publication in *EURASIP J. Wireless Commun. Netw.* and submitted to *Dokuz Eylul Univ. Eng. Faculty J. Sci. Eng.*.

5. TWO-WAY FULL-DUPLEX AF RELAY SYSTEMS

In comparison to the conventional dual-hop cooperative systems, two-way transmission can be completed in a single time-slot with the two-way full-duplex amplify-and-forward (TWFD-AF) systems, where all nodes operate in FD mode and two source nodes exchange their information via the relay node using variable-gain AF relaying. The performance of TWFD-AF systems over Nakagami- m fading channels is analyzed in this chapter. Firstly, the exact, lower-bound and asymptotic outage probability expressions are derived for the TWFD-AF systems. In order to improve the outage performance of the TWFD-AF systems having multiple relays, the relay selection is employed according to the max-min policy. We also provide the exact, lower-bound and asymptotic outage probability expressions for the relay selection case. The accuracy of the theoretical analyses are verified via Monte Carlo type computer simulation results.

5.1 TWFD Variable-Gain AF Relay Systems over Nakagami- m Fading Channels

In the literature, there are a few works about the performance analyses of the variable-gain TWFD-AF systems. However, there is no outage performance analysis for the variable-gain TWFD-AF systems over Nakagami- m fading channels. In this work, the exact outage probability expression of the proposed system is obtained in a single-integral form by using the CDF of end-to-end SINR. Moreover, the tight lower-bound and asymptotic outage probability expressions are found in the closed-forms. The accuracy of the derived analytical expressions are supported by various Monte Carlo type computer simulation results. It is shown that the outage performance of the system can be enhanced as long as either the transmit power or the efficiency of LI cancellation process increases. We also observe that the outage probability converges to an error floor at high SINR values due to the residual LI component occurred at the source and relay nodes.

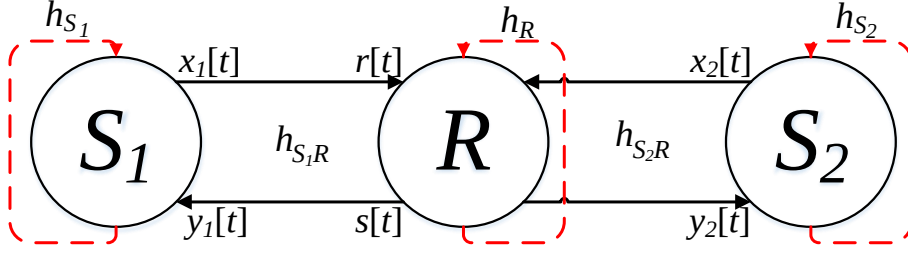


Figure 5.1 : TWFD-AF system model.

5.1.1 System model

A dual-hop transmission scheme is considered as shown in Figure 5.1 where there are two sources ($S_i, i \in \{1, 2\}$) and one relay (R). All the nodes in the TWFD-AF system are equipped with one transmit and one receive antenna for two-way transmissions. We assume that the sources communicate with each other via the relay node and there is no direct link between the sources. As seen from the system model, the complex channel coefficients between $S_i - R$ link are denoted by h_{S_iR} and reciprocal as in [32, 80]. Similarly, h_{S_i} and h_R are the complex residual LI channel coefficients at S_i and R nodes, respectively. The envelope of all the channel coefficients ($|h_{S_iR}|, |h_{S_i}|, |h_R|$) are modeled as Nakagami- m distribution random variables with average powers of $\Omega_{S_iR}, \Omega_{S_i}$ and Ω_R , respectively. The unit-power signals $x_1[t]$ and $x_2[t]$ are transmitted from S_1 and S_2 , respectively. Then, the received signal at the receive antenna of the relay node becomes

$$r[t] = h_{S_1R}\sqrt{P_S}x_1[t] + h_{S_2R}\sqrt{P_S}x_2[t] + h_Rs[t] + n_R[t] \quad (5.1)$$

where P_S is the transmission power of the sources, $s[t]$ is the transmitted signal by the relay and $n_R[t]$ is the complex Gaussian noise term at the receiver of the relay distributed according to $\mathcal{CN}(0, \sigma_R^2)$. The transmitted signal from the relay is

$$s[t] = Gr[t-1] \quad (5.2)$$

where G is the amplification factor. As seen from above expression, there is a one time-slot delay due to the relay processing [10]. By substituting (5.2) into (5.1) recursively, the transmitted signal by the relay can be expanded as

$$s[t] = G \sum_{u=0}^{\infty} (Gh_R)^u q[t-u] \quad (5.3)$$

where $q[t] = h_{S_1R}\sqrt{P_S}x_1[t] + h_{S_2R}\sqrt{P_S}x_2[t] + n_R[t]$. In order to normalize the transmission power at the relay as P_R , the amplification factor is chosen as

$$G = \sqrt{\frac{P_R}{P_S|h_{S_2R}|^2 + P_S|h_{S_1R}|^2 + P_R|h_R|^2 + \sigma_R^2}}. \quad (5.4)$$

By taking into consideration that $G|h_R| < 1$ given in (5.3), the instantaneous transmission power of the relay for any given h_{S_1R} , h_{S_2R} and h_R is found as

$$E\{|s[t]|^2 | h_{S_1R}, h_{S_2R}, h_R\} = \frac{G^2}{1-G^2|h_R|^2} [P_S|h_{S_1R}|^2 + P_S|h_{S_2R}|^2 + \sigma_R^2] = P_R. \quad (5.5)$$

Finally, the received signal at source node S_i is given by

$$y_i[t] = h_{S_iR}s[t] + h_{S_i}\sqrt{P_S}x_i[t] + n_{S_i}[t] \quad (5.6)$$

where $n_{S_i}[t]$ is the complex Gaussian noise term at the receiver of S_i distributed according to $\mathcal{CN}(0, \sigma_S^2)$. By combining (5.1), (5.2) and (5.6), the received signal at the source is rewritten as follows

$$\begin{aligned} y_i[t] = & \underbrace{Gh_{S_iR}h_{S_jR}\sqrt{P_S}x_j[t]}_{\text{Desired Signal}} + \underbrace{Gh_{S_iR}^2\sqrt{P_S}x_i[t]}_{\text{Removed by PLNC}} \\ & + \underbrace{Gh_{S_iR}h_Rs[t-1] + h_{S_i}\sqrt{P_S}x_i[t]}_{\text{Residual Loop-Interference}} \\ & + \underbrace{Gh_in_R[t-1] + n_{S_i}[t]}_{\text{Noise}} \end{aligned} \quad (5.7)$$

where $j \in \{1, 2\}$ and $j \neq i$. The information bearing signal for S_i node is the first term at the right hand side of the first line given in (5.7). The other term at the right hand side of the same line can be completely removed by S_i node since each source node has the information about its transmitted signal and the channel coefficients. The residual LI terms at the relay and source are given in the second line of (5.7). Similarly, the noise terms at the relay and source are given in the third line of (5.7). After some manipulations, the instantaneous end-to-end SINR at S_i node is obtained as

$$\gamma_{end,i} = \frac{\alpha \frac{\gamma_{S_iR}}{(\gamma_{S_i}+1)} \frac{\gamma_{S_jR}}{(\gamma_R+1)}}{\alpha \frac{\gamma_{S_iR}}{\gamma_{S_i}+1} + \frac{\gamma_{S_iR}+\gamma_{S_jR}}{\gamma_R+1} + 1} \quad (5.8)$$

where $\alpha = (P_R\sigma_R^2)/(P_S\sigma_S^2)$, $\gamma_{S_iR} = P_S|h_{S_iR}|^2/\sigma_R^2$, $\gamma_{S_jR} = P_S|h_{S_jR}|^2/\sigma_R^2$, $\gamma_{S_i} = P_S|h_{S_i}|^2/\sigma_S^2$ and $\gamma_R = P_R|h_R|^2/\sigma_R^2$. As mentioned above, the envelope of all channel coefficients are assumed to be Nakagami- m distributed. By defining

$\bar{\gamma}_{S_i R} = P_S \Omega_{S_i R} / \sigma_R^2$, $\bar{\gamma}_{S_j R} = P_S \Omega_{S_j R} / \sigma_R^2$, $\bar{\gamma}_{S_i} = P_S \Omega_{S_i} / \sigma_S^2$ and $\bar{\gamma}_R = P_R \Omega_R / \sigma_R^2$, the PDF and CDF of those random variables can be given by $f_{\gamma_A}(x) = \frac{\mu_A^{m_A} x^{m_A-1}}{\Gamma(m_A)} e^{-x\mu_A}$ and $F_{\gamma_A}(x) = \frac{\gamma(m_A, x\mu_A)}{\Gamma(m_A)}$ with $\mu_A = m_A / \bar{\gamma}_A$ and $A \in \{S_i R, S_j R, S_i, R\}$.

5.1.2 Outage probability analyses

In this section, we derive the exact, lower-bound and asymptotic outage probability expressions for the TWFD-AF system. The exact outage probability of the TWFD-AF system at S_i node is given by $P_{out,i}(\gamma_{th}) = F_{\gamma_{end,i}}(\gamma_{th})$ where $F_{\gamma_{end,i}}(\cdot)$ is the CDF of $\gamma_{end,i}$. By neglecting +1 term at the denominator of (5.8) for the simplicity, the instantaneous end-to-end SINR can be rewritten as

$$\begin{aligned} \gamma_{end,i} &\approx \frac{\alpha \frac{\gamma_{S_i R}}{(\gamma_{S_i}+1)} \frac{\gamma_{S_j R}}{(\gamma_R+1)}}{\alpha \frac{\gamma_{S_i R}}{\gamma_{S_i}+1} + \frac{\gamma_{S_i R} + \gamma_{S_j R}}{\gamma_R+1}} = \frac{\alpha \frac{\gamma_{S_i R} \gamma_{S_j R}}{\gamma_{S_i}+1}}{\alpha \frac{\gamma_{S_i R} [\gamma_R+1 + (\gamma_{S_i}+1)/\alpha]}{\gamma_{S_i}+1} + \gamma_{S_j R}} \\ &= \frac{Y_i Z_i}{Y_i + Z_i} \end{aligned} \quad (5.9)$$

where $Y_i = \alpha \frac{\gamma_{S_i R}}{\gamma_{S_i}+1}$ and $Z_i = \frac{\gamma_{S_j R}}{\gamma_R+1 + (\gamma_{S_i}+1)/\alpha}$. As well known in the FD literature, the power of residual LI channels should be much smaller than the direct link in order to ensure performance improvement with respect to the conventional HD transmission. Therefore, the dominant parameters are $\gamma_{S_i R}$ and $\gamma_{S_j R}$ inside Y_i and Z_i . For the sake of mathematical calculations, we assume that Y_i and Z_i are independent from each other. As we will show in the Numerical Results Section, this is an appropriate assumption for the TWFD-AF systems. Hence, by using [32, Eq. (37)], $F_{\gamma_{end,i}}(\cdot)$ can be given by

$$\begin{aligned} F_{\gamma_{end,i}}(x) &= P\left(\frac{Y_i Z_i}{Y_i + Z_i} \leq x\right) = \int_{y=0}^{\infty} P\left(\frac{y Z_i}{y + Z_i} \leq x\right) f_{Y_i}(y) dy \\ &= \int_{y=0}^x f_{Y_i}(y) dy + \int_{y=x}^{\infty} \left[1 - \bar{F}_{Z_i}\left(\frac{xy}{y-x}\right)\right] f_{Y_i}(y) dy \\ &= 1 - \int_{y=0}^{\infty} \bar{F}_{Z_i}\left(x + \frac{x^2}{y}\right) f_{Y_i}(y+x) dy \end{aligned} \quad (5.10)$$

where $\bar{F}_{Z_i}(\cdot) = 1 - F_{Z_i}(\cdot)$. The PDF of Y_i can be easily written by using [81] as

$$f_{Y_i}(y) = \frac{[\mu_{S_i R} / (\alpha \mu_{S_i}^*)]^{m_{S_i R}} y^{m_i-1}}{B(m_{S_i R}, m_{S_i}^*)} \left(1 + y \frac{\mu_{S_i R}}{\alpha \mu_{S_i}^*}\right)^{-m_{S_i R} - m_{S_i}^*} \quad (5.11)$$

where $\mu_{S_i}^* = \mu_{S_i}(\bar{\gamma}_{S_i} + 1)/\bar{\gamma}_{S_i}$, $m_{S_i}^* = m_{S_i}[(\bar{\gamma}_{S_i} + 1)/\bar{\gamma}_{S_i}]^2$, $B(\cdot, \cdot)$ is the Beta function [65, Eq. (8.384.1)]. Then, we can find the CDF of Z_i as follows

$$\begin{aligned}
F_{Z_i}(z) &= P(Z_i \leq z) = P\left(\frac{\gamma_{S_j R}}{\gamma_R + 1 + (\gamma_{S_i} + 1)/\alpha} \leq z\right) \\
&= \int_{x=0}^{\infty} \int_{y=0}^{\infty} F_{\gamma_{S_j R}}\left[\left(y + 1 + \frac{x+1}{\alpha}\right)z\right] f_{\gamma_{S_i}}(x) f_{\gamma_R}(y) dy dx \\
&= \frac{\mu_{S_i}^{m_{S_i}} \mu_R^{m_R}}{\Gamma(m_{S_j R}) \Gamma(m_{S_i}) \Gamma(m_R)} \int_{x=0}^{\infty} \int_{y=0}^{\infty} \frac{x^{m_{S_i}-1} y^{m_R-1}}{e^{x\mu_{S_i} + y\mu_{S_i}}} \\
&\quad \times \gamma\left[m_{S_j R}, \left(y + 1 + \frac{x+1}{\alpha}\right)z\mu_{S_j R}\right] dy dx.
\end{aligned} \tag{5.12}$$

For the integer values of $m_{S_j R}$, by using [65, Eq. (8.352.6)] and also the binomial expansion,

$$\begin{aligned}
F_{Z_i}(z) &= 1 - \frac{\mu_{S_i}^{m_{S_i}} \mu_R^{m_R} e^{-(1+\frac{1}{\alpha})z\mu_{S_j R}}}{\Gamma(m_{S_i}) \Gamma(m_R)} \sum_{n=0}^{m_{S_j R}-1} \sum_{u=0}^n \sum_{v=0}^u \frac{(z\mu_{S_j R})^n}{(n-u)!} \\
&\quad \times \frac{(1+\alpha)^{u-v}}{\alpha^u (u-v)! v!} \int_{x=0}^{\infty} \int_{y=0}^{\infty} \frac{y^{m_R+n-u-1}}{e^{y(z\mu_{S_j R} + \mu_R)}} \frac{x^{m_{S_i}+v-1}}{e^{x(\mu_{S_i} + z\mu_{S_j R}/\alpha)}} dy dx
\end{aligned} \tag{5.13}$$

can be obtained. As seen, two integrals are independent and very similar, so both of them can be easily solved by using [65, Eq. (3.351.3)]. Hence, the closed-form of $F_{Z_i}(\cdot)$ is given by

$$\begin{aligned}
F_{Z_i}(z) &= 1 - \frac{\mu_{S_i}^{m_{S_i}} \mu_R^{m_R} e^{-(1+\frac{1}{\alpha})z\mu_{S_j R}}}{\Gamma(m_{S_i}) \Gamma(m_R)} \sum_{n=0}^{m_{S_j R}-1} \sum_{u=0}^n \sum_{v=0}^u \frac{(z\mu_{S_j R})^n}{(n-u)!} \\
&\quad \times \frac{(1+\alpha)^{u-v} \Gamma(m_R + n - u) \Gamma(m_{S_i} + v)}{\alpha^u (u-v)! v! (z\mu_{S_j R} + \mu_R)^{m_R+n-u} (\mu_{S_i} + z\mu_{S_j R}/\alpha)^{m_{S_i}+v}}.
\end{aligned} \tag{5.14}$$

Finally, by substituting (5.11) and (5.14) into (5.10), the exact outage probability of the TWFD-AF system in a single-integral form can be obtained as

$$\begin{aligned}
P_{out,i}(\gamma_{th}) &= 1 - \frac{\left(\frac{\mu_{S_i R}}{\alpha \mu_{S_i}^*}\right)^{m_{S_i R}} \mu_{S_i}^{m_{S_i}} \mu_R^{m_R} e^{-(1+\frac{1}{\alpha})\gamma_{th} m_{S_j R}}}{B(m_{S_i R}, m_{S_i}^*) \Gamma(m_{S_i}) \Gamma(m_R)} \sum_{n=0}^{m_{S_j R}-1} \sum_{u=0}^n \sum_{v=0}^u \frac{\mu_{S_j R}^n}{\alpha^u} \\
&\quad \times \frac{\Gamma(m_R + n - u) \Gamma(m_{S_i} + v)}{(n-u)! (u-v)! v! (1+\alpha)^{v-u}} \int_{y=0}^{\infty} \frac{(\gamma_{th} + y)^{m_{S_i R}-1}}{e^{\gamma_{th}^2 (\alpha+1) \mu_{S_j R} / (\alpha y)}} \\
&\quad \times \frac{\left[1 + (\gamma_{th} + y) \frac{\mu_{S_i R}}{\alpha \mu_{S_i}^*}\right]^{-m_{S_i R}-m_{S_i}^*} \left(\gamma_{th} + \frac{\gamma_{th}^2}{y}\right)^n}{\left[\left(\gamma_{th} + \frac{\gamma_{th}^2}{y}\right) \mu_{S_j R} + \mu_R\right]^{m_R+n-u} \left[\mu_{S_i} + \left(\gamma_{th} + \frac{\gamma_{th}^2}{y}\right) \frac{\mu_{S_j R}}{\alpha}\right]^{m_{S_i}+v}} dy.
\end{aligned} \tag{5.15}$$

Unfortunately, to the best of our knowledge, there is no closed-form solution for $P_{out,i}(\cdot)$. However, it can be efficiently evaluated with common mathematical softwares such as MATLAB, MATHEMATICA or MAPLE.

In order to obtain a lower-bound for the outage probability of the TWFD-AF systems, we use the property of $\gamma_{end,i} = Y_i Z_i / (Y_i + Z_i) < \min \{Y_i, Z_i\}$. So, the lower-bound expression can be written as

$$\begin{aligned} P_{out,low,i}(\gamma_{th}) &= P(\min \{Y_i, Z_i\} \leq \gamma_{th}) = 1 - P(\min \{Y_i, Z_i\} > \gamma_{th}) \\ &= 1 - \bar{F}_{Y_i}(\gamma_{th}) \bar{F}_{Z_i}(\gamma_{th}) \end{aligned} \quad (5.16)$$

where $\bar{F}_{Y_i}(\cdot) = 1 - F_{Y_i}(\cdot)$. The CDF of Z_i was already found in (5.14). Therefore, we need to derive the CDF of Y_i by integrating its PDF given in (5.11) as follows

$$F_{Y_i}(y) = \int_0^y f_{Y_i}(x) dx = \frac{[\mu_{S_i R} / (\alpha \mu_{S_i}^*)]^{m_{S_i R}}}{B(m_{S_i R}, m_{S_i}^*)} \int_0^y \frac{x^{m_{S_i R}-1}}{\left(1 + x \frac{\mu_{S_i R}}{\alpha \mu_{S_i}^*}\right)^{m_{S_i R} + m_{S_i}^*}} dx. \quad (5.17)$$

By using [65, Eq. (3.194.1)], the CDF of Y_i can be easily obtained as

$$F_{Y_i}(y) = \frac{[y \mu_{S_i R} / (\alpha \mu_{S_i}^*)]^{m_{S_i R}}}{B(m_{S_i R}, m_{S_i}^*) m_{S_i R}} {}_2F_1\left(m_{S_i R} + m_{S_i}^*, m_{S_i R}; m_{S_i R} + 1; -y \frac{\mu_{S_i R}}{\alpha \mu_{S_i}^*}\right) \quad (5.18)$$

where ${}_2F_1(\cdot, \cdot; \cdot; \cdot)$ is Gauss hypergeometric function [65, Eq. (9.100)]. Now, we can derive $P_{out,low,i}$ in closed-form by substituting (5.14) and (5.18) into (5.16) as

$$\begin{aligned} P_{out,low,i}(\gamma_{th}) &= 1 - \left[\frac{\mu_{S_i}^{m_{S_i}} \mu_R^{m_R} e^{-(1+\frac{1}{\alpha})\gamma_{th}\mu_{S_j R}}}{\Gamma(m_{S_i}) \Gamma(m_R)} \sum_{n=0}^{m_{S_j R}-1} \sum_{u=0}^n \sum_{v=0}^u \frac{(\alpha+1)^{u-v}}{\alpha^u (n-u)!} \right. \\ &\quad \times \left. \frac{(\gamma_{th}\mu_{S_j R})^n \Gamma(m_R + n - u) \Gamma(m_{S_i} + v)}{(u-v)! v! (\gamma_{th}\mu_{S_j R} + \mu_R)^{m_R + n - u} (\mu_{S_i} + \gamma_{th}\mu_{S_j R}/\alpha)^{m_{S_i} + v}} \right] \\ &\quad \times \left[1 - \frac{[\gamma_{th}\mu_{S_i R} / (\alpha \mu_{S_i}^*)]^{m_i}}{B(m_{S_i R}, m_{S_i}^*) m_{S_i R}} {}_2F_1\left(m_{S_i R} + m_{S_i}^*, m_{S_i R}; m_{S_i R} + 1; -\gamma_{th} \frac{\mu_{S_i R}}{\alpha \mu_{S_i}^*}\right) \right]. \end{aligned} \quad (5.19)$$

For the further analysis, we also investigate the asymptotic outage probability of the TWFD-AF systems at high SNR region, i.e., $\{P_S, P_R\} \rightarrow \infty$ for fixed σ_S^2 and σ_R^2 . When SNR goes to infinity, the outage probability of the system is asymptotically obtained as

$$\begin{aligned} P_{out,i}^\infty(\gamma_{th}) &= \lim_{\substack{P_S \rightarrow \infty \\ P_R \rightarrow \infty}} P\left(\frac{Y_i Z_i}{Y_i + Z_i} \leq \gamma_{th}\right) \approx \lim_{\substack{P_S \rightarrow \infty \\ P_R \rightarrow \infty}} P(\min \{Y_i, Z_i\} \leq \gamma_{th}) \\ &= P(\min \{Y_i^\infty, Z_i^\infty\} \leq \gamma_{th}) \\ &= 1 - \bar{F}_{Y_i^\infty}(\gamma_{th}) \bar{F}_{Z_i^\infty}(\gamma_{th}) \end{aligned} \quad (5.20)$$

where $Y_i^\infty = \lim_{P_S, P_R \rightarrow \infty} Y_i$ and $Z_i^\infty = \lim_{P_S, P_R \rightarrow \infty} Z_i$, so their CDF expressions are required to obtain $P_{out,i}^\infty(\cdot)$. Firstly, the CDF of Y_∞ can be expressed as

$$\begin{aligned} F_{Y_i^\infty}(y) &= \lim_{\substack{P_S \rightarrow \infty \\ P_R \rightarrow \infty}} P(Y_i < y) = P\left(\frac{\alpha \gamma_{S_i R}}{\gamma_{S_i}} < y\right) = \int_{x=0}^{\infty} F_{\gamma_{S_i R}}\left(\frac{xy}{\alpha}\right) f_{\gamma_{S_i}}(x) dx \\ &= \frac{\mu_{S_i}^{m_{S_i}}}{\Gamma(m_{S_i R}) \Gamma(m_{S_i})} \int_{x=0}^{\infty} \frac{\gamma(m_i, xy \mu_{S_i R} / \alpha)}{e^{x \mu_{S_i}} x^{1-m_{S_i}}} dx. \end{aligned} \quad (5.21)$$

For the integer values of $m_{S_i R}$, by using [65, Eq. (8.352.6)] and the binomial expansion, $F_{Y_i^\infty}(\cdot)$ can be given by

$$F_{Y_i^\infty}(y) = 1 - \frac{\mu_{S_i}^{m_{S_i}}}{\Gamma(m_{S_i})} \sum_{k=0}^{m_{S_i R}-1} \frac{(y \mu_{S_i R} / \alpha)^k}{k!} \int_{x=0}^{\infty} \frac{x^{k+m_{S_i}-1}}{e^{x(\mu_{S_i} + y \mu_{S_i R} / \alpha)}} dx. \quad (5.22)$$

By using [65, Eq. (3.351.3)], the above integral is solved as

$$F_{Y_i^\infty}(y) = 1 - (1 + y \omega_1)^{-m_{S_i}} \sum_{k=0}^{m_{S_i R}-1} \frac{[1 + (y \omega_1)^{-1}]^{-k}}{B(m_{S_i}, k)} \quad (5.23)$$

where $\omega_1 = \frac{\mu_{S_i R}}{\alpha \mu_{S_i}} = \frac{P_S \Omega_{S_i} m_{S_i R}}{P_R \Omega_{S_i R} m_{S_i}}$. Then, the second CDF term $F_{Z_i^\infty}(\cdot)$ can be written as

$$\begin{aligned} F_{Z_i^\infty}(z) &= \lim_{\substack{P_S \rightarrow \infty \\ P_R \rightarrow \infty}} P(Z_i < z) = P\left(\frac{\gamma_{S_j R}}{\gamma_R + \gamma_{S_i} / \alpha} < z\right) \\ &= \int_{x=0}^{\infty} \int_{y=0}^{\infty} F_{\gamma_{S_j R}}\left[\left(y + \frac{x}{\alpha}\right) z\right] f_{\gamma_{S_i}}(x) f_{\gamma_R}(y) dy dx \\ &= \frac{\mu_{S_i}^{m_{S_i}} \mu_R^{m_R}}{\Gamma(m_{S_j R}) \Gamma(m_{S_i}) \Gamma(m_R)} \int_{x=0}^{\infty} \int_{y=0}^{\infty} \frac{x^{m_{S_i}-1} y^{m_R-1}}{e^{x \mu_{S_i} + y \mu_R}} \\ &\quad \times \gamma\left[m_{S_j R}, \left(y + \frac{x}{\alpha}\right) z \mu_{S_j R}\right] dy dx. \end{aligned} \quad (5.24)$$

Similarly as in the derivation of $F_{Y_i^\infty}(\cdot)$, by using [65, Eq. (8.352.6)], the binomial expansion and also [65, Eq. (3.351.3)], the above integration can be obtained as

$$F_{Z_i^\infty}(z) = 1 - \frac{(1 + z \omega_2)^{-m_{S_i}}}{(1 + z \omega_3)^{m_R}} \sum_{n=0}^{m_{S_j R}-1} \sum_{u=0}^n \frac{[1 + (z \omega_3)^{-1}]^{-n}}{B(m_{S_i}, u) B(m_R, n-u)} \left[\frac{\omega_3 \left(z + \frac{1}{\omega_2}\right)}{1 + z \omega_3} \right]^{-u} \quad (5.25)$$

where $\omega_2 = \frac{\mu_{S_j R}}{\alpha \mu_{S_i}} = \frac{P_S \Omega_{S_i} m_{S_j R}}{P_R \Omega_{S_j R} m_{S_i}}$ and $\omega_3 = \frac{\mu_{S_j R}}{\mu_R} = \frac{P_R \Omega_R m_{S_j R}}{P_S \Omega_{S_j R} m_R}$. At the end, the asymptotic outage probability is obtained in the closed-form as

$$\begin{aligned} P_{out,i}^\infty(\gamma_{th}) &= 1 - \left\{ (1 + \gamma_{th} \omega_1)^{-m_{S_i}} \sum_{k=0}^{m_{S_i R}-1} \frac{[1 + (\gamma_{th} \omega_1)^{-1}]^{-k}}{B(m_{S_i}, k)} \right\} \\ &\quad \times \left\{ \frac{(1 + \gamma_{th} \omega_2)^{-m_{S_i}}}{(1 + \gamma_{th} \omega_3)^{m_R}} \sum_{n=0}^{m_{S_j R}-1} \sum_{u=0}^n \frac{[1 + (\gamma_{th} \omega_3)^{-1}]^{-n}}{B(m_{S_i}, u) B(m_R, n-u)} \left[\frac{\omega_3 (\gamma_{th} + 1 / \omega_2)}{\gamma_{th} \omega_3 + 1} \right]^{-u} \right\} \end{aligned} \quad (5.26)$$

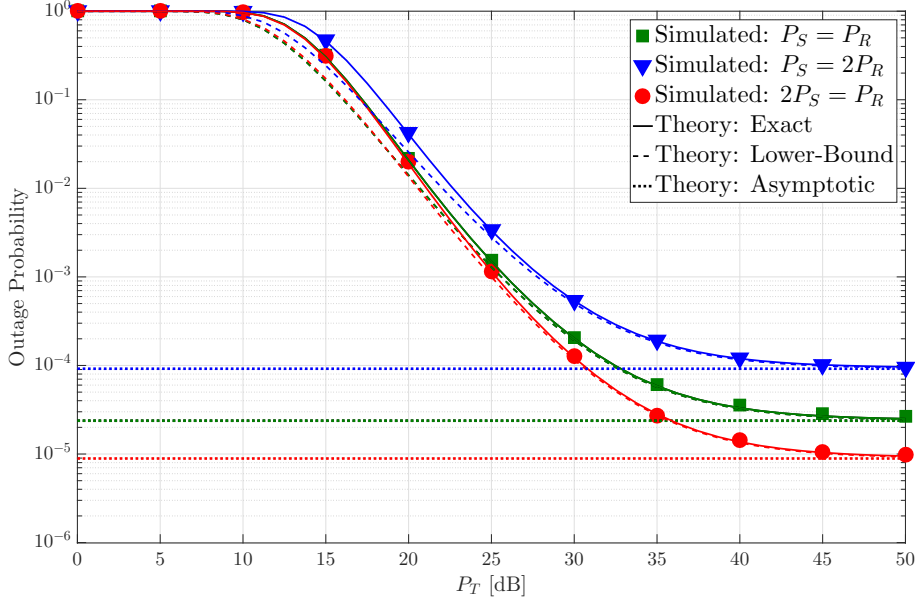


Figure 5.2 : Outage probability of the TWFD-AF systems at S_1 versus P_T when $\gamma_{th}=3$, $\Omega_{LI} = -30$ dB, $m_{S_1R} = 2$, $m_{S_2R} = m_R = 3$ and $m_{S_1} = 4$.

As seen, $P_{out,i}^\infty(\cdot)$ provides zero-diversity order as expected at high SNR region. Hence, $P_{out,i}^\infty(\cdot)$ is constant.

5.1.3 Numerical results

In this section, the outage theoretical and simulation results of the TWFD-AF systems over Nakagami- m fading channel are represented for different scenarios. The Monte Carlo simulations are performed for the outage probability at S_1 node, i.e. $P_{out,1}(\cdot)$. Without loss of generality, we assume that total transmission power is $P_T = P_S + P_R$, power of the relay-source links are chosen as $\Omega_{S_1R} = \Omega_{S_2R} = 1$, the residual LI terms at the relay and source nodes have the same power as $\Omega_{LI} = \Omega_R = \Omega_{S_1}$ and all the noise terms have unity variance $\sigma_R^2 = \sigma_S^2 = 1$ for all the simulation scenarios.

Figure 5.2 plots the outage performance of the TWFD-AF systems versus different values of P_T for $\gamma_{th} = 3$, $\Omega_{LI} = -30$ dB, $m_{S_1R} = 2$, $m_{S_2R} = m_R = 3$ and $m_{S_1} = 4$. It is shown that the exact outage probability results obtained from (5.15) are perfectly matched with Monte Carlo simulation results. The first important observation is that all the curves reach an error floor regardless of the ratio between P_S and P_R due to the residual LI. When P_R is selected as twice of P_S , the better outage performance can be achieved compared to the other curves in that scenario. Although there is a small gap between the lower-bound and simulation results, the lower-bound given

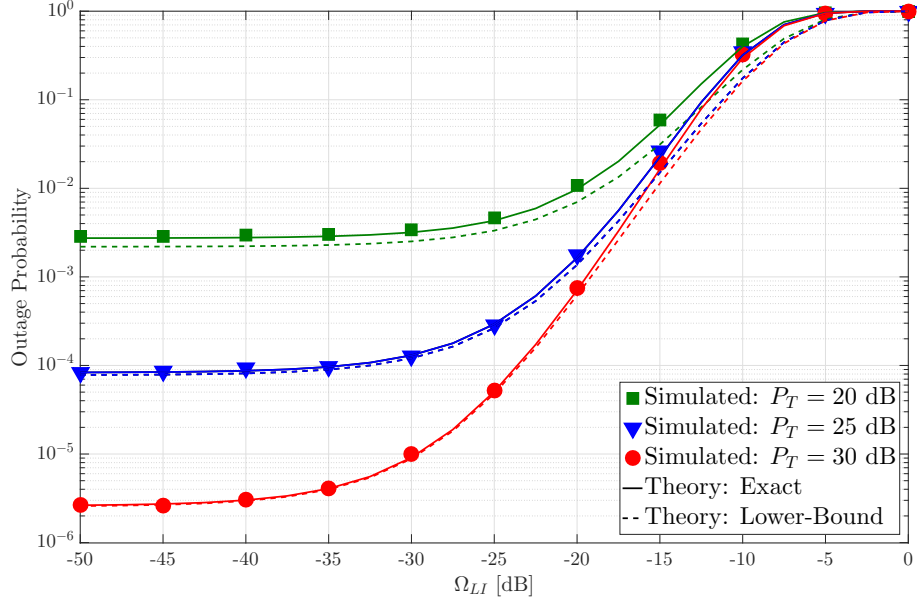


Figure 5.3 : Outage probability of the TWFD-AF systems at S_1 versus Ω_{LI} when $\gamma_{th}=2$, $P_S = P_R$, $m_{S_1R} = m_{S_2R} = 3$ and $m_R = m_{S_1} = 5$.

by (5.19) converges to the exact performance especially at medium and high SNR region. Furthermore, the asymptotic outage results obtained from (5.26) are perfectly consistent with the error floors.

Figure 5.3 represents the outage probability versus Ω_{LI} for $P_T \in \{20, 25, 30\}$ dB when $\gamma_{th} = 2$, $P_S = P_R$, $m_{S_1R} = m_{S_2R} = 3$ and $m_R = m_{S_1} = 5$. As seen from the curves, the outage probability of the TWFD-AF relay system takes lower values as long as either the transmission power is increased or the efficiency of the LI cancellation process is enhanced, i.e., the residual LI power Ω_{LI} is decreased. Additionally, the lower-bound results approach to the actual performance more tightly when the total transmission power increases.

5.2 Relay Selection for TWFD Variable-Gain AF Relay Systems over Nakagami- m Fading Channels

After analyzing the performance of the TWFD-AF systems, the relay selection algorithm can be applied in order to improve the outage performance and decrease the error floors. The outage probability of the proposed variable-gain two-way full-duplex amplify-and-forward relay selection (TWFD-AF-RS) systems is investigated over Nakagami- m fading channels. In this thesis, new exact outage probability expression of the TWFD-AF-RS systems with max-min relay selection policy is obtained in a

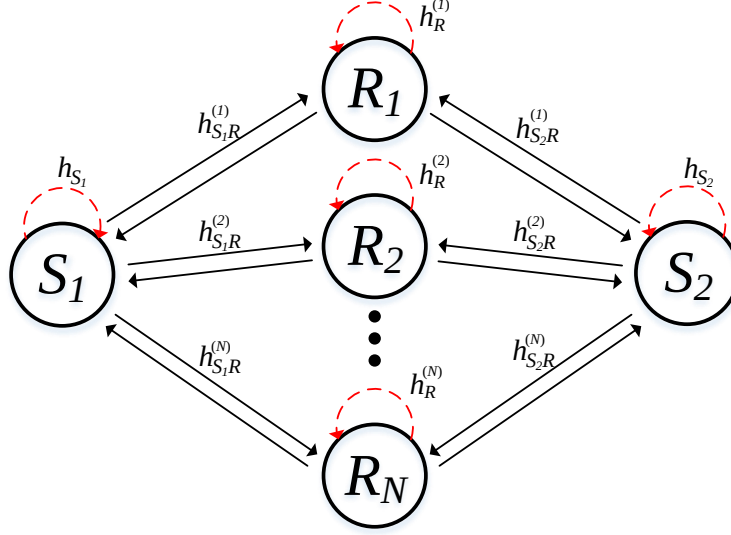


Figure 5.4 : TWFD-AF-RS system model.

single-integral form. As mentioned earlier, the performance of the TWFD-AF-RS systems was investigated in [32] for Rayleigh fading channels. In our work, besides more general fading model assumption (Nakagami- m) for all the source-relay channels, all the residual LI channels are also assumed as Nakagami- m fading, which is different from [32] where all the residual LI channels are modeled as complex Gaussian noise. Furthermore, we also derive tight lower-bound and asymptotic outage probability expressions in the closed-forms. The analytical results are verified by Monte Carlo type computer simulation results. The numerical results reveal that the relay selection in the TWFD-AF systems provides a significant advantage so as to obtain better outage performance and lower error-floors. It is also shown that the error floors can be determined easily by means of the derived asymptotic outage probability expression.

5.2.1 System model

We consider a multiple relay network topology as shown in Figure 5.4 consisting of two source nodes ($S_i, i \in \{1, 2\}$) and N relay nodes ($R_c, c \in \{1, 2, \dots, N\}$). In the TWFD-AF-RS system, each node is equipped with one transmitter and one receiver. The direct link between S_1 and S_2 is assumed strongly attenuated and they can only communicate with each other via the cooperative relays. During the two-way transmission, $N - 1$ relay nodes remain silent and transmission operation is

accomplished by using the selected relay node where the selection policy is described in the following section.

The complex channel coefficients between the $S_i - R_c$ links are denoted by $h_{S_i R}^{(c)}$ and they are reciprocal as in [32,34,35]. Likewise, h_{S_i} and $h_R^{(c)}$ are the complex residual LI channel coefficients at S_i and R_c , respectively. It is assumed that envelopes of all the channel coefficients $(|h_{S_i R}^{(c)}|, |h_{S_i}|, |h_R^{(c)}|)$ are Nakagami- m distributed with average powers of $\Omega_{S_i R}^{(c)}$, Ω_{S_i} and $\Omega_R^{(c)}$, respectively. For the simplicity, the average power of all links between S_i and R_c are considered to be equal and shown with $\Omega = \Omega_{S_i R}^{(c)}$. Similarly, the average power of residual LI channels are also assumed equal and given by $\Omega_{LI} = \Omega_{S_i} = \Omega_R^{(c)}$. Additionally, the complex Gaussian noise terms at receive antenna of each relay are assumed to be i.i.d. according to $\mathcal{CN}(0, \sigma_R^2)$.

By using (5.8), when two-way transmission is completed via R_c , the end-to-end SINR of the variable-gain TWFD-AF-RS systems becomes

$$\gamma_{end,i}^{(c)} = \frac{\alpha \frac{\gamma_{S_i R}^{(c)}}{(\gamma_{S_i} + 1)} \frac{\gamma_{S_j R}^{(c)}}{(\gamma_R^{(c)} + 1)}}{\alpha \frac{\gamma_{S_i R}^{(c)}}{\gamma_{S_i} + 1} + \frac{\gamma_{S_i R}^{(c)} + \gamma_{S_j R}^{(c)}}{\gamma_R^{(c)} + 1} + 1} \quad (5.27)$$

where $\gamma_{S_i R}^{(c)} = P_S |h_{S_i R}^{(c)}|^2 / \sigma_R^2$, $\gamma_{S_j R}^{(c)} = P_S |h_{S_j R}^{(c)}|^2 / \sigma_R^2$, $\gamma_R^{(c)} = P_R |h_R^{(c)}|^2 / \sigma_R^2$, $\gamma_{S_i} = P_S |h_{S_i}|^2 / \sigma_S^2$ and $\alpha = (P_R \sigma_R^2) / (P_S \sigma_S^2)$. It is also defined as $\bar{\gamma} = \bar{\gamma}_{S_i R}^{(c)} = \bar{\gamma}_{S_i R}^{(c)} = P_S \Omega / \sigma_R^2$ and $\bar{\gamma}_{LI} = \bar{\gamma}_R^{(c)} = \alpha \bar{\gamma}_{S_i} = P_R \Omega_{LI} / \sigma_R^2$. All channel coefficients are assumed to be Nakagami- m distributed. Hence, the PDF and CDF of $\gamma_{S_i R}^{(c)}$, $\gamma_{S_j R}^{(c)}$, $\gamma_R^{(c)}$ and γ_{S_i} can be respectively given as $f_A(x) = \frac{\lambda^\beta x^{\beta-1}}{\Gamma(\beta)} e^{-\lambda x}$ and $F_A(x) = \frac{\gamma(\beta, \lambda x)}{\Gamma(\beta)}$ with $(A, \beta, \lambda) \in \{(\gamma_{S_i R}^{(c)}, m, \mu), (\gamma_{S_j R}^{(c)}, m, \mu), (\gamma_R^{(c)}, m_{LI}, \mu_{LI}), (\gamma_{S_i}, m_{LI}, \alpha \mu_{LI})\}$, $\mu = m / \bar{\gamma}$ and $\mu_{LI} = m_{LI} / \bar{\gamma}_{LI}$. It is worthwhile to mention that $\gamma_{S_i R}^{(c)}$ and $\gamma_{S_j R}^{(c)}$ are i.i.d random variables.

5.2.2 Relay selection policy

The optimal relay selection algorithm for the two-way relay channels is to maximize the minimum of SINRs at S_1 and S_2 [32, 82]. Therefore, in order to improve the outage performance of the TWFD-AF-RS system, the max-min relay selection policy is constructed as follows

$$b = \arg \max_{1 \leq c \leq N} \min \left\{ \gamma_{S_1 R}^{(c)}, \gamma_{S_2 R}^{(c)} \right\}. \quad (5.28)$$

After selecting the b^{th} relay among N available relays considering the max-min selection algorithm, the end-to-end SINR given in (5.27) becomes

$$\gamma_{end,i}^{(b)} = \frac{\alpha \frac{\gamma_{S_i R}^{(b)}}{(\gamma_{S_i}^{(b)} + 1)} \frac{\gamma_{S_j R}^{(b)}}{(\gamma_R^{(b)} + 1)}}{\alpha \frac{\gamma_{S_i R}^{(b)}}{\gamma_{S_i}^{(b)} + 1} + \frac{\gamma_{S_i R}^{(b)} + \gamma_{S_j R}^{(b)}}{\gamma_R^{(b)} + 1} + 1} \quad (5.29)$$

where $\gamma_R^{(b)}$ and γ_{S_i} are independent from the max-min selection, so their distributions remain the same as given in the previous section. Although $\gamma_{S_i R}^{(b)}$ and $\gamma_{S_j R}^{(b)}$ are still i.i.d., they are not Gamma distributed random variables. The PDF and CDF expressions of i.i.d. random variables $\gamma_{S_i R}^{(b)}$ and $\gamma_{S_j R}^{(b)}$ should be derived before the outage probability analyses.

5.2.3 Outage probability analyses

The exact, lower-bound and asymptotic outage probability expressions of the TWFD-AF-RS systems are investigated in this section. In order to analyze the outage performance, we need both the CDF and PDF of $\gamma_{S_i R}^{(b)}$ according to the max-min relay selection algorithm. By using [32, Eq. (35)], the CDF of $\gamma_{S_i R}^{(b)}$ can be given by

$$\begin{aligned} F_{\gamma_{S_i R}^{(b)}}(z) &= \sum_{c=1}^N \int_{x=0}^z f_{\gamma_{S_i R}^{(c)}}(x) P(x \leq \gamma_{S_j R}^{(c)}) \prod_{\substack{p=1 \\ p \neq c}}^N P[\min(\gamma_{S_i R}^{(p)}, \gamma_{S_j R}^{(p)}) \leq x] dx \\ &+ \sum_{c=1}^N \int_{x=0}^z f_{\gamma_{S_i R}^{(c)}}(x) P(x > \gamma_{S_j R}^{(c)}) \prod_{\substack{p=1 \\ p \neq c}}^N P[\min(\gamma_{S_i R}^{(p)}, \gamma_{S_j R}^{(p)}) \leq \gamma_{S_j R}^{(c)}] dx \\ &= N \int_{x=0}^z f_{\gamma_{S_i R}^{(c)}}(x) \left[\bar{F}_{\gamma_{S_i R}^{(c)}}(x) \left(1 - \bar{F}_{\gamma_{S_i R}^{(c)}}^2(x) \right)^{N-1} \right. \\ &\quad \left. + \int_{y=0}^x f_{\gamma_{S_i R}^{(c)}}(y) \left(1 - \bar{F}_{\gamma_{S_i R}^{(c)}}^2(y) \right)^{N-1} dy \right] dx \end{aligned} \quad (5.30)$$

where $\bar{F}_{\gamma_{S_i R}^{(b)}}(\cdot) = 1 - F_{\gamma_{S_i R}^{(b)}}(\cdot)$. By substituting the PDF and CDF expressions of $\gamma_{S_i R}^{(c)}$ given in Section 5.2.1, the above expression is rewritten as

$$\begin{aligned} F_{\gamma_{S_i R}^{(b)}}(z) &= \frac{N\mu^m}{\Gamma(m)} \int_{x=0}^z \frac{x^{m-1} \Gamma(m, \mu x)}{e^{\mu x} \Gamma(m)} \left[1 - \left(\frac{\Gamma(m, \mu x)}{\Gamma(m)} \right)^2 \right]^{N-1} \\ &+ \frac{x^{m-1} \mu^m}{e^{\mu x} \Gamma(m)} \int_{y=0}^x \frac{y^{m-1}}{e^{\mu y}} \left[1 - \left(\frac{\Gamma(m, \mu y)}{\Gamma(m)} \right)^2 \right]^{N-1} dy dx. \end{aligned} \quad (5.31)$$

For the integer values of m , by using binomial expansion and [65, Eq. (8.352.4)],

$$F_{\gamma_{S_i R}^{(b)}}(z) = \frac{N\mu^m}{\Gamma(m)} \sum_{k=0}^{N-1} \binom{N-1}{k} (-1)^k \left\{ \underbrace{\int_{x=0}^z \frac{x^{m-1}}{e^{(2k+2)\mu x}} \left[\sum_{q=0}^{m-1} \frac{(\mu x)^q}{q!} \right]^{2k+1} dx}_{I_1} \right. \\ \left. + \underbrace{\frac{\mu^m}{\Gamma(m)} \int_{x=0}^z \frac{x^{m-1}}{e^{\mu x}} \int_{y=0}^x \frac{y^{m-1}}{e^{(2k+1)\mu y}} \left[\sum_{q=0}^{m-1} \frac{(\mu y)^q}{q!} \right]^{2k} dy dx}_{I_2} \right\} \quad (5.32)$$

can be obtained. For the solution of the first integration I_1 , we need to use multinomial coefficients as in [78, Eq. (7)] and [65, Eq. (3.351.1)]. Then, I_1 is found as

$$I_1 = \sum_{t=0}^{\theta_1} \Phi_t^{2k+1} \mu^t \int_{x=0}^z \frac{x^{t+m-1}}{e^{(2k+2)\mu x}} dx \\ = \sum_{t=0}^{\theta_1} \frac{\Phi_t^{2k+1} \gamma[t+m, (2k+2)\mu z]}{\mu^m (2k+2)^{t+m}} \quad (5.33)$$

where $\theta_1 = (2k+1)(m-1)$ and Φ_t^{2k+1} is the coefficient of $(\mu x)^t$ in the expansion of

$$\left[\sum_{q=0}^{m-1} \frac{(\mu x)^q}{q!} \right]^{2k+1}. \quad (5.34)$$

The other integration I_2 is composed of double-integrals. Initially, the inner integration inside I_2 can be found by following the same steps applied to I_1 and it is given by

$$I_2 = \sum_{t=0}^{\theta_2} \frac{\Phi_t^{2k} \mu^{-m}}{(2k+1)^{t+m}} \int_{x=0}^z \frac{\gamma[t+m, (2k+1)\mu x]}{x^{1-m} e^{\mu x}} dx \quad (5.35)$$

where $\theta_2 = 2k(m-1)$ and Φ_t^{2k} is also the coefficient of $(\mu y)^t$ in the expansion of

$$\left[\sum_{q=0}^{m-1} \frac{(\mu y)^q}{q!} \right]^{2k}. \quad (5.36)$$

By applying both [65, Eq. (8.352.6)] and [65, Eq. (3.351.1)], respectively, I_2 can be easily found in closed-form as

$$I_2 = \sum_{t=0}^{\theta_2} \frac{\Phi_t^{2k} \Gamma(m+t)}{\mu^{2m} (2k+1)^{t+m}} \\ \times \left[\gamma(m, \mu z) - \sum_{v=0}^{t+m-1} \frac{\gamma[m+v, (2k+2)\mu z]}{v! (2k+2)^{m+v} (2k+1)^{-v}} \right]. \quad (5.37)$$

Finally, by substituting (5.33) and (5.37) into (5.32), the CDF of $\gamma_{S_i R}^{(b)}$ can be obtained in a closed-form as follows

$$F_{\gamma_{S_i R}^{(b)}}(z) = \frac{N}{\Gamma(m)} \sum_{k=0}^{N-1} \binom{N-1}{k} (-1)^k \left\{ \sum_{t=0}^{\theta_1} \frac{\Phi_t^{2k+1} \gamma[t+m, (2k+2)\mu z]}{(2k+2)^{t+m}} \right. \\ \left. + \sum_{t=0}^{\theta_2} \frac{\Phi_t^{2k} \Gamma(t+m)}{\Gamma(m)(2k+1)^{t+m}} \left[\gamma(m, \mu z) - \sum_{v=0}^{t+m-1} \frac{\gamma[v+m, (2k+2)\mu z]}{v!(2k+2)^{v+m} (2k+1)^{-v}} \right] \right\}. \quad (5.38)$$

Then, it is convenient to derive the PDF of $\gamma_{S_i R}^{(b)}$ with the derivative of its CDF. By using $\frac{\partial}{\partial z} \gamma(m, \mu z) = e^{-\mu z} \mu^m z^{m-1}$ [65, Eq. (8.356.4)] and [65, Eq. (8.352.6)], the PDF expression is obtained as

$$f_{\gamma_{S_i R}^{(b)}}(z) = \frac{\partial}{\partial z} F_{\gamma_{S_i R}^{(b)}}(z) \\ = \frac{N \mu^m z^{m-1}}{\Gamma(m) e^{\mu z}} \sum_{k=0}^{N-1} \binom{N-1}{k} (-1)^k \left\{ \sum_{t=0}^{\theta_1} \frac{\Phi_t^{2k+1}}{\mu^{-t}} \right. \\ \left. \times \frac{z^t}{e^{(2k+1)\mu z}} + \sum_{t=0}^{\theta_2} \frac{\Phi_t^{2k} \gamma[t+m, (2k+1)\mu z]}{\Gamma(m) (2k+1)^{t+m}} \right\}. \quad (5.39)$$

The exact outage probability of the TWFD-AF-RS system at node S_i with max-min relay selection algorithm is defined as $P_{out,i}(\gamma_{th}) = F_{\gamma_{end,i}^{(b)}}(\gamma_{th})$ where $F_{\gamma_{end,i}^{(b)}}(\cdot)$ is the CDF of end-to-end SINR given in (5.29). For the sake of simplicity, +1 term at the denominator of end-to-end SINR is neglected and it can be reorganized as

$$\gamma_{end,i}^{(b)} \approx \frac{\alpha \frac{\gamma_{S_i R}^{(b)}}{(\gamma_{S_i}+1)} \frac{\gamma_{S_j R}^{(b)}}{(\gamma_R^{(b)}+1)}}{\alpha \frac{\gamma_{S_i R}^{(b)}}{\gamma_{S_i}+1} + \frac{\gamma_{S_i R}^{(b)} + \gamma_{S_j R}^{(b)}}{\gamma_R^{(b)}+1}} = \frac{\alpha \frac{\gamma_{S_i R}^{(b)} \gamma_{S_j R}^{(b)}}{\gamma_{S_i}+1}}{\alpha \frac{\gamma_{S_i R}^{(b)} [\gamma_R^{(b)}+1+(\gamma_{S_i}+1)/\alpha]}{\gamma_{S_i}+1} + \gamma_{S_j R}^{(b)}} \\ = \frac{Y_i^{(b)} Z_i^{(b)}}{Y_i^{(b)} + Z_i^{(b)}} \quad (5.40)$$

where $Y_i^{(b)} = \alpha \frac{\gamma_{S_i R}^{(b)}}{\gamma_{S_i}+1}$ and $Z_i^{(b)} = \frac{\gamma_{S_j R}^{(b)}}{\gamma_R^{(b)}+1+(\gamma_{S_i}+1)/\alpha}$. As in Section 5.1.2, we again assume that $Y_i^{(b)}$ and $Z_i^{(b)}$ are independent from each other due to the dominant effect of $\gamma_{S_i R}^{(b)}$ and $\gamma_{S_j R}^{(b)}$ with respect to the residual LI terms. By using (5.10), the CDF of end-to-end SINR can be written as

$$F_{\gamma_{end,i}^{(b)}}(\gamma_{th}) = 1 - \int_{y=0}^{\infty} \bar{F}_{Z_i^{(b)}} \left(x + \frac{x^2}{y} \right) f_{Y_i^{(b)}}(y+x) dy \quad (5.41)$$

where $\bar{F}_{Z_i^{(b)}}(\cdot) = 1 - F_{Z_i^{(b)}}(\cdot)$. The PDF of $Y_i^{(b)}$ is given by

$$\begin{aligned}
f_{Y_i^{(b)}}(y) &= \int_{x=0}^{\infty} \left(\frac{x+1}{\alpha} \right) f_{\gamma_{S_i R}^{(b)}} \left[\frac{(x+1)y}{\alpha} \right] f_{\gamma_{S_i}}(x) dx \\
&= \sum_{k=0}^{N-1} \binom{N-1}{k} \frac{N(\alpha\mu_{LI})^{m_{LI}} (-1)^k}{y\Gamma(m)\Gamma(m_{LI})} \left\{ \sum_{t=0}^{\theta_1} \frac{\Phi_t^{2k+1} \left(\frac{\mu y}{\alpha} \right)^{t+m}}{e^{(2k+2)\mu y/\alpha}} \right. \\
&\quad \times \underbrace{\int_{x=0}^{\infty} \frac{(x+1)^{t+m} x^{m_{LI}-1}}{e^{x(\alpha\mu_{LI} + (2k+2)\mu y/\alpha)}} dx}_{I_3} + \sum_{t=0}^{\theta_2} \frac{\Phi_t^{2k} \left(\frac{\mu y}{\alpha} \right)^m e^{-\mu y/\alpha}}{\Gamma(m)(2k+1)^{t+m}} \\
&\quad \times \underbrace{\int_{x=0}^{\infty} \frac{\gamma[t+m, (2k+1)(x+1)\mu y/\alpha]}{x^{1-m_{LI}} (x+1)^{-m} e^{x(\alpha\mu_{LI} + \mu y/\alpha)}} dx}_{I_4} \Bigg\}. \tag{5.42}
\end{aligned}$$

As seen from above expression, we need to derive two independent integrals denoted as I_3 and I_4 . By applying [65, Eq. (9.211.4)], I_3 can be easily obtained as

$$I_3 = \frac{U\left(m_{LI}, m_{LI} + m + t + 1, \alpha\mu_{LI} + \frac{(2k+2)\mu y}{\alpha}\right)}{[\Gamma(m_{LI})]^{-1}}. \tag{5.43}$$

By using the expansion of lower incomplete Gamma function [65, Eq. (8.352.6)], other integration term I_4 is also found as

$$\begin{aligned}
I_4 &= \Gamma(t+m) \left\{ \frac{U\left(m_{LI}, m_{LI} + m + 1, \alpha\mu_{LI} + \frac{\mu y}{\alpha}\right)}{[\Gamma(m_{LI})]^{-1}} - \sum_{v=0}^{t+m-1} \frac{[(2k+1)\mu y/\alpha]^v}{e^{(2k+1)\mu y/\alpha} v!} \right. \\
&\quad \times \left. \frac{U\left(m_{LI}, m_{LI} + m + v + 1, \alpha\mu_{LI} + \frac{(2k+2)\mu y}{\alpha}\right)}{[\Gamma(m_{LI})]^{-1}} \right\}. \tag{5.44}
\end{aligned}$$

Finally, the PDF of $Y_i^{(b)}$ is obtained by substituting (5.43) and (5.44) into (5.42) and it is given by

$$\begin{aligned}
f_{Y_i^{(b)}}(y) &= \sum_{k=0}^{N-1} \binom{N-1}{k} (-1)^k \left\{ \sum_{t=0}^{\theta_1} \Delta_{Y,1}(y; \Phi_t^{2k+1}, t, 2k+2) \right. \\
&\quad \left. + \sum_{t=0}^{\theta_2} \frac{\Gamma(t+m)}{\Gamma(m)} \left[\frac{\Delta_{Y,1}(y; \Phi_t^{2k}, 0, 1)}{(2k+1)^{t+m}} - \sum_{v=0}^{t+m-1} \frac{\Delta_{Y,1}(y; \Phi_t^{2k}, v, 2k+2)}{(2k+1)^{t+m-v} v!} \right] \right\} \tag{5.45}
\end{aligned}$$

where

$$\Delta_{Y,1}(y; \Phi, \delta, \varepsilon) = \frac{N\Phi y^{m+\delta-1} \mu^{m+\delta} \mu_{LI}^{m_{LI}}}{\Gamma(m) e^{\varepsilon\mu y/\alpha} \alpha^{m+\delta-m_{LI}}} U\left(m_{LI}, m_{LI} + m + \delta + 1, \alpha\mu_{LI} + \frac{\varepsilon\mu y}{\alpha}\right). \tag{5.46}$$

According to (5.41), the CDF of $Z_i^{(b)}$ is also required for the calculation of the exact outage probability and given by

$$\begin{aligned}
F_{Z_i^{(b)}}(z) &= \int_{x=0}^{\infty} \int_{y=0}^{\infty} f_{\gamma_R^{(b)}}(x) f_{\gamma_{S_i}}(y) F_{\gamma_{S_j R}^{(b)}} \left[\left(x + 1 + \frac{y+1}{\alpha} \right) z \right] dy dx \\
&= \sum_{k=0}^{N-1} \binom{N-1}{k} (-1)^k \left\{ \sum_{t=0}^{\theta_1} \frac{\Delta_Z(z; \Phi_t^{2k+1}, t, 2k+2)}{(2k+2)^{t+m}} \right. \\
&\quad + \sum_{t=0}^{\theta_2} \frac{\Gamma(t+m)}{\Gamma(m)} \left[\frac{\Delta_Z(z; \Phi_t^{2k}, 0, 1)}{(2k+1)^{t+m}} \right. \\
&\quad \left. \left. - \sum_{v=0}^{t+m-1} \frac{\Delta_Z(z; \Phi_t^{2k}, v, 2k+2)}{v!(2k+1)^{t+m-v}(2k+2)^{v+m}} \right] \right\}
\end{aligned} \tag{5.47}$$

where $\Delta_Z(\cdot; \cdot, \cdot, \cdot, \cdot)$ containing double-integrals is written as

$$\Delta_Z(z; \Phi, \delta, \varepsilon) = \frac{N\Phi\alpha^{m_{LI}}\mu_{LI}^{2m_{LI}}}{[\Gamma(m_{LI})]^2\Gamma(m)} \int_{x=0}^{\infty} \int_{y=0}^{\infty} \frac{\gamma[m+\delta, (x+1+\frac{y+1}{\alpha})\varepsilon\mu z]}{(xy)^{1-m_{LI}} e^{x\mu_{LI}+\alpha y\mu_{LI}}} dy dx. \tag{5.48}$$

Applying [65, Eq. (8.352.6)], [65, Eq. (3.351.3)] and the binomial expansion, respectively, $\Delta_Z(\cdot; \cdot, \cdot, \cdot, \cdot)$ can be rewritten as

$$\begin{aligned}
\Delta_Z(z; \Phi, \delta, \varepsilon) &= \frac{N\Phi\Gamma(\delta+m)}{\Gamma(m)} \left[1 - \sum_{u=0}^{\delta+m-1} \sum_{f=0}^u \frac{\alpha^{m_{LI}-f} \mu_{LI}^{2m_{LI}} (\varepsilon\mu z)^u}{[\Gamma(m_{LI})]^2 (u-f)! f! e^{(1+\alpha^{-1})\varepsilon\mu z}} \right. \\
&\quad \left. \times \int_{x=0}^{\infty} \int_{y=0}^{\infty} \frac{(xy)^{m_{LI}-1} (x+1)^{u-f} (y+1)^f}{e^{x(\mu_{LI}+\varepsilon\mu z)+y(\alpha\mu_{LI}+\varepsilon\mu z/\alpha)}} dy dx \right].
\end{aligned} \tag{5.49}$$

One can easily show that the double-integrals in above expression consist of the multiplication of independent single-integrals and both of the single-integrals are quite similar to each other. Hence, by using [65, Eq. (9.211.4)],

$$\begin{aligned}
\Delta_Z(z; \Phi, \delta, \varepsilon) &= \frac{N\Phi\Gamma(\delta+m)}{\Gamma(m)} \left[1 - \sum_{u=0}^{\delta+m-1} \sum_{f=0}^u \frac{U(m_{LI}, u-f+m_{LI}+1, \mu_{LI}+\varepsilon\mu z)}{\alpha^{f-m_{LI}} \mu_{LI}^{-2m_{LI}} (\varepsilon\mu z)^{-u}} \right. \\
&\quad \left. \times \frac{U(m_{LI}, f+m_{LI}+1, \alpha\mu_{LI}+\varepsilon\mu z/\alpha)}{(u-f)! f! e^{(1+\alpha^{-1})\varepsilon\mu z}} \right]
\end{aligned} \tag{5.50}$$

can be obtained. Now, it is convenient to derive the CDF of $Z_i^{(b)}$ in a closed-form by only substituting (5.50) into (5.47).

Finally, the exact outage probability expression of the TWFD-AF-RS system can be obtained in a single-integral form by combining (5.41), (5.45) and (5.47) as

$$\begin{aligned}
P_{out,i}(\gamma_{th}) = & 1 - \int_{y=0}^{\infty} \left\{ \sum_{k=0}^{N-1} \binom{N-1}{k} (-1)^k \left[\sum_{t=0}^{\theta_1} \Delta_{Y,1}(y + \gamma_{th}; \Phi_t^{2k+1}, t, 2k+2) \right. \right. \\
& + \sum_{t=0}^{\theta_2} \frac{\Gamma(t+m)}{\Gamma(m)} \left(\frac{\Delta_{Y,1}(y + \gamma_{th}; \Phi_t^{2k}, 0, 1)}{(2k+1)^{t+m}} - \sum_{v=0}^{t+m-1} \frac{\Delta_{Y,1}(y + \gamma_{th}; \Phi_t^{2k}, v, 2k+2)}{(2k+1)^{t+m-v} v!} \right) \Bigg] \Bigg\} \\
& \times \left\{ 1 - \sum_{k=0}^{N-1} \binom{N-1}{k} (-1)^k \left[\sum_{t=0}^{\theta_1} \frac{\Delta_Z(\gamma_{th} + \frac{\gamma_{th}^2}{y}; \Phi_t^{2k+1}, t, 2k+2)}{(2k+2)^{t+m}} + \sum_{t=0}^{\theta_2} \frac{\Gamma(t+m)}{\Gamma(m)} \right. \right. \\
& \times \left. \left(\frac{\Delta_Z(\gamma_{th} + \frac{\gamma_{th}^2}{y}; \Phi_t^{2k}, 0, 1)}{(2k+1)^{t+m}} - \sum_{v=0}^{t+m-1} \frac{\Delta_Z(\gamma_{th} + \frac{\gamma_{th}^2}{y}; \Phi_t^{2k}, v, 2k+2)}{v! (2k+1)^{t+m-v} (2k+2)^{v+m}} \right) \right] \Bigg\} dy.
\end{aligned} \tag{5.51}$$

Unfortunately, to the best of our knowledge, $P_{out,i}(\cdot)$ does not have a closed-form solution. But, it can be efficiently calculated by using common mathematical softwares such as MATLAB, MATHEMATICA or MAPLE.

The end-to-end SINR of the TWFD-AF-RS system given in (5.29) is upper-bounded as $\gamma_{end,i,b} \approx \frac{Y_{i,b} Z_{i,b}}{Y_{i,b} + Z_{i,b}} \leq \min(Y_{i,b}, Z_{i,b})$. The upper-bound for the end-to-end SINR implies the lower-bound for the outage probability as follows

$$\begin{aligned}
P_{out,low,i}(\gamma_{th}) &= P \left[\min(Y_{S_{iR}}^{(b)}, Z_{S_{iR}}^{(b)}) \leq \gamma_{th} \right] \\
&= 1 - \bar{F}_{Y_{S_{iR}}^{(b)}}(\gamma_{th}) \bar{F}_{Z_{S_{iR}}^{(b)}}(\gamma_{th})
\end{aligned} \tag{5.52}$$

where $\bar{F}_{Y_{S_{iR}}^{(b)}}(\cdot) = 1 - F_{Y_{S_{iR}}^{(b)}}(\cdot)$. We have already found the CDF of $Z_{S_{iR}}^{(b)}$ given in (5.47). Therefore, it is required to find the CDF of $Y_{S_{iR}}^{(b)}$ from its PDF in (5.45) as $F_{Y_{S_{iR}}^{(b)}}(y) = \int_{q=0}^y f_{Y_{S_{iR}}^{(b)}}(q) dq$. The only term inside $f_{Y_{i,b}}(\cdot)$ belongs to the integration is $\Delta_{Y,1}(\cdot; \cdot, \cdot, \cdot)$ given in (5.46). In order to simplify the analysis, we only evaluate

$$\begin{aligned}
\Delta_{Y,2}(y; \Phi, \delta, \varepsilon) &= \frac{N \Phi \mu^{m+\delta} \mu_{LI}^{m_{LI}}}{\Gamma(m) \alpha^{m+\delta-m_{LI}}} \int_{q=0}^y \frac{q^{m+\delta-1}}{e^{\varepsilon \mu q / \alpha}} \\
&\times U\left(m_{LI}, m_{LI} + m + \delta + 1, \alpha \mu_{LI} + \frac{\varepsilon \mu q}{\alpha}\right) dq
\end{aligned} \tag{5.53}$$

where $\Delta_{Y,2}(\cdot; \cdot, \cdot, \cdot)$ describes the integration of $\Delta_{Y,1}(\cdot; \cdot, \cdot, \cdot)$. Then, expanding the confluent hypergeometric function as an integration [65, Eq. (9.211.4)], it can be rewritten as

$$\begin{aligned}\Delta_{Y,2}(y; \Phi, \delta, \varepsilon) &= \frac{N\Phi\mu^{m+\delta}\mu_{LI}^{m_{LI}}}{\Gamma(m)\Gamma(m_{LI})\alpha^{m+\delta-m_{LI}}} \\ &\times \int_{q=0}^y \int_{x=0}^{\infty} \frac{q^{m+\delta-1}x^{m_{LI}-1}(1+x)^{m+\delta}}{e^{x(\alpha\mu_{LI}+\varepsilon\mu q/\alpha)+\varepsilon\mu q/\alpha}} dx dq.\end{aligned}\quad (5.54)$$

Before starting to the solution of the double-integration, we should change the order of integrals as follows

$$\begin{aligned}\Delta_{Y,2}(y; \Phi, \delta, \varepsilon) &= \frac{N\Phi\mu^{m+\delta}\mu_{LI}^{m_{LI}}}{\Gamma(m)\Gamma(m_{LI})\alpha^{m+\delta-m_{LI}}} \\ &\times \int_{x=0}^{\infty} \frac{x^{m_{LI}-1}(1+x)^{m+\delta}}{e^{\alpha x\mu_{LI}}} \int_{q=0}^y \frac{q^{m+\delta-1}}{e^{(1+x)\varepsilon\mu q/\alpha}} dq dx.\end{aligned}\quad (5.55)$$

The inner integration can be easily solved as

$$\begin{aligned}\Delta_{Y,2}(y; \Phi, \delta, \varepsilon) &= \frac{N\Phi(\alpha\mu_{LI})^{m_{LI}}}{\Gamma(m)\Gamma(m_{LI})\varepsilon^{m+\delta}} \\ &\times \int_{x=0}^{\infty} \frac{x^{m_{LI}-1}\gamma[m+\delta, (1+x)\varepsilon\mu y/\alpha]}{e^{\alpha x\mu_{LI}}} dx.\end{aligned}\quad (5.56)$$

Afterwards, by again using [65, Eq. (8.352.6)] and [65, Eq. (9.211.4)], respectively, it is obtained as

$$\begin{aligned}\Delta_{Y,2}(y; \Phi, \delta, \varepsilon) &= \frac{N\Phi\Gamma(m+\delta)}{\Gamma(m)\varepsilon^{m+\delta}} \left[1 - \frac{(\alpha\mu_{LI})^{m_{LI}}}{e^{\varepsilon\mu y/\alpha}} \right. \\ &\times \left. \sum_{u=0}^{m+\delta-1} \frac{U(m_{LI}, m_{LI}+u+1, \alpha\mu_{LI} + \frac{\varepsilon\mu y}{\alpha})}{(\varepsilon\mu y/\alpha)^{-u}u!} \right].\end{aligned}\quad (5.57)$$

At the end, the CDF of $Y_{S_iR}^{(b)}$ can be easily written as

$$\begin{aligned}F_{Y_{S_iR}^{(b)}}(y) &= \sum_{k=0}^{N-1} \binom{N-1}{k} (-1)^k \left\{ \sum_{t=0}^{\theta_1} \Delta_{Y,2}(y; \Phi_t^{2k+1}, t, 2k+2) + \sum_{t=0}^{\theta_2} \frac{\Gamma(t+m)}{\Gamma(m)} \right. \\ &\times \left. \left[\frac{\Delta_{Y,2}(y; \Phi_t^{2k}, 0, 1)}{(2k+1)^{t+m}} - \sum_{v=0}^{t+m-1} \frac{\Delta_{Y,2}(y; \Phi_t^{2k}, v, 2k+2)}{(2k+1)^{t+m-v}v!} \right] \right\}.\end{aligned}\quad (5.58)$$

By substituting (5.47) and (5.58) into (5.52), the lower-bound outage probability $P_{out,low,i}(\cdot)$ is derived in a closed-form.

Another significant performance metric, which is asymptotic outage probability of the TWFD-AF-RS system, is also analyzed for the high SNR region. In this analysis, we again assume that $\{P_S, P_R\} \rightarrow \infty$ for fixed σ_S^2 and σ_R^2 . By using (5.20), the asymptotic outage probability is given by

$$P_{out,i}^{\infty}(\gamma_{th}) = 1 - \bar{F}_{Y_{i,\infty}^{(b)}}(\gamma_{th}) \bar{F}_{Y_{i,\infty}^{(b)}}(\gamma_{th}) \quad (5.59)$$

where $Y_{i,\infty}^{(c)} = \lim_{P_S, P_R \rightarrow \infty} Y_i^{(c)}$ and $Z_{i,\infty}^{(c)} = \lim_{P_S, P_R \rightarrow \infty} Z_i^{(c)}$. Therefore, their CDF expressions are briefly defined as $F_{Y_{i,\infty}^{(c)}}(\cdot) = \lim_{P_S, P_R \rightarrow \infty} F_{Y_i^{(b)}}(\cdot)$ and $F_{Z_{i,\infty}^{(c)}}(\cdot) = \lim_{P_S, P_R \rightarrow \infty} F_{Z_i^{(b)}}(\cdot)$. In order to compute the asymptotic outage probability, we need to obtain new CDF expressions for the limit condition. Actually, in the limit condition, $\{P_S, P_R\} \rightarrow \infty$ means that $\{\mu, \mu_{LI}\} \rightarrow 0$. As seen from the CDF of $Y_i^{(b)}$ in (5.58), the only term related to the limit is $\Delta_{Y,2}(\cdot; \cdot, \cdot, \cdot)$ that is defined in (5.57). At the high SNR region, $\Delta_{Y,2}(\cdot; \cdot, \cdot, \cdot)$ is given by

$$\Delta_{Y,2}^\infty(y; \Phi, \delta, \varepsilon) = \frac{N\Phi\Gamma(m+\delta)}{\Gamma(m)\varepsilon^{m+\delta}} \left\{ 1 - \sum_{u=0}^{m+\delta-1} \frac{\alpha^{m_{LI}}(\varepsilon\zeta y)^u}{u!} \right. \\ \left. \times \left[\lim_{\substack{P_S \rightarrow \infty \\ P_R \rightarrow \infty}} \frac{U(m_{LI}, m_{LI} + u + 1, \alpha\mu_{LI} + \frac{\varepsilon\mu y}{\alpha})}{\mu_{LI}^{-m_{LI}-u} e^{\varepsilon\mu y/\alpha}} \right] \right\} \quad (5.60)$$

where $\zeta = \frac{\mu}{\alpha\mu_{LI}} = \frac{\sigma_S^2\Omega_{LI}m}{\sigma_R^2\Omega_{LI}m_{LI}}$. In order to handle the limit operation, we need to transform the confluent hypergeometric function to a single-integral [65, Eq. (9.211.4)]. It is given by

$$\Delta_{Y,2}^\infty(y; \Phi, \delta, \varepsilon) = \frac{N\Phi\Gamma(m+\delta)}{\Gamma(m)\varepsilon^{m+\delta}} \left\{ 1 - \sum_{u=0}^{m+\delta-1} \frac{\alpha^{m_{LI}}(\varepsilon\zeta y)^u}{u!} \right. \\ \left. \times \left[\lim_{\substack{P_S \rightarrow \infty \\ P_R \rightarrow \infty}} \frac{\mu_{LI}}{e^{\varepsilon\mu y/\alpha}} \int_{x=0}^{\infty} \frac{(x\mu_{LI})^{m_{LI}-1}(\mu_{LI} + x\mu_{LI})^u}{e^{x(\alpha\mu_{LI} + \varepsilon\mu y/\alpha)}} dx \right] \right\}. \quad (5.61)$$

When we apply the variable transformation as $q = x\mu_{LI}$ inside the integration and use $\lim_{x \rightarrow 0} \frac{(x+q)^t}{e^x} = q^t$, the limit of the above expression can be easily computed. By using [65, Eq. (3.351.3)], it can be found as

$$\Delta_{Y,2}^\infty(y; \Phi, \delta, \varepsilon) = \frac{N\Phi\Gamma(m+\delta)}{\Gamma(m)\varepsilon^{m+\delta}} \left[1 - \sum_{u=0}^{m+\delta-1} \frac{\alpha^{m_{LI}}(\varepsilon\zeta y)^u}{\Gamma(m_{LI})u!} \frac{\Gamma(m_{LI}+u)}{(\alpha + \varepsilon\zeta y)^{m_{LI}+u}} \right]. \quad (5.62)$$

By simply using the above expression and (5.58), the CDF of $Y_{i,\infty}^{(c)}$ is obtained as

$$F_{Y_{i,\infty}^{(c)}}(y) = \sum_{k=0}^{N-1} \binom{N-1}{k} (-1)^k \left\{ \sum_{t=0}^{\theta_1} \Delta_{Y,2}^\infty(y; \Phi_t^{2k+1}, t, 2k+2) + \sum_{t=0}^{\theta_2} \frac{\Gamma(t+m)}{\Gamma(m)} \right. \\ \left. \times \left[\frac{\Delta_{Y,2}^\infty(y; \Phi_t^{2k}, 0, 1)}{(2k+1)^{t+m}} - \sum_{u=0}^{t+m-1} \frac{\Delta_{Y,2}^\infty(y; \Phi_t^{2k}, u, 2k+2)}{(2k+1)^{t+m-u} u!} \right] \right\}. \quad (5.63)$$

Afterwards, similar steps have to be accomplished so as to derive the CDF of $Z_{i,\infty}^{(c)}$. If we analyze the CDF of $Z_i^{(c)}$ given in (5.47), it is convenient to realize that the only term

relevant to the limit operator is $\Delta_Z(\cdot; \cdot, \cdot, \cdot)$ expressed in (5.50). When $\{P_S, P_R\} \rightarrow \infty$, $\Delta_Z(\cdot; \cdot, \cdot, \cdot)$ will be

$$\begin{aligned} \Delta_Z^\infty(z; \Phi, \delta, \varepsilon) &= \frac{N\Phi\Gamma(\delta+m)}{\Gamma(m)} \left\{ 1 - \sum_{u=0}^{\delta+m-1} \sum_{f=0}^u \frac{\alpha^{m_{LI}+u-f}}{(u-f)!} \right. \\ &\quad \times \frac{(\varepsilon\eta z)^u}{f!} \left[\lim_{\substack{P_S \rightarrow \infty \\ P_R \rightarrow \infty}} \frac{U(m_{LI}, u-f+m_{LI}+1, \mu_{LI}+\varepsilon\mu z)}{\mu_{LI}^{-2m_{LI}-u}} \right. \\ &\quad \left. \left. \times \frac{U(m_{LI}, f+m_{LI}+1, \alpha\mu_{LI} + \frac{\varepsilon\mu z}{\alpha})}{e^{(1+\alpha^{-1})\varepsilon\mu z}} \right] \right\}. \end{aligned} \quad (5.64)$$

As mentioned above, we again need to expand the hypergeometric functions as integrations, perform variable transformation and use limit property. After some manipulations, $\Delta_Z^\infty(\cdot; \cdot, \cdot, \cdot)$ is obtained as

$$\begin{aligned} \Delta_Z^\infty(z; \Phi, \delta, \varepsilon) &= \frac{N\Phi\Gamma(\delta+m)}{\Gamma(m)} \left\{ 1 - \sum_{u=0}^{\delta+m-1} \sum_{f=0}^u \frac{\alpha^{m_{LI}+u-f}}{(u-f)!f!} \right. \\ &\quad \left. \times \frac{(\varepsilon\eta z)^u \Gamma(m_{LI}+u-f) \Gamma(m_{LI}+f)}{[\Gamma(m_{LI})]^2 (1+\alpha\varepsilon\eta z)^{m_{LI}+u-f} (\alpha+\varepsilon\eta z)^{m_{LI}+f}} \right\}. \end{aligned} \quad (5.65)$$

Utilizing both (5.47) and (5.65), the CDF of $Z_{i,\infty}^{(c)}$ is written as

$$\begin{aligned} F_{Z_{i,b}^\infty}(z) &= \sum_{k=0}^{N-1} \binom{N-1}{k} (-1)^k \left\{ \sum_{t=0}^{\theta_1} \frac{\Delta_Z^\infty(z; \Phi_t^{2k+1}, t, 2k+2)}{(2k+2)^{t+m}} + \sum_{t=0}^{\theta_2} \frac{\Gamma(t+m)}{\Gamma(m)} \right. \\ &\quad \left. \times \left[\frac{\Delta_Z^\infty(z; \Phi_t^{2k}, 0, 1)}{(2k+1)^{t+m}} - \sum_{v=0}^{t+m-1} \frac{\Delta_Z^\infty(z; \Phi_t^{2k}, v, 2k+2)}{v!(2k+1)^{t+m-v} (2k+2)^{v+m}} \right] \right\}. \end{aligned} \quad (5.66)$$

Eventually, it is possible to derive $P_{out,i}^\infty(\cdot)$ in a closed-form by substituting (5.63) and (5.66) into (5.59). For the further investigation, it is observed that the CDF of both $Y_{i,\infty}^{(c)}$ and $Z_{i,\infty}^{(c)}$, in terms of transmission powers, only depend on the ratio between P_S and P_R . Hence, TWFD-AF-RS exhibits zero-diversity order at the high SNR region.

5.2.4 Numerical results

In this section, we provide Monte Carlo simulation results in order to evaluate outage performance of the TWFD-AF-RS systems. Accuracy of the theoretical analyses are verified via different scenarios. In the TWFD-AF-RS systems, the total transmission power of one active relay and two sources is denoted by $P_T = 2P_S + P_R$. For all the simulation scenarios, we assume that $\Omega = 1$ and $\sigma_S^2 = \sigma_R^2 = 1$ and the outage probability curves are produced according to S_1 node, i.e., $P_{out,1}(\cdot)$.

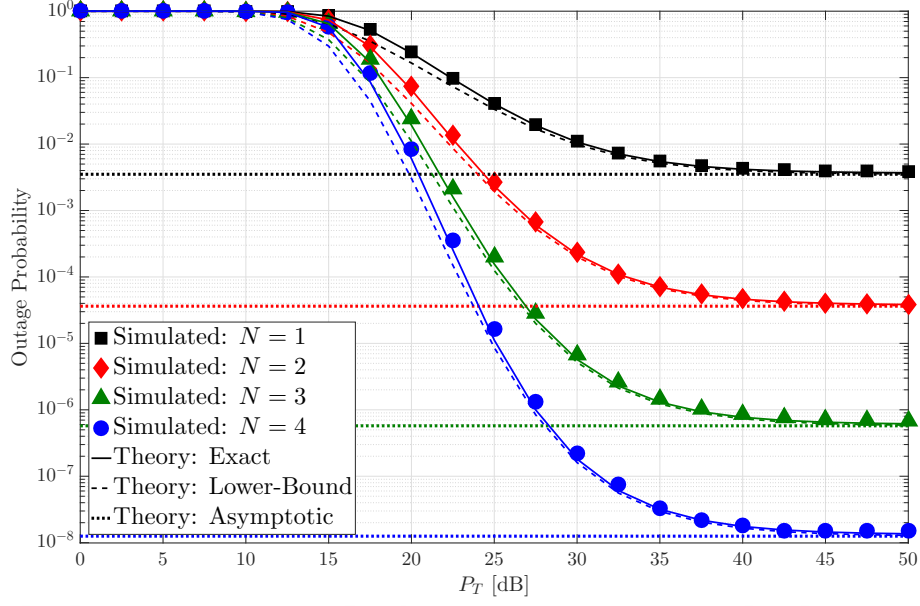


Figure 5.5 : Outage probability of the TWFD-AF-RS systems at S_1 versus P_T when $2P_S = P_R$, $\gamma_{th} = 5$, $\Omega_{LI} = -25$ dB, $m = 2$ and $m_{LI} = 4.5$.

In Figure 5.5, outage performance of the TWFD-AF-RS systems is plotted versus different values of total transmission power P_T when $2P_S = P_R$, $\gamma_{th} = 5$, $\Omega_{LI} = -25$ dB, $m = 2$ and $m_{LI} = 4.5$ for $N \in \{1, 2, 3, 4\}$. It is shown that the theoretical curves for the exact outage probability obtained from (5.51) are perfectly consistent with the Monte Carlo simulation results, verifying the accuracy of our analyses. As seen from the curves, outage probability of the TWFD-AF-RS systems can be enhanced by increasing the number of available relays for the max-min selection policy. Although slope of the outage curves become more steep thanks to the rise of N , error floors appeared due to the residual LI cannot be prevented. In terms of the error floor characteristic, the increase of N provides a significant decrease. For example, outage performance of the TWFD-AF-RS system for $N \in \{1, 2, 3, 4\}$ is limited at 3.7×10^{-3} , 3.8×10^{-5} , 6.6×10^{-7} and 1.5×10^{-8} , respectively, in this scenario. On the other hand, regardless of the total number of relays present in the system, the error floors are observed in the same P_T interval, i.e. every curve reach its error floor, when the total transmission power is approximately between 45 dB and 50 dB. Moreover, the exact outage probability curves are tightly lower-bounded with (5.52) especially in the medium and high SNR regions. Additionally, the error floors appeared at the higher values of P_T are fully compatible with the asymptotic outage probability results generated from (5.59). Therefore, the error floors can be easily determined by using the derived asymptotic outage probability expression.

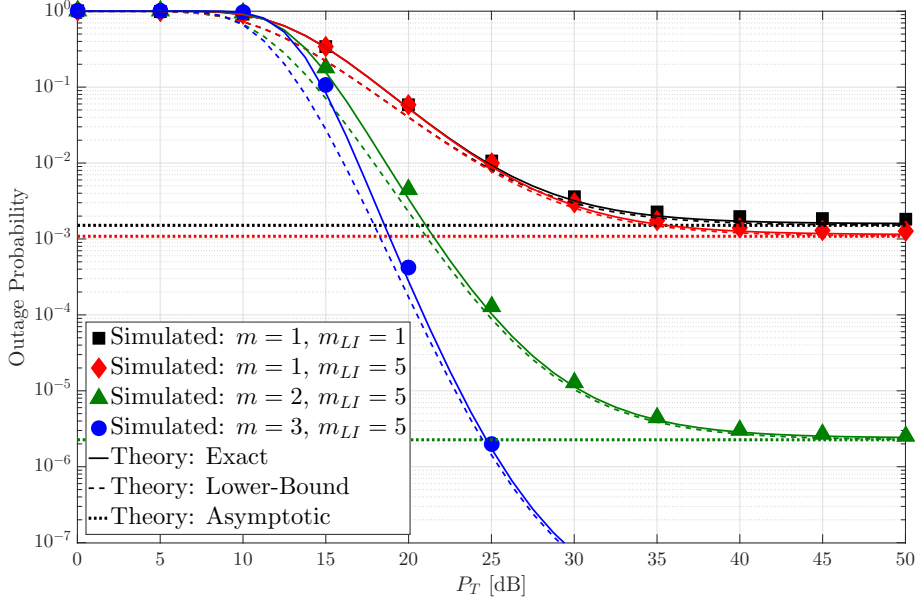


Figure 5.6 : Outage probability of the TWFD-AF-RS systems at S_1 versus P_T when $N = 2$, $P_S = P_R$, $\gamma_{th} = 2$ and $\Omega_{LI} = -23$ dB.

Figure 5.6 illustrates the outage probability versus P_T when $N = 2$, $P_S = P_R$, $\gamma_{th} = 2$ and $\Omega_{LI} = -23$ dB for various m and m_{LI} values. It is shown that increasing m swiftly improves the outage probability and decreases the error floors appeared at the high SNR region. To illustrate, the error floors for $m = 1$ and $m = 2$ are observed at approximately 1.2×10^{-3} and 2.4×10^{-6} , respectively, when $m_{LI} = 5$. On the other hand, a dramatic increase of m_{LI} has a limited effect on the outage performance. For instance, the outage probability curves for $m_{LI} = 1$ and $m_{LI} = 5$ when $m = 1$ get close values until they reach the error floors occurred at 1.8×10^{-2} and 1.2×10^{-2} , respectively. Therefore, the effect of increasing m in terms of outage probability is more dominant with respect to the effect of m_{LI} in the TWFD-AF-RS system.

In Figure 5.7, outage performance versus different values of the residual LI power is presented for $P_T \in \{20 \text{ dB}, 30 \text{ dB}, 40 \text{ dB}\}$ when $N = 2$, $P_S = 2P_R$, $\gamma_{th} = 3$, $m = 1$ and $m_{LI} = 3$. It is seen that the outage probability curves have lower values when either P_T increases or Ω_{LI} decreases. Hence, the increase in the efficiency of the LI cancellation process provides better outage results. Furthermore, the small gap between the exact and lower-bound curves vanishes as long as total transmission power increases.

Figure 5.8 demonstrates the outage performance of the TWFD-AF-RS systems versus different γ_{th} values when $N = 2$, $4P_S = P_R$, $P_T = 20$ dB and $\Omega_{LI} = -30$ dB. During

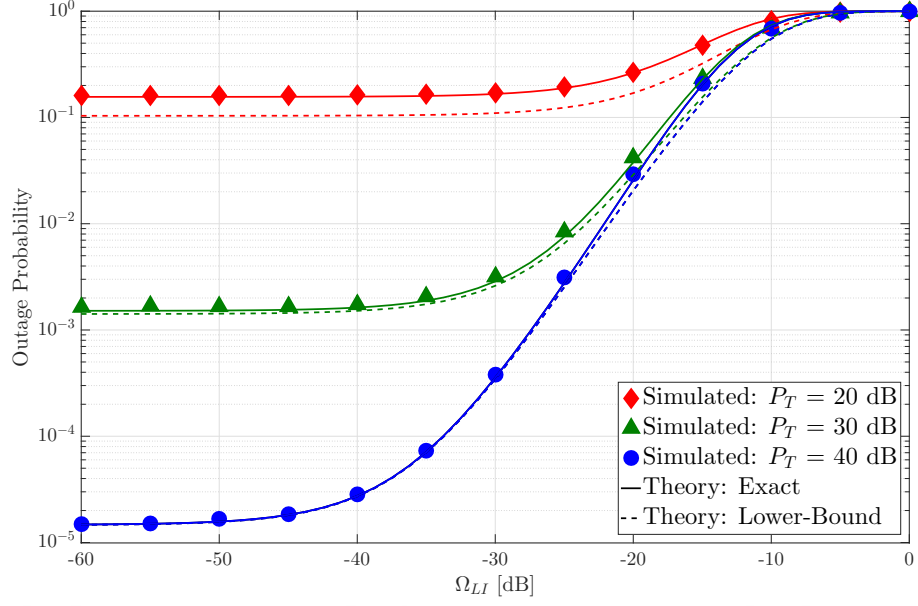


Figure 5.7 : Outage probability of the TWFD-AF-RS systems at S_1 versus Ω_{LI} when $N = 2$, $P_S = 2P_R$, $\gamma_{th} = 3$, $m = 1$ and $m_{LI} = 3$.

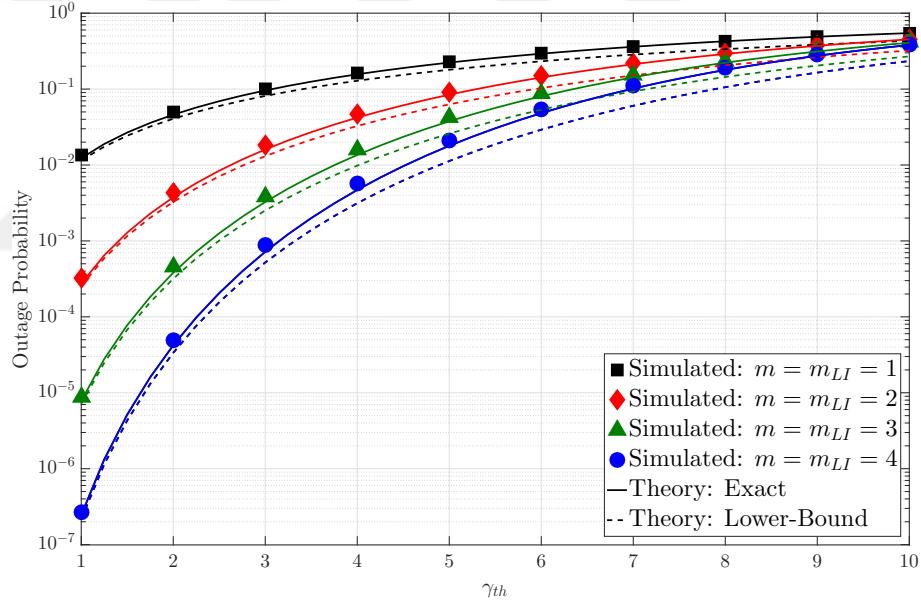


Figure 5.8 : Outage probability of the TWFD-AF-RS systems at S_1 versus γ_{th} when $N = 2$, $4P_S = P_R$, $P_T = 20$ dB and $\Omega_{LI} = -30$ dB.

that analysis, we consider that m and m_{LI} are equal to each other and take the integer values from 1 to 4. It is shown that the difference between the exact and lower-bound results decreases for the lower values of γ_{th} . Moreover, the outage performance can be improved by the higher values of m and m_{LI} as expected.

In this chapter, the outage probability of the TWFD-AF and TWFD-AF-RS systems enabling two-way transmission in a single time-slot has been investigated over Nakagami- m fading channels in the presence of residual LI at both source and relay

nodes^{1,2}. The exact outage probability expressions for each system have been derived in a single-integral form by using the CDF of the end-to-end SINR. The accuracy of the theoretical analysis has been verified by Monte Carlo type computer simulation results. Moreover, the tight lower-bound and asymptotic outage probability expressions have been obtained in the closed-forms. The numerical results have been revealed that outage performance of the TWFD-AF-RS systems with max-min relay selection policy can be improved by increasing either the number of available relay nodes and/or the efficiency of LI cancellation process. It has been demonstrated that the error floors occurred due to residual LI at the relay and source can be easily calculated with the asymptotic results.

¹Separate parts of this chapter are presented in *IEEE 84th Veh. Technol. Conf. (VTC Fall)* [83] and *24th Signal Process. Commun. Applicat. Conf. (SIU 2016)* [84].

²A part of this chapter is also submitted to *Ann. Telecommun.*

6. FULL-DUPLEX SPATIAL MODULATION SYSTEMS

As mentioned earlier, the spectral efficiency can be simply doubled by using the FD communications instead of the conventional HD communications. For further improvement in terms of the spectral efficiency and reliability of wireless networks, FD communications can be combined with the promising MIMO technique as spatial modulation (SM) [38], where the information bits are conveyed via both the modulated M -ary signals and the spatial domain by activating a single transmit antenna. Hence, the requirement of multiple transmit RF chains is relaxed in the SM technique. Moreover, it is shown that SM with the optimum ML detection provides error performance gains over popular MIMO techniques, e.g., V-BLAST architecture [42]. Additionally, instead of using the M -ary signals, all information bits can be conveyed via the transmit antenna indices by using a constant known signal (e.g., $+1$), which is named as space-shift-keying (SSK) [39–41] as a special case of SM.

In the literature, there are a few studies that combine FD communications with SM technique for the non-cooperative [45–47] and cooperative [48, 49] schemes. In this section of the thesis, we first analyze the BER performance of two-way full-duplex spatial modulation (TWFD-SM) systems over Rayleigh fading channels under imperfect channel state information (I-CSI). The upper-bound BER expression of the proposed system is derived in a closed-form. It is shown that the TWFD-SM systems can outperform the conventional two-way half-duplex spatial modulation (TWHD-SM) systems. Besides, the error performance of one-way full-duplex space-shift-keying with amplify-and-forward (OWFD-SSK-AF) relay systems is investigated over Rayleigh fading channels and the upper-bound BER expression is also derived in a closed-form. The BER performance of the conventional one-way half-duplex space-shift-keying with amplify-and-forward (OWHD-SSK-AF) relay systems is compared with the proposed OWFD-SSK-AF systems, which can provide quite better performance especially for the higher values of spectral efficiency.

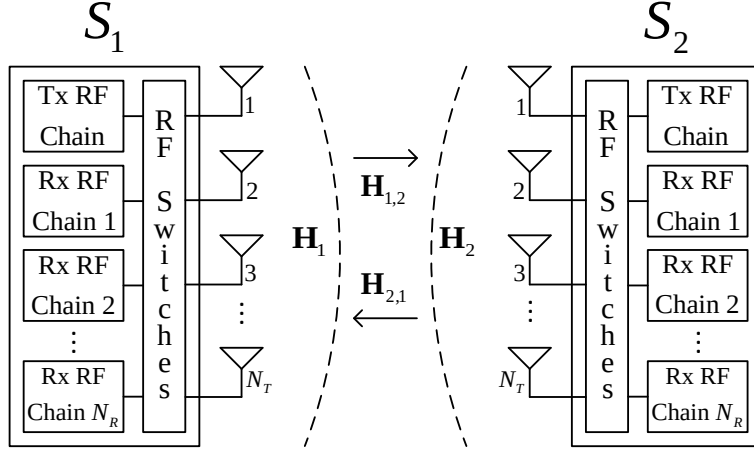


Figure 6.1 : TWFD-SM system model.

6.1 TWFD-SM Systems under I-CSI

The first proposed FD system using the SM technique is the TWFD-SM systems. The performance of the TWFD-SM systems over Rayleigh fading channels is investigated in the presence of channel estimation errors. Therefore, we assume that the receiver has I-CSI so as to recover the information bearing bits. We provide upper-bound BER expressions of TWFD-SM systems for both PSK and QAM signals. The theoretical analyses are supported by Monte Carlo type computer simulation results.

6.1.1 System model

In the proposed TWFD-SM system illustrated in Figure 6.1, bi-directional communications is established between two source nodes ($S_i, i \in \{1, 2\}$). Each source node is equipped with N_T antennas that can either transmit or receive the signals by using the RF switches. According to the SM technique, only one antenna is employed for the transmission among N_T antennas in each symbol period, hence, it is required to place one transmit RF chain at each node. The remaining $N_T - 1$ antennas can be used as receive antennas; however, it could be difficult to place that many receive RF chains for the increasing values of N_T . Therefore, we assume that each source node has N_R receive RF chains, where $1 \leq N_R \leq N_T - 1$. The connection of the RF switches between N_R receive RF chains and $N_T - 1$ available antennas is selected arbitrarily. Thus, the total number of RF chains employed for TWFD-SM systems is equal to $N_R + 1$ at each source node. An example TWFD-SM system configuration is given in Figure 6.2, where $N_T = 4$ and $N_R = 2$. According to the input bit sequences,

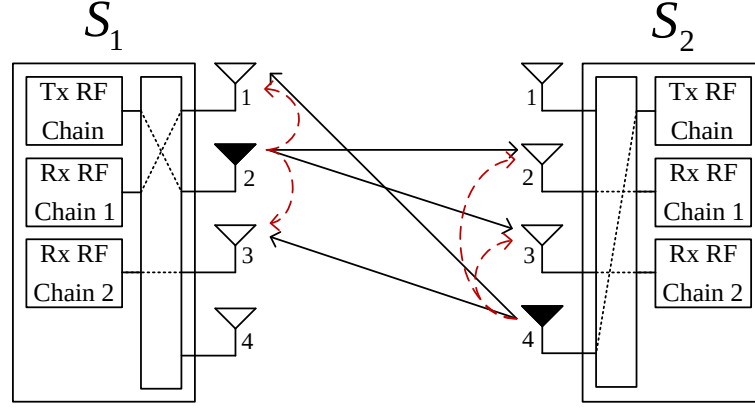


Figure 6.2 : An example configuration for a TWFD-SM system with $N_T = 4$ and $N_R = 2$ where the black solid and red dotted lines represent the channels between sources and residual LI channels, respectively.

the second and fourth antenna is connected to the transmit RF chain at S_1 and S_2 , respectively. Then, the receive RF chains are randomly connected to the available antennas. It is worthwhile to mention that the channels between sources and residual LI channels are represented via the black solid and red dotted lines, respectively.

In one symbol period, $\log_2(N_T)$ bits are conveyed by the transmit antenna index, while $\log_2(M)$ bits are transmitted with M -ary modulation (e.g., M -PSK/QAM) where M is the size of signal constellation. By means of the simultaneous transmission and reception in TWFD-SM systems, a total of $2\log_2(N_TM)$ bits are conveyed bi-directionally. The vector of received signals at S_i is given as

$$\mathbf{y}_i = \underbrace{\sqrt{P_S}\mathbf{H}_{j,i}\mathbf{x}_j}_{\text{Desired Signal}} + \underbrace{\sqrt{\beta P_S^\lambda}\mathbf{H}_i\mathbf{x}_i}_{\text{Residual Loop-Interference}} + \underbrace{\mathbf{n}_i}_{\text{Noise}} \quad (6.1)$$

where $j \in \{1, 2\}$, $j \neq i$, P_S is the transmission power, $\mathbf{x}_i$ and $\mathbf{x}_j$ are the unit-power SM signal vectors transmitted from S_i and S_j (i.e., $\mathbb{E}\{\mathbf{x}_i^H \mathbf{x}_i\} = 1$), respectively, $\mathbf{H}_{j,i}$ is the $N_R \times N_T$ channel matrix between all antennas of S_j and selected receive antennas of S_i , $\mathbf{H}_i$ is the $N_R \times N_T$ residual LI channel matrix and $\mathbf{n}_i$ is the $N_R \times 1$ complex Gaussian noise vector. β and λ are the constants that represent the quality of the LI cancellation process [8, 20, 48]. As mentioned in Section 3.3 as the second residual LI channel model, the experimental results show that the residual LI is proportional to the λ^{th} power of the transmission power ($0 \leq \lambda \leq 1$) and the value of λ can be reduced to 0.21 after the cancellation processes [8]. The entries of $\mathbf{H}_{j,i}$ and $\mathbf{H}_i$ are i.i.d. according to $\mathcal{CN}(0, 1)$. Similarly, the entries of $\mathbf{n}_i$ are i.i.d. as $\mathcal{CN}(0, \sigma_S^2)$.

In the TWFD-SM systems, $\mathbf{x}_j$ can be given by

$$\mathbf{x}_j = [0 \ \cdots \ 0 \ \underbrace{x_j^{q_j}}_{k_j^{th} \text{ position}} \ 0 \ \cdots \ 0]^T \quad (6.2)$$

where k_j is the selected transmit antenna index at S_j according to first $\log_2(N_T)$ bits and $x_j^{q_j}$ denotes the q_j^{th} symbol from the M -ary constellation according to following $\log_2(M)$ bits. Similarly, one can easily obtain $\mathbf{x}_i$ in terms of k_i and $x_i^{q_i}$. The received signal vector given in (6.1) can be rewritten as

$$\mathbf{y}_i = \underbrace{\sqrt{P_S} \mathbf{h}_{j,i}^{k_j} x_j^{q_j}}_{\text{Desired Signal}} + \underbrace{\sqrt{\beta P_S^\lambda} \mathbf{h}_i^{k_i} x_i^{q_i}}_{\text{Loop Interference}} + \underbrace{\mathbf{n}_i}_{\text{Noise}} \quad (6.3)$$

where $\mathbf{h}_{j,i}^{k_j}$ and $\mathbf{h}_i^{k_i}$ are respectively the k_j^{th} and k_i^{th} columns of $\mathbf{H}_{j,i}$ and $\mathbf{H}_i$. When $\mathbf{h}_{j,i}^{k_j}$ is given, then $\mathbf{y}_i \sim \mathcal{CN}(\sqrt{P_S} \mathbf{h}_{j,i}^{k_j} x_j^{q_j}, \beta P_S^\lambda |x_i^{q_i}|^2 + \sigma_S^2)$. If the receiver has P-CSI, the optimum detector according to the ML principle is given by [42]

$$[\hat{k}_j, \hat{q}_j] = \arg \min_{\tilde{k}_j, \tilde{q}_j} \left\| \mathbf{y}_i - \sqrt{P_S} \mathbf{z}(\tilde{k}_j, \tilde{q}_j) \right\|^2 \quad (6.4)$$

where $\hat{k}_j$ and $\hat{q}_j$ are the recovered transmit antenna and M -ary symbol index, respectively, and $\mathbf{z}(k_j, q_j) = \mathbf{h}_{j,i}^{k_j} x_j^{q_j}$.

In the practice, the receiver has I-CSI because of the channel estimation errors. According to the pilot-assisted least square (LS) estimation [85], the estimated channel vector in the presence of channel estimation errors is expressed as $\hat{\mathbf{h}}_{j,i}^{k_j} = \mathbf{h}_{j,i}^{k_j} + \epsilon_{r,i}$ where $\epsilon_{r,i} \sim \mathcal{CN}(0, \sigma_\epsilon^2)$. The entries of $\epsilon_{r,i}$ are independent from $\mathbf{h}_{j,i}^{k_j}$. By defining T as the number of pilot signals and P_S/σ_S^2 as the SNR, the variance of the estimation error is given by $\sigma_\epsilon^2 = (TP_S/\sigma_S^2)^{-1}$ [43]. Hence, the entries of $\hat{\mathbf{h}}_{j,i}^{k_j}$ are i.i.d. according to $\mathcal{CN}(0, 1 + \sigma_\epsilon^2)$ and the correlation coefficient for the entries of $\hat{\mathbf{h}}_{j,i}^{k_j}$ and $\mathbf{h}_{j,i}^{k_j}$ becomes $\Upsilon = 1/\sqrt{1 + \sigma_\epsilon^2}$. For higher SNR values and/or increasing number of pilot signals, the channel estimation error decreases and the correlation between the exact and the estimated channel coefficients increases.

In order to formulate the optimal ML detector under I-CSI, it is required to obtain the distribution characteristics of $\mathbf{y}_i$ for a given $\hat{\mathbf{h}}_{j,i}^{k_j}$. First, we need to derive the conditional PDF of $\mathbf{h}_{j,i}^{k_j}$ for a given $\hat{\mathbf{h}}_{j,i}^{k_j}$ by the Bayes' rule as follows:

$$f_{\mathbf{h}_{j,i}^{k_j} | \hat{\mathbf{h}}_{j,i}^{k_j}}(\mathbf{x} | \mathbf{y}) = \frac{f_{\mathbf{h}_{j,i}^{k_j}, \hat{\mathbf{h}}_{j,i}^{k_j}}(\mathbf{x}, \mathbf{y})}{f_{\hat{\mathbf{h}}_{j,i}^{k_j}}(\mathbf{y})} = \frac{e^{-\|\mathbf{x} - \Upsilon \mathbf{y}\|^2 / (1 - \Upsilon^2)}}{[\pi (1 - \Upsilon^2)]^{N_R/2}}. \quad (6.5)$$

Using (6.3) and (6.5), we obtain $\mathbf{y}_i \sim \mathcal{CN}(\Upsilon^2 \sqrt{P_S} \hat{\mathbf{z}}(k_j, q_j), \sigma_{y,i}^2)$ for a given $\hat{\mathbf{h}}_{j,i}(k_j)$, where $\hat{\mathbf{z}}(k_j, q_j) = \hat{\mathbf{h}}_{j,i}^{k_j} x_j^{q_j}$ and $\sigma_{y,i}^2 = (1 - \Upsilon^2) P_S |x_j^{q_j}|^2 + \beta P_S^\lambda |x_i^{q_i}|^2 + \sigma_S^2$. After some manipulations, the optimal ML detector for TWFD-SM systems under I-CSI can be obtained as

$$[\hat{k}_j, \hat{q}_j] = \arg \min_{\tilde{k}_j, \tilde{q}_j} \left\{ \frac{\left\| \mathbf{y}_i - \Upsilon^2 \sqrt{P_S} \hat{\mathbf{z}}(\tilde{k}_j, \tilde{q}_j) \right\|^2}{\sigma_{y,i}^2 N_R} + \ln(\sigma_{y,i}^2) \right\}. \quad (6.6)$$

It can be noted that $|x_j^{q_j}| = 1$ for M -PSK signals, hence, $\sigma_{y,i}^2$ does not depend on neither k_j nor q_j . Therefore, the optimal ML detector under I-CSI reduces to

$$[\hat{k}_j, \hat{q}_j] = \arg \min_{k_j, q_j} \left\| \mathbf{y}_i - \Upsilon^2 \sqrt{P_S} \hat{\mathbf{z}}(k_j, q_j) \right\|^2. \quad (6.7)$$

6.1.2 BER analyses

In this section, we derive BER expressions of the TWFD-SM systems for both M -PSK and M -QAM signals by using PEP of the TWFD-SM systems.

6.1.2.1 BER analysis for M -PSK

By using (6.7), conditional pairwise error probability (CPEP) of the TWFD-SM systems for M -PSK is given by

$$\begin{aligned} P(\mathbf{x}_j \rightarrow \hat{\mathbf{x}}_j | \hat{\mathbf{H}}_{j,i}) &= P \left\{ \left\| \mathbf{y}_i - \Upsilon^2 \sqrt{P_S} \hat{\mathbf{z}}(k_j, q_j) \right\|^2 \right. \\ &\quad \left. \geq \left\| \mathbf{y}_i - \Upsilon^2 \sqrt{P_S} \hat{\mathbf{z}}(\hat{k}_j, \hat{q}_j) \right\|^2 \middle| \hat{\mathbf{H}}_{j,i} \right\} \\ &= P \left\{ \Upsilon^2 \sqrt{P_S} \left(\left\| \hat{\mathbf{z}}(k_j, q_j) \right\|^2 - \left\| \hat{\mathbf{z}}(\hat{k}_j, \hat{q}_j) \right\|^2 \right) \right. \\ &\quad \left. - 2\Re \left[\mathbf{y}_i^H \left(\hat{\mathbf{z}}(k_j, q_j) - \hat{\mathbf{z}}(\hat{k}_j, \hat{q}_j) \right) \right] \geq 0 \middle| \hat{\mathbf{H}}_{j,i} \right\} \\ &= P(\xi_i \geq 0 | \hat{\mathbf{H}}_{j,i}) \end{aligned} \quad (6.8)$$

where $\hat{\mathbf{x}}_j$ is the decided symbol when $\mathbf{x}_j$ is transmitted and $\hat{\mathbf{H}}_{j,i}$ is the estimated channel matrix. Then, the expected value and the variance of ξ_i for a given $\hat{\mathbf{H}}_{j,i}$ are obtained as

$$\begin{aligned} \mathbb{E} \left\{ \xi_i | \hat{\mathbf{H}}_{j,i} \right\} &= -\Upsilon^2 \sqrt{P_S} \kappa, \\ \text{Var} \left\{ \xi_i | \hat{\mathbf{H}}_{j,i} \right\} &= 2\kappa \sigma_{y,i}^2 \end{aligned} \quad (6.9)$$

where $\kappa = \left\| \hat{\mathbf{z}}(k_j, q_j) - \hat{\mathbf{z}}(\hat{k}_j, \hat{q}_j) \right\|^2$ is a chi-squared random variable with $2N_R$ degrees of freedom. Therefore, κ can be also defined as $\kappa = \sum_{n=1}^{2N_R} \alpha_n^2$ where

$\alpha_n \sim \mathcal{N}\{0, (1 + \sigma_\epsilon^2) \sigma_\alpha^2/2\}$ and

$$\sigma_\alpha^2 = \begin{cases} |x_j^{q_j}|^2 + |x_j^{\hat{q}_j}|^2, & k_j \neq \hat{k}_j \\ |x_j^{q_j} - x_j^{\hat{q}_j}|^2, & k_j = \hat{k}_j. \end{cases} \quad (6.10)$$

Considering $|x_j^{q_j}|^2 = 1$ for M -PSK signals, $\sigma_\alpha^2 = 2$ when $k_j \neq \hat{k}_j$. By using the Gaussian random variable ξ_i , the CPEP expression given in (6.8) can be rewritten in terms of Q -function as follows:

$$P(\mathbf{x}_j \rightarrow \hat{\mathbf{x}}_j | \hat{\mathbf{H}}_{j,i}) = Q\left(\sqrt{\frac{\Upsilon^4 P_S \kappa}{2\sigma_{y,i}^2}}\right). \quad (6.11)$$

By using Craig's formula [86, Eq. (4.2)] and integrating the CPEP over the PDF of κ , the PEP can be obtained as

$$\begin{aligned} P(\mathbf{x}_j \rightarrow \hat{\mathbf{x}}_j) &= \frac{1}{\pi} \int_{u=0}^{\infty} \int_{\theta=0}^{\pi/2} \exp\left(-\frac{\Upsilon^4 P_S u}{4\sigma_{y,i}^2 \sin^2 \theta}\right) p_\kappa(u) d\theta du \\ &= \frac{1}{\pi} \int_{\theta=0}^{\pi/2} M_\kappa\left(\frac{\Upsilon^4 P_S}{4\sigma_{y,i}^2 \sin^2 \theta}\right) d\theta \end{aligned} \quad (6.12)$$

where $M_\kappa(s) = [1 + s(1 + \sigma_\epsilon^2) \sigma_\alpha^2]^{-N_R}$ is the moment generating function (MGF) of κ [86]. Afterwards, the PEP expression can be calculated as

$$P(\mathbf{x}_j \rightarrow \hat{\mathbf{x}}_j) = \frac{1}{\pi} \int_{\theta=0}^{\pi/2} \left(\frac{\sin^2 \theta}{\sin^2 \theta + \frac{\Upsilon^2 P_S \sigma_\alpha^2}{4\sigma_{y,i}^2}} \right)^{N_R} d\theta. \quad (6.13)$$

By using [86, Eq. (5A.4b)], the above integration can be solved and the PEP expression found in a closed-form is given by

$$P(\mathbf{x}_j \rightarrow \hat{\mathbf{x}}_j) = \mu_i^{N_R} \sum_{b=0}^{N_R-1} \binom{N_R + b - 1}{b} (1 - \mu_i)^b \quad (6.14)$$

where $\mu_i = \frac{1}{2} \left(1 - \sqrt{\Upsilon^2 P_S \sigma_\alpha^2 / (4\sigma_{y,i}^2 + \Upsilon^2 P_S \sigma_\alpha^2)} \right)$. Finally, the BER expression at S_i can be derived by using the well known union-bound technique [42, 43, 47–49] and given by

$$P_b^{(i)} \leq \sum_{k_j=1}^{N_T} \sum_{q_j=1}^M \sum_{\hat{k}_j=1}^{N_T} \sum_{\hat{q}_j=1}^M \frac{N(\mathbf{x}_j, \hat{\mathbf{x}}_j) P(\mathbf{x}_j \rightarrow \hat{\mathbf{x}}_j)}{\log_2(N_T M) N_T M} \quad (6.15)$$

where $N(\mathbf{x}_j, \hat{\mathbf{x}}_j)$ represents the number of bit errors associated with the corresponding pairwise error event.

6.1.2.2 BER analysis for M -QAM

In contrast to M -PSK signals, the envelope of M -QAM signals is non-constant and it changes in each symbol period. Therefore, the analysis for M -QAM is quite complicated for the decision metric given in (6.6). Thus, for the BER analysis in TWFD-SM systems for M -QAM signals under I-CSI, the mismatched ML detector is employed at the receiver as in [43] and its decision metric is given by

$$\left[\hat{k}_j, \hat{q}_j \right] = \arg \min_{\tilde{k}_j, \tilde{q}_j} \left\| \mathbf{y}_i - \sqrt{P_S} \hat{\mathbf{z}} \left(\tilde{k}_j, \tilde{q}_j \right) \right\|^2. \quad (6.16)$$

The decision metric for the mismatched ML detector is quite similar to the P-CSI case given in (6.4). The only difference between them is that the receiver uses $\hat{\mathbf{h}}_{j,i}^{k_j}$ instead of $\mathbf{h}_{j,i}^{k_j}$ in the mismatched ML detector. Similar to (6.8), the CPEP for TWFD-SM systems with M -QAM signals is written as

$$\begin{aligned} P \left(\mathbf{x}_j \rightarrow \hat{\mathbf{x}}_j | \hat{\mathbf{H}}_{j,i} \right) &= P \left\{ \sqrt{P_S} \left(\left\| \hat{\mathbf{z}}(k_j, q_j) \right\|^2 - \left\| \hat{\mathbf{z}}(\hat{k}_j, \hat{q}_j) \right\|^2 \right) \right. \\ &\quad \left. - 2\Re \left[\mathbf{y}_i^H \left(\hat{\mathbf{z}}(k_j, q_j) - \hat{\mathbf{z}}(\hat{k}_j, \hat{q}_j) \right) \right] \geq 0 \mid \hat{\mathbf{H}}_{j,i} \right\} \\ &= P \left(\psi_i \geq 0 \mid \hat{\mathbf{H}}_{j,i} \right) \end{aligned} \quad (6.17)$$

where the expected value and the variance of ψ_i as a Gaussian distributed random variable for a given $\hat{\mathbf{H}}_{j,i}$ are obtained as

$$\begin{aligned} E \left\{ \psi_i \mid \hat{\mathbf{H}}_{j,i} \right\} &= \sqrt{P_S} \left[2\Re \left\{ \left[\hat{\mathbf{z}}(k_j, q_j) \right]^H \hat{\mathbf{z}}(\hat{k}_j, \hat{q}_j) \right\} \right. \\ &\quad \left. - (2\Upsilon^2 - 1) \left\| \hat{\mathbf{z}}(k_j, q_j) \right\|^2 - \left\| \hat{\mathbf{z}}(\hat{k}_j, \hat{q}_j) \right\|^2 \right], \end{aligned} \quad (6.18)$$

$$\text{Var} \left\{ \psi_i \mid \hat{\mathbf{H}}_{j,i} \right\} = 2\kappa\sigma_{y,i}^2.$$

In the practice, the variance of the channel estimation error should be $\sigma_\epsilon^2 \ll 1$. Thus, we can take $(2\Upsilon^2 - 1) \approx 1$ so as to simplify the expectation of ψ_i which provides a more tractable expression for the BER calculations. Then, one can easily obtain that $E \left\{ \psi_i \mid \hat{\mathbf{H}}_{j,i} \right\} \approx -\sqrt{P_S}\kappa$. By using Q -function, the CPEP is calculated as

$$P \left(\mathbf{x}_j \rightarrow \hat{\mathbf{x}}_j \mid \hat{\mathbf{H}}_{j,i} \right) \approx Q \left(\sqrt{\frac{P_S\kappa}{2\sigma_{y,i}^2}} \right). \quad (6.19)$$

By applying Craig's formula [86, Eq. (4.2)], averaging on κ and using the MGF of κ , respectively, the PEP is given by

$$P \left(\mathbf{x}_j \rightarrow \hat{\mathbf{x}}_j \right) \approx \frac{1}{\pi} \int_{\theta=0}^{\pi/2} \left(\frac{\sin^2 \theta}{\sin^2 \theta + \frac{P_S \sigma_\epsilon^2}{4\Upsilon^2 \sigma_{y,i}^2}} \right)^{N_R} d\theta. \quad (6.20)$$

Similar to (6.13), the above integration can be solved as

$$P(\mathbf{x}_j \rightarrow \hat{\mathbf{x}}_j) \approx \omega_i^{N_R} \sum_{b=0}^{N_R-1} \binom{N_R+b-1}{b} (1-\omega_i)^b \quad (6.21)$$

where $\omega_i = \frac{1}{2} \left(1 - \sqrt{P_S \sigma_\alpha^2 / 4 \Upsilon^2 \sigma_{y,i}^2 + P_S \sigma_\alpha^2} \right)$. It is worthwhile to mention that the value of $\sigma_{y,i}^2$ is independent from $x_i^{q_i}$ for M -PSK signals due to $|x_i^{q_i}|^2 = 1$. In M -QAM, however, the value of $\sigma_{y,i}^2$ changes with different $x_i^{q_i}$ and it should be taken into consideration for the BER calculations. As in M -PSK, the upper-bound for the BER expression of M -QAM derived by using the union-bound approach is given by

$$P_b^{(i)} \leq \sum_{q_i=1}^M \sum_{k_j=1}^{N_T} \sum_{q_j=1}^M \sum_{\hat{k}_j=1}^{N_T} \sum_{\hat{q}_j=1}^M \frac{N(\mathbf{x}_j, \hat{\mathbf{x}}_j) P(\mathbf{x}_j \rightarrow \hat{\mathbf{x}}_j)}{\log_2(N_T M) N_T M^2}. \quad (6.22)$$

Both of the derived BER expressions of the TWFD-SM systems for M -PSK and M -QAM signals, given in (6.15) and (6.22), are also valid for the TWHD-SM systems by considering $\beta = 0$ in (6.1).

6.1.3 Spectral efficiency of TWFD-SM and TWHD-SM

The TWFD-SM systems allow the simultaneous transmission and reception operations over the same frequency band and a total of $2 \log_2(N_T M)$ bits are conveyed in each time-slot. In the TWHD-SM systems, however, the transmission and the reception operations are accomplished by using two different time slots and only $\log_2(N_T M)$ bits are transmitted in each time slot. Consequently, spectral efficiency expressions of the TWFD-SM and TWHD-SM systems are respectively given by

$$\begin{aligned} \eta_{FD} &= 2 \log_2(N_T M), \\ \eta_{HD} &= \log_2(N_T M). \end{aligned} \quad (6.23)$$

In TWHD-SM, either M -ary constellation size or N_T should be increased to obtain the same spectral efficiency with TWFD-SM.

6.1.4 Numerical results

We present the theoretical and Monte Carlo simulation results for the BER performance of the TWFD-SM systems over Rayleigh fading channels in order to prove the accuracy of the analyses. Furthermore, the performances of the TWFD-SM and TWHD-SM systems under I-CSI are compared. In order to make a fair comparison

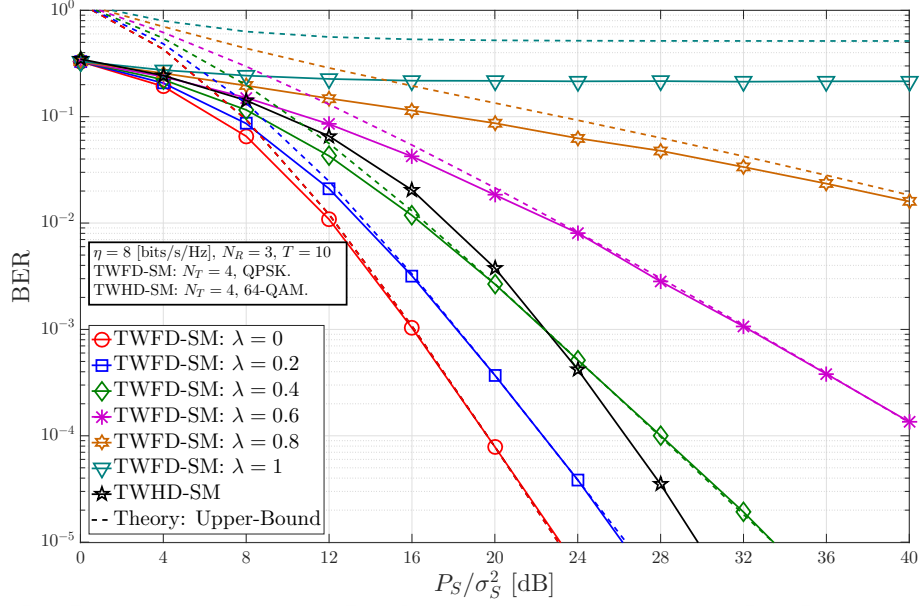


Figure 6.3 : BER performance comparison of TWFD-SM and TWHD-SM under I-CSI for $\eta = 8$ [bits/s/Hz], $N_T = 4$, $N_R = 3$ and $T = 10$.

between TWFD-SM and TWHD-SM, the total number of available RF chains are taken as the same. For all computer simulation scenarios, BER curves are produced for the received signal at S_1 (i.e., $P_b^{(1)}$) and we assume that $\beta = 1$ for TWFD-SM as in [20].

In Figure 6.3, the BER performance of the TWFD-SM systems with QPSK is plotted versus P_S/σ_S^2 for different values of λ between the optimistic ($\lambda = 0$) and pessimistic ($\lambda = 1$) cases, where $\eta = 8$ [bits/s/Hz], $N_T = 4$, $N_R = 3$ and $T = 10$. Note that T represents the number of the pilot signals. The theoretical results are generated by using (6.15). It is seen that the computer simulation results are tightly upper-bounded by the theoretical results as long as either the SNR increases or λ decreases. Moreover, the performance of the proposed system is compared with TWHD-SM having the same configuration with 64-QAM to ensure the same spectral efficiency. We observe that TWFD-SM outperforms TWHD-SM when λ decreases, i.e., the quality of LI cancellation increases. For instance, TWFD-SM provides approximately 6.7 dB and 4 dB performance improvement over TWHD-SM at a BER value of 10^{-4} for $\lambda = 0$ and $\lambda = 0.2$, respectively. Furthermore, TWFD-SM having $\lambda = 0.4$ performs better than TWHD-SM until 22.5 dB. However, TWHD-SM provides lower BER values for $\lambda > 0.4$ in this scenario. Although the case of $\lambda = 0$ is an optimistic scenario, as stated in Section 6.1.1, the experimental results show that λ could be decreased to 0.21 after the LI cancellation processes [8].

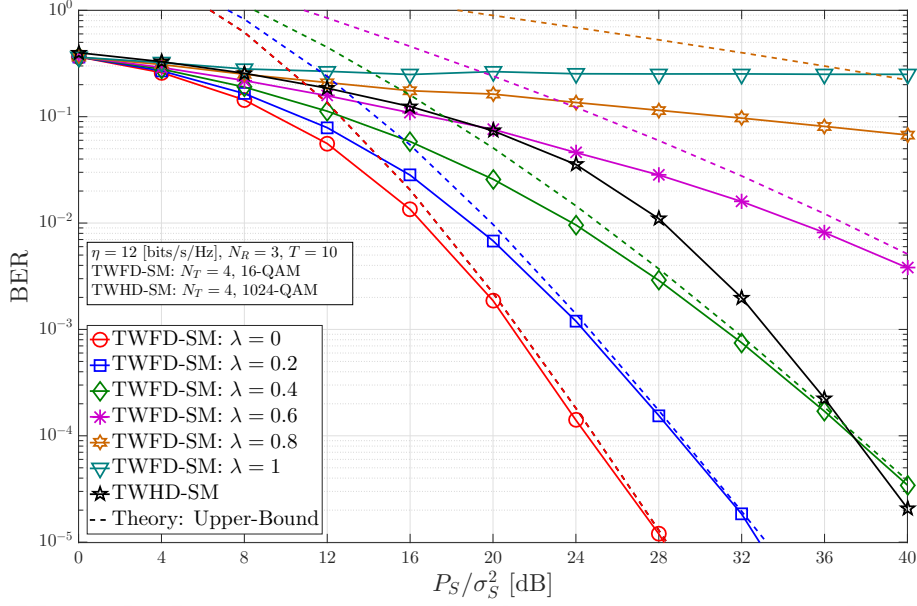


Figure 6.4 : BER performance comparison of TWFD-SM and TWHD-SM under I-CSI for $\eta = 12$ [bits/s/Hz], $N_T = 4$, $N_R = 3$ and $T = 10$.

Figure 6.4 represents the BER performance of the TWFD-SM systems with 16-QAM for $\lambda = \{0, 0.2, 0.4, 0.6, 0.8, 1\}$. Here, $\eta = 12$ [bits/s/Hz], $N_T = 4$, $N_R = 3$ and $T = 10$. It is observed that the theoretical results obtained by using (6.22) are quite tight with the computer simulation results and the gap between them vanishes for higher SNR and/or lower λ values. The BER curve of TWHD-SM with 1024-QAM at the same spectral efficiency is also included into the analysis. In the light of previous results given in Figure 6.3, it is observed that the performance improvement of TWFD-SM with respect to TWHD-SM increases for higher spectral efficiencies. For example, the performance degradation of TWHD-SM at a BER value of 10^{-4} approaches to 12.8 dB and 8.5 dB compared to TWFD-SM with $\lambda = 0$ and $\lambda = 0.2$, respectively.

Figure 6.5 plots the BER performance of the TWFD-SM systems with BPSK under P-CSI and I-CSI with different number of pilot signals, where $\eta = 8$ [bits/s/Hz], $N_T = 8$, $N_R = 2$ and $\lambda = 0.2$. We observe that TWFD-SM is also robust to the channel estimation errors, e.g., the performance degradation of I-CSI against P-CSI at a BER value of 10^{-5} is only 0.1 dB, 0.3 dB and 0.7 dB when T is 10, 3 and 1, respectively.

In Figure 6.6, the BER performance of the TWFD-SM systems with BPSK under I-CSI is demonstrated for $N_R \in \{1, 2, 3\}$, where $\eta = 6$ [bits/s/Hz], $N_T = 4$, $T = 5$ and $\lambda = 0.25$. According to the increase of N_R , the performance of the proposed system

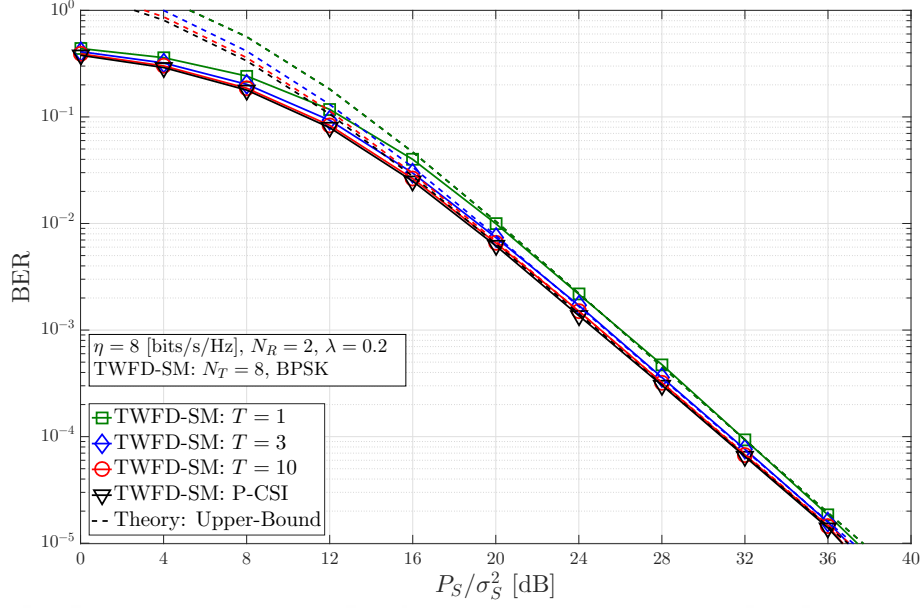


Figure 6.5 : BER performance of TWFD-SM with BPSK under P-CSI and under I-CSI for different number of pilot signals, where $\eta = 8$ [bits/s/Hz], $N_T = 8$, $N_R = 2$ and $\lambda = 0.2$.

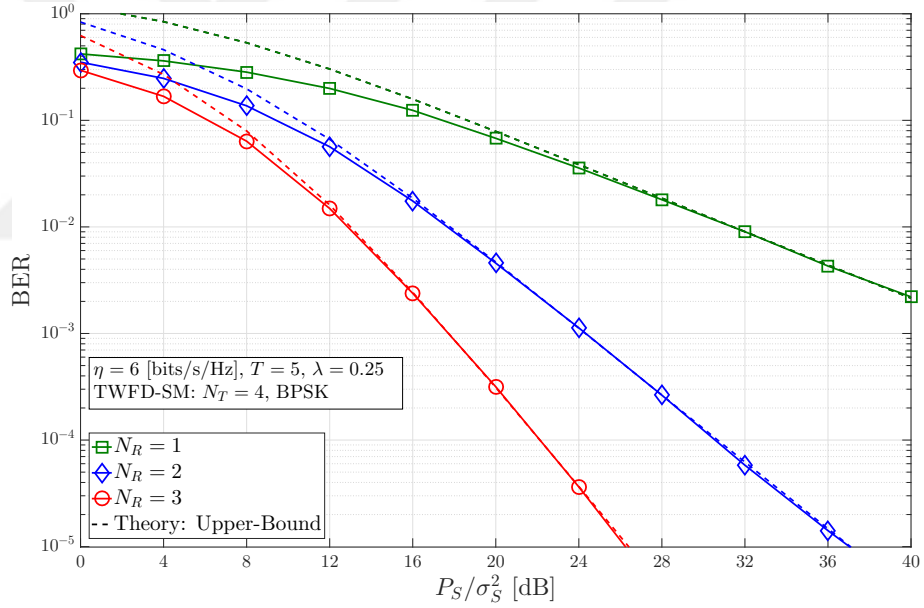


Figure 6.6 : BER performance of TWFD-SM with BPSK under I-CSI for different values of N_R , where $\eta = 6$ [bits/s/Hz], $N_T=4$, $T=5$ and $\lambda=0.25$.

can be enhanced; however, it costs an increase in the number of RF chains. Therefore, there is a trade-off between the number of RF chains and BER performance.

6.2 SSK with OWFD Fixed-Gain AF Relay Systems

SSK as a special case of SM uses only the spatial constellation for the information transmission instead of M -PSK/QAM signals [39]. To the best of our knowledge, although there are two different studies that analyze SM technique with FD relaying

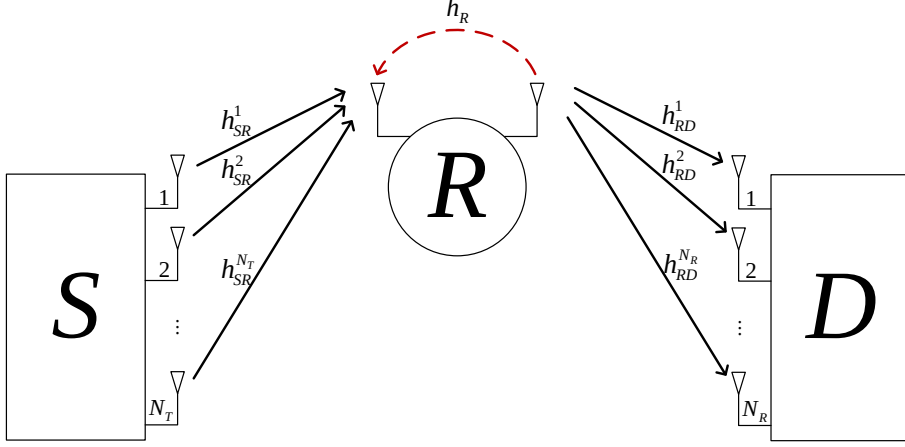


Figure 6.7 : OWFD-SSK-AF system model.

using DF protocol [48, 49], there is no performance analyses of the SSK technique with FD relaying with AF protocol. In the scope of this thesis, the error performance of the OWFD-SSK-AF systems is investigated over Rayleigh fading channels and an upper-bound for BER expression is derived in a closed-form. The theoretical analyses are verified by Monte Carlo type computer simulations. We observe that OWFD-SSK-AF outperforms the conventional OWHD-SSK-AF systems as long as the quality of LI cancellation improves and/or the spectral efficiency increases.

6.2.1 System model

We consider a dual-hop one-way transmission scheme for the OWFD-SSK-AF system as shown in Figure 6.7. As seen from the system model, the source node (S) has N_T transmit antennas and the destination node (D) has N_R receive antennas. SSK transmission scheme is employed at S in order to convey $\log_2(N_T)$ information bits in each symbol period. We assume that the direct link between the source and destination is strongly attenuated. Therefore, the communication between them can be established only via the relay node (R) consisting of single transmit and receive antenna. The relay operates in FD mode and uses the AF protocol to retransmit the received signal to the destination.

The received signal from the relay at time index t is given by

$$r[t] = \sqrt{P_S} \mathbf{h}_{SR} \mathbf{x}[t] + h_{RS} s[t] + n_R[t] \quad (6.24)$$

where P_S is the transmission power of the source, $\mathbf{x}[t]$ is the $N_T \times 1$ deterministic unit-power signal vector transmitted from the source, $s[t]$ is the transmitted signal

from the relay, $\mathbf{h}_{SR}$ is the $1 \times N_T$ channel coefficient vector between $S - R$ link, h_R is the residual LI channel coefficient at the relay and $n_R[t]$ is the complex Gaussian noise signal. The entries of $\mathbf{h}_{SR}$ are i.i.d. according to $\mathcal{CN}(0, \Omega_{SR})$. Similarly, h_R and $n_R[t]$ are respectively distributed as $h_R \sim \mathcal{CN}(0, \Omega_R)$ and $n_R[t] \sim \mathcal{CN}(0, \sigma_R^2)$. Similar to the TWFD-SM systems, power of the residual LI channel Ω_R is related to transmission power of the relay as P_R and it is modeled as $\Omega_R = \beta P_R^{\lambda-1}$ where β and λ are the constants that reflects the quality of LI cancellation ($0 \leq \lambda \leq 1$). Then, the transmitted signal from the source can be given by

$$\mathbf{x}[t] = [0 \ \cdots \ 0 \ \underbrace{1}_{k[t]^{th} \text{ position}} \ 0 \ \cdots \ 0]^T \quad (6.25)$$

where $k[t] \in \{1, 2, \dots, N_T\}$ is the index of the selected transmit antenna for time index t . By combining (6.24) and (6.25), the received signal at the relay is rewritten as

$$r[t] = \sqrt{P_S} h_{SR}^{k[t]} + h_R s[t] + n_R[t] \quad (6.26)$$

where $h_{SR}^{k[t]}$ is the $k[t]^{th}$ entry of $\mathbf{h}_{SR}$. According to the AF protocol, the transmitted signal from the relay is given by

$$s[t] = G r[t-1] = G \sum_{u=0}^{\infty} (G h_R)^u q[t-u] \quad (6.27)$$

where G is the fixed-gain amplification factor and $q[t] = \sqrt{P_S} h_{SR}^{k[t-1]} + n_R[t-1]$. In order to guarantee finite relay transmit power and prevent oscillations, the relay gain must be limited by $G^2 < 1/|h_R|^2$ [10]. Otherwise, power of the residual LI goes to infinity. When the gain condition is satisfied, average transmission power at the relay for a given h_R is obtained as [10, Eq. (5)]

$$\begin{aligned} \mathbb{E} \{ |s[t]|^2 | h_{LI} \} &= G^2 (P_S \Omega_{SR} + \sigma_R^2) \sum_{u=0}^{\infty} (G^2 |h_R|^2)^u \\ &= \frac{P_S \Omega_{SR} + \sigma_R^2}{\frac{1}{G^2} - |h_R|^2}. \end{aligned} \quad (6.28)$$

The fixed gain OWFD-AF systems are analyzed in Section 4.1.1.2, where the fixed amplification gain is given in (4.9) in order to adjust the average transmission power at R to P_R . One can easily show that the average transmission power at R is equal to

$$\mathbb{E} \{ |s[t]|^2 \} = \mathbb{E} \left\{ \frac{P_S \Omega_{SR} + \sigma_R^2}{\frac{1}{G^2} - |h_R|^2} \right\} = \frac{P_S \Omega_{SR} + \sigma_R^2}{\frac{1}{G^2} - \Omega_R} = P_R. \quad (6.29)$$

By using (6.24), (6.29) and $\Omega_R = \beta P_R^{\lambda-1}$, the average LI power at the relay is found as $E\{|h_{RS}[t]|^2\} = \beta P_R^\lambda$. Afterwards, the received signal vector at the destination is given by

$$\begin{aligned} \mathbf{y}[t] &= \mathbf{h}_{RD} s[t] + \mathbf{n}_D[t] \\ &= \underbrace{G\sqrt{P_S}h_{SR}^{k[t-1]}\mathbf{h}_{RD}}_{\text{Desired Signal}} + \underbrace{G\mathbf{h}_{RD}h_{RS}[t-1]}_{\text{Residual Loop-Interference}} + \underbrace{G\mathbf{h}_{RD}n_R[t-1] + \mathbf{n}_D[t]}_{\text{Noise}} \end{aligned} \quad (6.30)$$

where $\mathbf{h}_{RD}$ denotes the $N_R \times 1$ channel coefficient vector and $\mathbf{n}_D[t]$ represents the complex Gaussian noise vector at the destination. The entries of $\mathbf{h}_{RD}$ and $\mathbf{n}_D[t]$ are i.i.d. according to $\mathcal{CN}(0, \Omega_{RD})$ and $\mathcal{CN}(0, \sigma_D^2)$, respectively. For given $\mathbf{h}_{SR}$ and $\mathbf{h}_{RD}$, the expected value and variance of the complex Gaussian distributed $\mathbf{y}[t]$ are respectively given by

$$\begin{aligned} E\{\mathbf{y}[t] | \mathbf{h}_{SR}, \mathbf{h}_{RD}\} &= G\sqrt{P_S}h_{SR}^{k[t-1]}\mathbf{h}_{RD}, \\ \text{Var}\{\mathbf{y}[t] | \mathbf{h}_{SR}, \mathbf{h}_{RD}\} &= G\|\mathbf{h}_{RD}\|^2 [\Omega_R P_R + \sigma_R^2] + \sigma_D^2. \end{aligned} \quad (6.31)$$

For the detection rule, we have omitted the time index t for brevity. Therefore, the optimum ML detection rule of the OWFD-SSK-AF system can be given by

$$\hat{k} = \arg \min_{1 \leq \hat{k} \leq N_T} \left\| \mathbf{y} - G\sqrt{P_S}h_{SR}^{\hat{k}}\mathbf{h}_{RD} \right\|^2 \quad (6.32)$$

where $\hat{k}$ is the determined transmit antenna index.

6.2.2 BER analysis

In this section, an upper-bound for the BER performance of the OWFD-SSK-AF systems is derived by using PEP approach. By taking into consideration the gain condition defined in the previous section, similar to (4.13), PEP expression of the OWFD-SSK-AF system can be written in terms of CPEP expressions and given by

$$P(k \rightarrow \hat{k}) = F_{\gamma_R}(C) P(k \rightarrow \hat{k} | \gamma_R < C) + \bar{F}_{\gamma_R}(C) P(k \rightarrow \hat{k} | \gamma_R \geq C) \quad (6.33)$$

where $\gamma_R = P_R|h_R|^2/\sigma_R^2$, $\bar{\gamma}_R = P_R\Omega_R/\sigma_R^2$, $F_{\gamma_R}(x) = 1 - \exp(-x/\bar{\gamma}_R)$, $\bar{F}_{\gamma_R}(\cdot) = 1 - F_{\gamma_R}(\cdot)$, and $C = P_R/G^2\sigma_R^2$. When $\gamma_R \geq C$ (i.e., $|h_R|^2 \geq 1/G^2$), the gain condition is not satisfied and power of the interference signal goes to infinity. Consequently, the received signal at D is purely random and $P(k \rightarrow \hat{k} | \gamma_R \geq C) = \frac{1}{N_T}$. So, PEP can be rewritten as

$$P(k \rightarrow \hat{k}) = \left[1 - \exp\left(\frac{-C}{\bar{\gamma}_R}\right) \right] \left[P(k \rightarrow \hat{k} | \gamma_R < C) + \frac{1}{N_T} \right] + \frac{1}{N_T}. \quad (6.34)$$

In order to calculate PEP, it is required to obtain the CPEP of OWFD-SSK-AF system for given $\mathbf{h}_{SR}$ and $\mathbf{h}_{RD}$ when $\gamma_R < C$ as follows

$$\begin{aligned}
P(k \rightarrow \hat{k} | \mathbf{h}_{SR}, \mathbf{h}_{RD}, \gamma_R < C) &= P\left\{\left\|\mathbf{y} - G\sqrt{P_S}h_{SR}^k\mathbf{h}_{RD}\right\|^2\right. \\
&\geq \left\|\mathbf{y} - G\sqrt{P_S}h_{SR}^{\hat{k}}\mathbf{h}_{RD}\right\|^2\Big\} \\
&= P\left\{G\sqrt{P_S}\|\mathbf{h}_{RD}\|^2\left(|h_{SR}^k|^2 - |h_{SR}^{\hat{k}}|^2\right)\right. \\
&\quad \left.- 2\Re\left[(h_{SR}^k - h_{SR}^{\hat{k}})\mathbf{y}^H\mathbf{h}_{RD}\right] \geq 0\right\} \\
&= P(\xi \geq 0)
\end{aligned} \tag{6.35}$$

where ξ is a Gaussian distributed random variable with the expected value and variance as follows

$$\begin{aligned}
E\{\xi | \mathbf{h}_{SR}, \mathbf{h}_{RD}, \gamma_R < C\} &= -G\sqrt{P_S}\|\mathbf{h}_{RD}\|^2|h_{SR}^k - h_{SR}^{\hat{k}}|^2, \\
\text{Var}\{\xi | \mathbf{h}_{SR}, \mathbf{h}_{RD}, \gamma_R < C\} &= 2|h_{SR}^k - h_{SR}^{\hat{k}}|^2\|\mathbf{h}_{RD}\|^2\left(\frac{P_R\|\mathbf{h}_{RD}\|^2}{D} + \sigma_D^2\right)
\end{aligned} \tag{6.36}$$

where $D = \frac{P_R}{G^2(\Omega_R P_R + \sigma_R^2)}$. Therefore, CPEP expression given in (6.35) is rewritten as

$$P(k \rightarrow \hat{k} | \mathbf{h}_{SR}, \mathbf{h}_{RD}, \gamma_R < C) = Q\left(\sqrt{\frac{\gamma_{SR}^{k \rightarrow \hat{k}} \gamma_{RD}}{\gamma_{RD} + D}}\right) \tag{6.37}$$

where $\gamma_{SR}^{k \rightarrow \hat{k}} = \frac{P_S|h_{SR}^k - h_{SR}^{\hat{k}}|^2}{2(\Omega_R P_R + \sigma_R^2)}$ and $\gamma_{RD} = \frac{P_R\|\mathbf{h}_{RD}\|^2}{\sigma_D^2}$. It is important to note that $\gamma_{SR}^{k \rightarrow \hat{k}}$ and γ_{RD} follow exponential and chi-square distributions, respectively. Therefore, their PDF expressions are given as $f_{\gamma_{SR}^{k \rightarrow \hat{k}}}(x) = \frac{\exp(-x/\bar{\gamma}_{SR})}{\bar{\gamma}_{SR}}$ and $f_{\gamma_{RD}}(x) = \frac{x^{N_R-1} \exp(-x/\bar{\gamma}_{RD})}{\Gamma(N_R) \bar{\gamma}_{RD}^{N_R}}$ where $\bar{\gamma}_{SR} = \frac{P_S \Omega_{SR}}{(\Omega_R P_R + \sigma_R^2)}$ and $\bar{\gamma}_{RD} = \frac{P_R \Omega_{RD}}{\sigma_D^2}$. By defining, $\gamma_{SRD}^{k \rightarrow \hat{k}} = \gamma_{SR}^{k \rightarrow \hat{k}} \gamma_{RD} / (\gamma_{RD} + D)$, integrating the above CPEP expression over the PDF of $\gamma_{SRD}^{k \rightarrow \hat{k}}$ and using [87, Eq. (32)], the required CPEP expression in (6.34) can be obtained as

$$\begin{aligned}
P(k \rightarrow \hat{k} | \gamma_R < C) &= \int_{x=0}^{\infty} Q(\sqrt{x}) f_{\gamma_{SRD}^{k \rightarrow \hat{k}}}(x) dx \\
&= \frac{1}{2\sqrt{2\pi}} \int_{x=0}^{\infty} \frac{\exp(-x/2)}{\sqrt{x}} F_{\gamma_{SRD}^{k \rightarrow \hat{k}}}(x) dx.
\end{aligned} \tag{6.38}$$

Then, the CDF of $\gamma_{SRD}^{k \rightarrow \hat{k}}$ can be obtained as follows

$$\begin{aligned}
F_{\gamma_{SRD}^{k \rightarrow \hat{k}}}(x) &= P\left(\frac{\gamma_{SR}^{k \rightarrow \hat{k}} \gamma_{RD}}{\gamma_{RD} + D} \leq x\right) \\
&= 1 - \frac{\exp(-x/\bar{\gamma}_{SR})}{\Gamma(N_R) \bar{\gamma}_{RD}^{N_R}} \int_{y=0}^{\infty} y^{N_R-1} \exp\left(-\frac{y}{\bar{\gamma}_{RD}} - \frac{x D}{y \bar{\gamma}_{SR}}\right) dy.
\end{aligned} \tag{6.39}$$

By applying [65, Eq. (3.471.9)], the above integration can be solved and the CDF of $\gamma_{SRD}^{k \rightarrow \hat{k}}$ is found as

$$F_{\gamma_{SRD}^{k \rightarrow \hat{k}}}(x) = 1 - \frac{2(x\mu)^{N_R/2}}{\Gamma(N_R) \exp(x/\bar{\gamma}_{SR})} K_{N_R}(2\sqrt{x\mu}) \quad (6.40)$$

where $\mu = D/(\bar{\gamma}_{SR}\bar{\gamma}_{RD})$. By substituting (6.40) into (6.38), we can write

$$P(k \rightarrow \hat{k} | \gamma_R < C) = \frac{1}{2} - \frac{\mu^{N_R/2}}{\sqrt{2\pi}\Gamma(N_R)} \int_0^\infty \frac{x^{(N_R-1)/2} K_{N_R}(2\sqrt{x\mu})}{\exp(\omega x)} dx \quad (6.41)$$

where $\omega = 0.5 + 1/\bar{\gamma}_{SR}$. By using [65, Eq. (6.643.3)], the required CPEP expression is derived as

$$P(k \rightarrow \hat{k} | \gamma_R < C) = \frac{1}{2} - \frac{(\mu/\omega)^{N_R/2} \Gamma(N_R + \frac{1}{2})}{2\sqrt{2\mu}\Gamma(N_R)} \exp\left(\frac{\mu}{2\omega}\right) W_{-\frac{N_R}{2}, \frac{N_R}{2}}\left(\frac{\mu}{\omega}\right) \quad (6.42)$$

where $W_{\cdot, \cdot}(\cdot)$ is the Whittaker function [65, Eq. (9.220.4)]. Finally, by combining (6.34) and (6.42), the PEP expression of the OWFD-SSK-AF system can be found as

$$P(k \rightarrow \hat{k}) = \left[1 - \exp\left(\frac{-C}{\bar{\gamma}_R}\right)\right] \left[\frac{1}{2} - \frac{(\mu/\omega)^{N_R/2} \Gamma(N_R + \frac{1}{2})}{2\sqrt{2\mu}\Gamma(N_R)} \exp\left(\frac{\mu}{2\omega}\right) W_{-\frac{N_R}{2}, \frac{N_R}{2}}\left(\frac{\mu}{\omega}\right) + \frac{1}{N_T}\right] + \frac{1}{N_T}. \quad (6.43)$$

At the last step, BER expression of the OWFD-SSK-AF system obtained via the union bound technique [42] is given by

$$P_b \leq \sum_{k=1}^{N_T} \sum_{\hat{k}=1}^{N_T} \frac{N(k, \hat{k}) P(k \rightarrow \hat{k})}{\log_2(N_T) N_T} \quad (6.44)$$

where $N(k, \hat{k})$ represents the number of bit errors associated with the corresponding pairwise error event. It is worth to mention that the above expression is equal to the exact BER of the OWFD-SSK-AF systems for $N_T = 2$ and it is the upper-bound BER expression for $N_T > 2$. Furthermore, the derived BER expression for the OWFD-SSK-AF systems are valid for the conventional OWHD-SSK-AF systems by letting $\beta = 0$ inside Ω_R .

6.2.3 Spectral efficiency of OWFD-SSK-AF and OWHD-SSK-AF

The transmission from source to destination can be accomplished in one time-slot in the OWFD-SSK-AF systems as expressed in the previous section. However, it can be completed in two time-slots with the conventional OWHD-SSK-AF systems, i.e., the

relay listens the signal transmitted by the source in the first time-slot and retransmits it to the destination in the second time-slot [40]. Thus, the spectral efficiency expressions of OWFD-SSK-AF and OWHD-SSK-AF are respectively given by

$$\begin{aligned}\eta_{FD} &= \log_2(N_T), \\ \eta_{HD} &= \frac{1}{2} \log_2(N_T),\end{aligned}\tag{6.45}$$

which means the number of transmit antennas in OWHD-SSK-AF should be as many as the square of the transmit antennas in OWFD-SSK-AF in order to achieve the same spectral efficiency.

6.2.4 Numerical results

Theoretical and Monte Carlo simulation results for the BER performance of the OWFD-SSK-AF systems over Rayleigh fading channels are represented throughout this section. The accuracy of the theoretical analysis is validated by different computer simulation scenarios. Moreover, BER performance of the OWFD-SSK-AF are compared with the conventional OWHD-SSK-AF systems. During the numerical analyses, we assume that $\Omega_{SR} = \Omega_{RD} = 1$ and $\sigma_D^2 = \sigma_R^2$.

Figure 6.8 demonstrates the BER performance of the OWFD-SSK-AF systems versus P_S/σ_D^2 for $\lambda \in \{0, 0.2, 0.4, 0.6, 0.8, 1\}$, where $\eta = 1$ [bit/s/Hz], $N_T = 2$, $N_R = 1$, $P_R = 4P_S$ and $\beta = 0$ dB. It is shown that the exact theoretical results obtained from (6.44) are perfectly matched with the computer simulations, which proves the correctness of our analysis. As seen from the curves, the BER values of the OWFD-SSK-AF systems takes lower values as long as the values of λ decrease, i.e., the efficiency of LI cancellation process increases. For instance, there is 5.5 dB difference for $\lambda = 0$ and $\lambda = 0.2$ at a BER value of 10^{-4} .

The BER performance of the OWFD-SSK-AF systems versus P_S/σ_D^2 for different values of N_R is plotted in Figure 6.9. Here, we assume that $\eta \in \{2, 4\}$ [bits/s/Hz], $N_T \in \{4, 16\}$, $P_R = P_S$, $\lambda = 0.25$ and $\beta = -10$ dB. It is observed that the computer simulation results are tightly upper-bounded by the theoretical results, especially for higher values of P_S/σ_D^2 . Additionally, we show that the increase of number of receive antennas at the destination (N_R) provides a coding gain. To illustrate, there is approximately 5.9 dB and 5.5 dB performance improvement when N_R is increased from 1 to 2 for $N_T = 4$ and $N_T = 16$ scenarios, respectively.

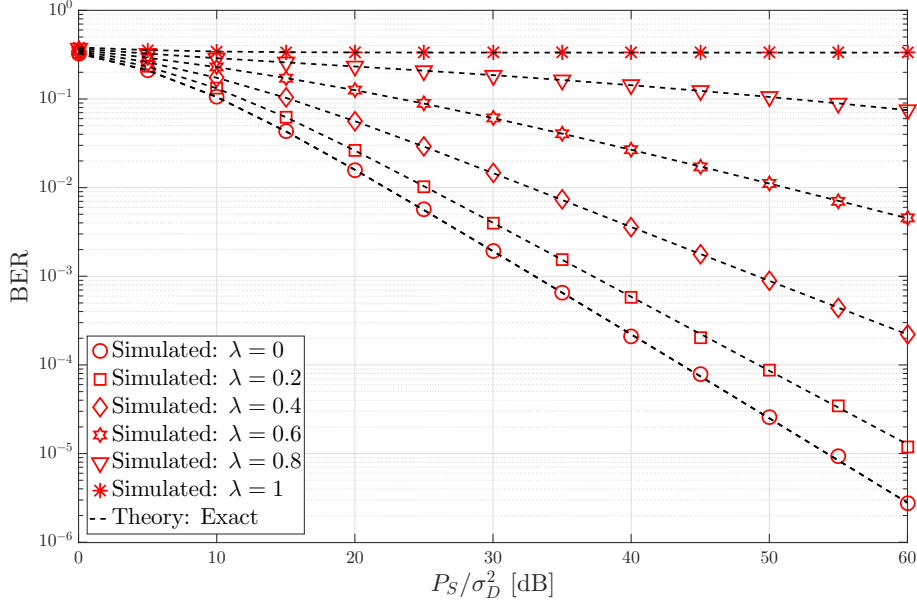


Figure 6.8 : BER performance of OWFD-SSK-AF versus P_S/σ_D^2 for different values of λ where $\eta = 1$ [bit/s/Hz], $N_T = 2$, $N_R = 1$, $P_R = 4P_S$ and $\beta = 0$ dB.

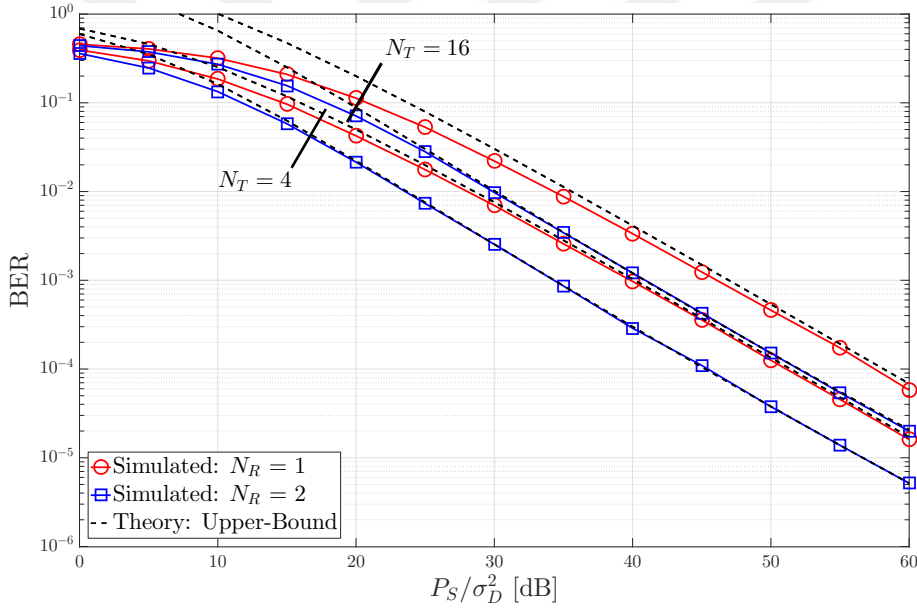


Figure 6.9 : BER performance of OWFD-SSK-AF versus P_S/σ_D^2 where $\eta \in \{2, 4\}$ [bits/s/Hz], $N_T \in \{4, 16\}$, $P_R = P_S$, $\lambda = 0.25$ and $\beta = -10$ dB.

In Figure 6.10, the BER performance comparison between OWFD-SSK-AF and OWHD-SSK-AF is represented for different spectral efficiencies as $\eta \in \{1, 2, 4\}$ [bits/s/Hz] when $N_R = 1$, $P_R = P_S$, $\lambda = 0.2$ and $\beta = 0$ dB. As explained in Section 6.2.3, the number of transmit antennas for both systems are different due to the FD and HD transmission. In order to ensure the same spectral efficiency values, $N_T \in \{2, 4, 16\}$ transmit antennas are employed for the OWFD-SSK-AF systems, while the number of transmit antennas for the OWHD-SSK-AF systems is set to $N_T \in \{4, 16, 256\}$. Although the BER performance of both systems are close to

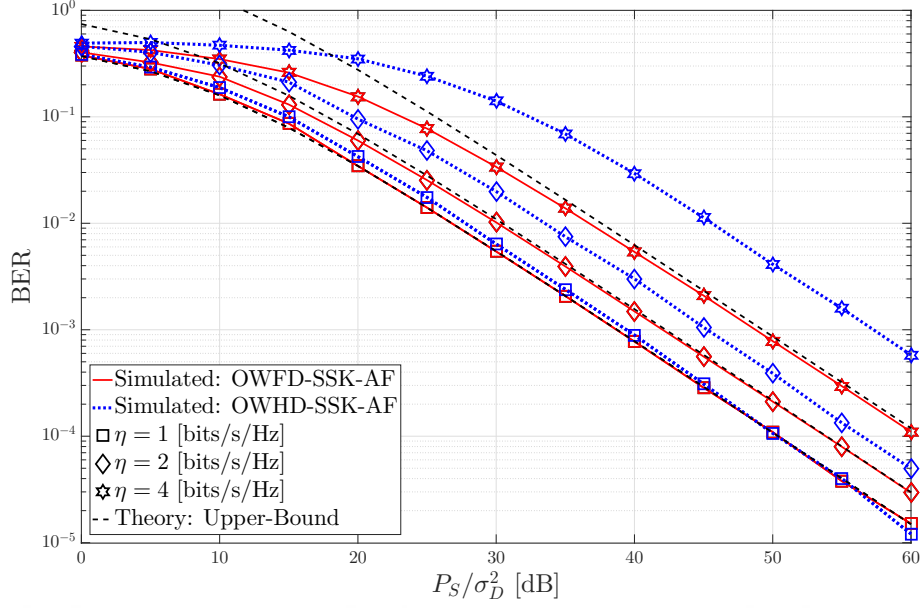


Figure 6.10 : BER performance of OWFD-SSK-AF and OWHD-SSK-AF versus P_S/σ_D^2 where $N_R = 2$, $P_R = P_S$, $\lambda = 0.2$ and $\beta = 0$ dB.

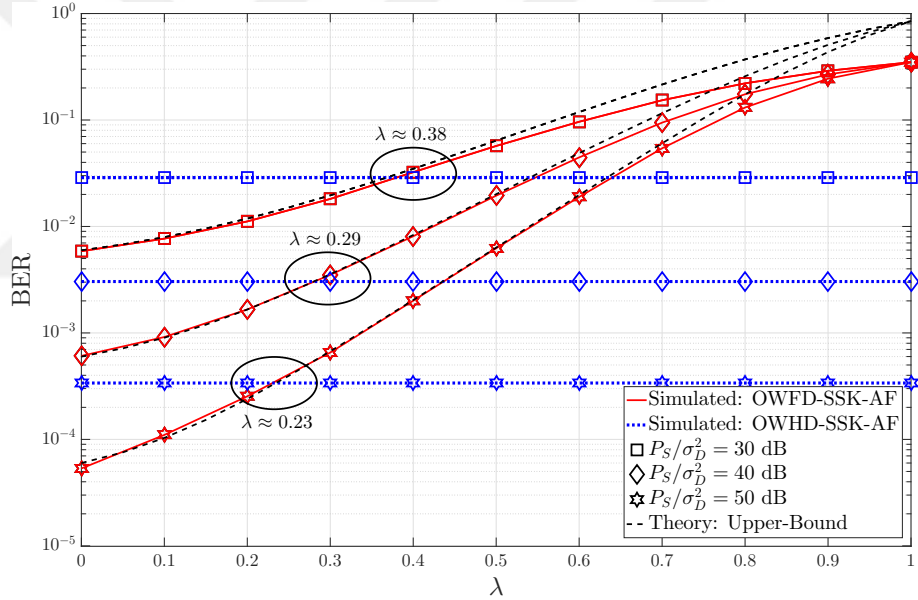


Figure 6.11 : BER performance of OWFD-SSK-AF and OWHD-SSK-AF versus λ where $\eta = 3$ [bits/s/Hz], $N_R = 2$, $P_R = P_S$ and $\beta = 0$ dB.

each other for $\eta = 1$ [bit/s/Hz], the OWFD-SSK-AF systems provide quite better performance compared to the OWHD-SSK-AF systems considering higher spectral efficiencies. For example, the performance improvement at a BER value of 10^{-3} for the increasing values of η are respectively found as 0.7 dB, 3.2 dB and 8.5 dB. Consequently, it is convenient to use the OWFD-SSK-AF to meet the higher spectral efficiency requirements.

In Figure 6.11, the BER performance of OWFD-SSK-AF and OWHD-SSK-AF systems is plotted versus λ for $P_S/\sigma_D^2 \in \{30, 40, 50\}$ dB where $\eta = 3$ [bits/s/Hz], $N_R = 2$, $P_R = P_S$ and $\beta = 0$ dB. As mentioned earlier, λ is the constant representing the quality of LI cancellation, so the performance of OWHD-SSK-AF systems is independent from λ . It is shown that OWFD-SSK-AF systems can provide better BER performance according to the quality of LI cancellations.

In summary, BER performances of both the TWFD-SM and OWFD-SSK-AF systems have been investigated over Rayleigh fading channels¹. The upper-bound expressions have been derived in closed-forms. The accuracy of those expressions has been validated via several Monte Carlo type computer simulation results. We have shown that the proposed FD systems can provide quite better error performance with respect to the conventional HD systems as long as the quality of LI cancellation improves and/or the requirement for the spectral efficiency increases.

¹Separate parts of this chapter are presented in *24th Int. Conf. Telecommun. (ICT 2017)* [88], *IEEE Int. Black Sea Conf. Commun. Netw. (IEEE BlackSeaCom 2017)* [89] and *25th Signal Process. Commun. Applicat. Conf. (SIU 2017)* [90].

7. FULL-DUPLEX SPATIAL MODULATION SYSTEMS WITH WIRELESS POWER TRANSFER

7.1 Simultaneous Wireless Information and Power Transfer

The wireless devices in the conventional communication systems use replaceable or rechargeable batteries, which induce the dependency on the limited battery lifetime and the necessity of external power supply. Furthermore, the battery replacement operations can be either impossible or dangerous under some circumstances (e.g., military conditions, chemical environment, etc.). In order to prolong the fixed operation time of the batteries, energy harvesting has gained great attention, where the major purpose is to scavenge the energy from natural resources (e.g., wind, solar, vibration, etc.). Although it achieves a significant breakthrough for the energy-efficient wireless networks, there could be wide fluctuations for the amount of harvested energy due to the unstable environmental conditions, which decreases the reliability of energy harvesting [91]. In addition to the natural resource, it shown that RF signals used for the information transmission can be also utilized for the energy harvesting, which is called as simultaneous wireless information and power transfer (SWIPT) proposed in [50]. At the beginning, it is assumed as the same RF signal can be entirely used for both information decoding and energy harvesting [50, 92]. However, this assumption is impractical because it is not possible to harvest energy after detection operation. Therefore, the information transmission and power transfer operations should be handled individually, which reveals a fundamental trade-off how much energy is used for information and harvesting. In [93], two common SWIPT protocols are proposed as power-splitting (PS) and time-switching (TS), which are expressed separately in the following sections.

The SWIPT technology is also very suitable for the cooperative networks in order to scavenge the energy at the relay node, which may not have an external power supply [54]. In the literature, OWFD and TWFD cooperative networks utilizing the SWIPT technique have been investigated in [55–58] and [59–61], respectively. The

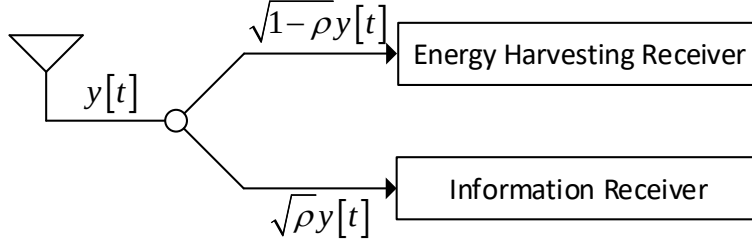


Figure 7.1 : Block diagram of the receiver in the PS protocol.

performance of both TS and PS protocols have been analyzed in the aforementioned works. The numerical results in [59] for the TWFD networks implies that the PS protocol provides higher throughput.

For the MIMO systems, the SM technique is combined with SWIPT in [94], where there is one source and one destination in the system (non-cooperative scheme). That system uses the PS protocol and an iterative PS algorithm is proposed so as to find the optimum ratio for the information transmission and energy harvesting.

7.1.1 Power-splitting

The block diagram of the receiver side in the PS protocol is represented in Figure 7.1, where the received signal is split into two streams. By assuming the total power of the received signal as $E\{|y[t]|^2\} = P$, receiver uses ρP for the information detection, at the same time, the energy harvesting is performed via the remainder of RF signals with the power of $(1 - \rho)P$, where $0 \leq \rho \leq 1$ [54, 93]. For the higher values of ρ , the amount of energy used for the information receiver increases; however, the harvested energy decreases. Therefore, it is required to find an optimum value for the ρ in the PS protocol.

7.1.2 Time-switching

In the TS protocol, the receiver employs switches for the energy harvesting and information receiver blocks as shown in Figure 7.2, where T is the symbol duration and α is the fraction of the symbol duration for information detection. The receiver first harvests its energy from the received RF signals during $(1 - \alpha)T$, then the remaining αT is utilized to recover the transmitted information [54, 93]. Similar to the PS protocol, we need to optimize the value of α in the TS protocol.

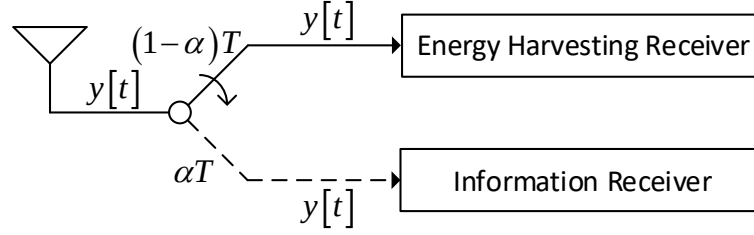


Figure 7.2 : Block diagram of the receiver in the TS protocol

7.2 SM with TWFD Fixed Gain AF Wireless Powered Relay Systems

The performance of TWFD-AF systems is comprehensively investigated for the single and multiple relay scenarios in Chapter 5. Then, FD transmission is combined with the SM technique for non-cooperative and cooperative networks in Chapter 6. In this section, a new two-way FD communication system consisting of two source and one relay node is proposed, where the sources employ SM and PLNC techniques and the relay uses fixed-gain AF relaying and the PS protocol for energy harvesting. Therefore, it is assumed that the assisted relay has no external power supply and the required energy for the information transmission is obtained from the RF signals. The proposed scheme is called as two-way full-duplex spatial modulation with amplify-and-forward and power-splitting (TWFD-SM-AF-PS). The BER performance of the TWFD-SM-AF-PS systems is analyzed over Rayleigh fading channels. We provide a closed-form upper-bound BER expression and the accuracy of analysis is verified various Monte Carlo type computer simulation results. Additionally, the derived BER expression is also valid for the HD counterpart of the proposed system as two-way half-duplex spatial modulation with amplify-and-forward and power-splitting (TWHD-SM-AF-PS). Hence, the BER performances of TWFD-SM-AF-PS and TWHD-SM-AF-PS are compared with each other. We show that TWFD-SM-AF-PS can provide quite better BER performance as long as the spectral efficiency increases and the quality and/or LI cancellations enhances.

7.2.1 System model

The system model of the proposed TWFD-SM-AF-PS and TWHD-SM-AF-PS systems composed of two sources ($S_i, i \in \{1, 2\}$) and one relay (R) is shown in Figure 7.3. We assume that there is no direct link between the sources and they can only communicate via the relay. The sources transmit their information to the relay by using the SM

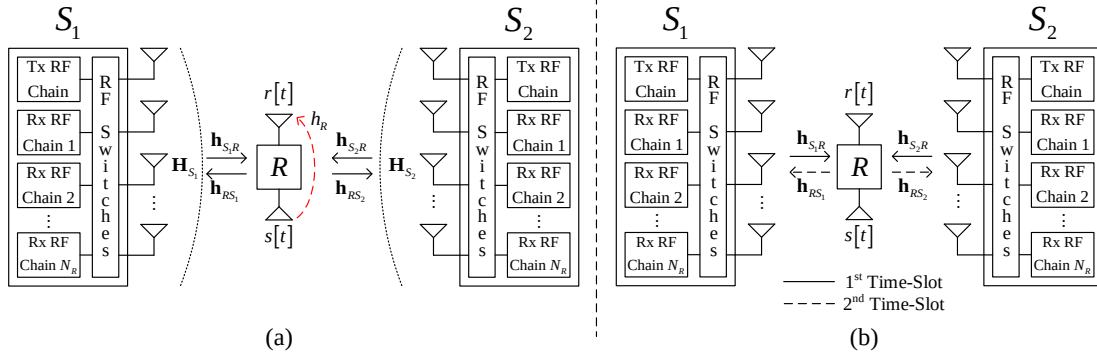


Figure 7.3 : The system models of (a) TWFD-SM-AF-PS and (b) TWHD-SM-AF-PS.

technique and M -PSK signal constellation. The relay using fixed-gain AF relaying has no external power supply and employs PS protocol in order to harvest energy from the received signal. For TWFD-SM-AF-PS, all nodes operate in FD mode and two-way transmission is accomplished in a single time-slot. So, the residual LI channels occur at both the sources and relay as in Figure 7.3(a). On the other hand, all nodes use the HD transmission and two time-slots are required in TWHD-SM-AF-PS as shown in Figure 7.3(b), where the sources transmit their information in the first time-slot and the relay retransmits the sum of the received signals to the sources in the second time-slot. Therefore, the residual LI channels do not exist in TWHD-SM-AF-PS.

Each source is equipped with N_T antennas that can either transmit or receive antenna by using the RF switches. Due to the SM transmission, only one antenna is utilized as a transmitter among N_T possible antennas in each symbol period, hence, it is required to place a single transmit RF chain at the source. The remaining $N_T - 1$ antennas can be used as receive antennas. Similar to Section 6.1, instead of using all of the remaining antennas to receive the signal, we place N_R receive RF chains ($1 \leq N_R \leq N_T - 1$) because it could be hard to employ that many receive RF chains for the increasing values of N_T . Therefore, there are $N_R + 1$ RF chains at each source node. As seen from the system model, the relay has one transmitter and one receiver. The block diagram of the relay node using the PS protocol is demonstrated in Figure 7.4. As mentioned above, the relay uses the fixed-gain AF relaying, and so output power of the relay varies with respect to the time as shown in Figure 4.2. Therefore, output power of the relay can be higher or lower than the harvested energy. An additional battery is placed into the energy harvesting block, which is either charged when the harvested energy is higher than the required at the power amplifier or discharged when

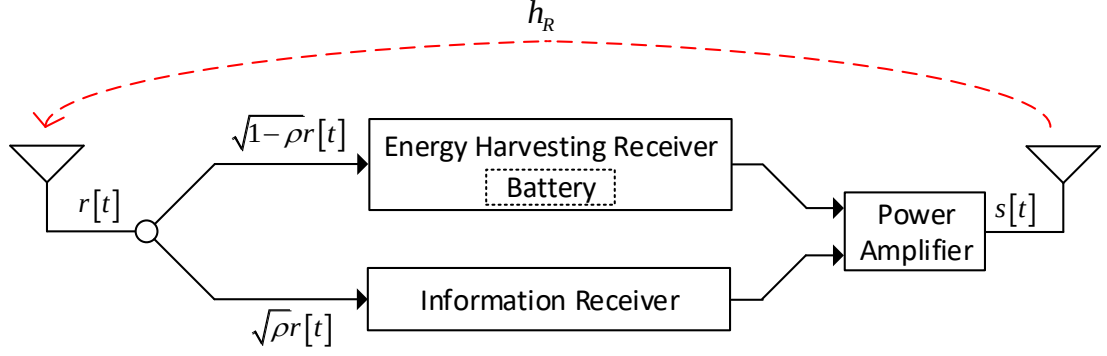


Figure 7.4 : The block diagram of relay using the PS protocol.

the harvested energy is insufficient for the power amplifier. We assume that the battery is fully charged at the beginning.

The received signal at the information receiver of the relay node is given by $r_{IR}[t] = \sqrt{\rho}r[t] + n_R[t]$, where

$$r[t] = \sqrt{P_S}\mathbf{h}_{S_1R}\mathbf{x}_1[t] + \sqrt{P_S}\mathbf{h}_{S_2R}\mathbf{x}_2[t] + h_Rs[t], \quad (7.1)$$

P_S is the transmission power of the sources, $\mathbf{x}_i[t]$ is the $N_T \times 1$ unit-power signal vector transmitted from S_i , $s[t]$ is the transmitted signal from the relay, $\mathbf{h}_{S_iR}$ is the $1 \times N_T$ channel coefficient vector between $S_i - R$ link, h_R is the residual LI channel coefficient at the relay and $n_R[t]$ is the complex Gaussian noise occurred at the information of the relay. The entries of $\mathbf{h}_{S_iR}$ are i.i.d. according to $\mathcal{CN}(0, \Omega_i)$. Similarly, h_R and $n_R[t]$ are distributed as $h_R \sim \mathcal{CN}(0, \Omega_R)$ and $n_R[t] \sim \mathcal{CN}(0, \sigma_R^2)$, respectively. As in Section 6.2, power of the residual LI channel is modeled as $\Omega_R = \beta P_R^{\lambda-1}$ where P_R is the average transmission power of the relay, β and λ are the constants representing the quality of LI cancellation ($0 \leq \lambda \leq 1$). It is important to note that $\beta = 0$ for the TWHD-SM-AF-PS systems. Therefore, remaining signal modeling used for the TWFD-SM-AF-PS systems is a more general model including the HD transmission scheme.

According to the SM technique, the transmitted signal vector at S_i can be written as

$$\mathbf{x}_i[t] = [0 \cdots 0 \quad \underbrace{x_i^{q_i[t]}}_{k_i[t]^{th} \text{ position}} \quad 0 \cdots 0] \quad (7.2)$$

where $k_i[t]$ is the selected transmit antenna index at S_i according to the first $\log_2(N_T)$ bits and $x_i^{q_i[t]}$ denotes the $q_i[t]$ symbol from the M -PSK constellation according to following $\log_2(M)$ bits. Hence, $\log_2(N_TM)$ bits are conveyed bidirectionally in the

proposed systems either in one or two time-slots. By using (7.1) and (7.2), the received signal at the information receiver can be rewritten as

$$r_{IR}[t] = \sqrt{P_S \rho} h_{S_1 R}^{k_1[t]} x_1^{q_1[t]} + \sqrt{P_S \rho} h_{S_2 R}^{k_2[t]} x_2^{q_2[t]} + \sqrt{\rho} h_{RS}[t] + n_R[t] \quad (7.3)$$

where $h_{S_i R}^{k_i[t]}$ is the $k_i[t]^{th}$ element of $\mathbf{h}_{S_i R}$. Afterwards, the transmitted signal from the relay is given by

$$s[t] = G r_{IR}[t-1] = G \sum_{u=0}^{\infty} (G \sqrt{\rho} h_R)^u q[t-u] \quad (7.4)$$

where $q[t] = \sum_{i=1}^2 \sqrt{P_S \rho} h_{S_i R}^{k_i[t-1]} x_i^{q_i[t-1]} + n_R[t-1]$ and G is the amplification factor. According to the fixed-gain AF relaying, the amplification factor is chosen as

$$G = \sqrt{\frac{P_R}{P_S \rho (\Omega_1 + \Omega_2) + P_R \rho \Omega_R + \sigma_R^2}}. \quad (7.5)$$

Although the instantaneous transmission power of the relay varies according to the time, by using (7.4) and (7.5), one can easily show that the average power consumption at the relay node is equal to $E\{|s[t]|^2\} = P_R$. The average transmission power of the relay is determined according to the harvested energy with the PS protocol. The energy harvesting operation is utilized by using $r_{EH}[t] = \sqrt{1 - \rho} r[t]$, which enters to the energy harvesting receiver. Then, the harvested energy during each symbol period is given by [59]

$$\begin{aligned} E_{EH}[t] &= \varphi(1 - \rho) \left(P_S |h_{S_1 R}^{k_1[t]}|^2 + P_S |h_{S_2 R}^{k_2[t]}|^2 + \frac{E_R[t-1]}{T} |h_R|^2 \right) T \\ &= \varphi(1 - \rho) T \sum_{u=0}^{\infty} (\varphi(1 - \rho) |h_R|^2)^u \left(P_S |h_{S_1 R}^{k_1[t-u]}|^2 + P_S |h_{S_2 R}^{k_2[t-u]}|^2 \right) \end{aligned} \quad (7.6)$$

where $0 \leq \varphi \leq 1$ is the energy conversion efficiency and T is the symbol period. By assuming $\varphi(1 - \rho) \Omega_R < 1$, the average harvested energy at the relay node can be obtained as

$$E\{E_{EH}[t]\} = \frac{\varphi(1 - \rho) P_S (\Omega_1 + \Omega_2) T}{1 - \varphi(1 - \rho) \Omega_R}. \quad (7.7)$$

Afterwards, the average transmission power of the relay can be found as

$$P_R = \frac{E\{E_{EH}[t]\}}{T} = \frac{\varphi(1 - \rho) P_S (\Omega_1 + \Omega_2)}{1 - \varphi(1 - \rho) \Omega_R} = \frac{\varphi(1 - \rho) P_S (\Omega_1 + \Omega_2)}{1 - \varphi(1 - \rho) \beta P_R^{\lambda-1}}. \quad (7.8)$$

To the best of our knowledge, unfortunately, it is not possible to find P_R analytically due to the above intractable expression. However, it can be numerically calculated by common mathematical software packages.

Finally, the received signal at S_i is given by

$$\begin{aligned} \mathbf{y}_i[t] = & \underbrace{\mathbf{h}_{RS_i} G \sqrt{P_S \rho} h_{S_j R}^{k_j[t-1]} x_j^{q_j[t-1]}}_{\text{Desired Signal}} + \underbrace{\mathbf{h}_{RS_i} G \sqrt{P_S \rho} h_{S_i R}^{k_i[t-1]} x_i^{q_i[t-1]}}_{\text{Removed by PLNC}} \\ & + \underbrace{\mathbf{h}_{RS_i} G \sqrt{\rho} h_{RS} [t-1] + \sqrt{P_S} \mathbf{h}_{S_i} x_i^{q_i[t-1]}}_{\text{Residual Loop-Interference}} + \underbrace{\mathbf{h}_{RS_i} G n_R [t-1] + \mathbf{n}_{S_i} [t]}_{\text{Noise}} \end{aligned} \quad (7.9)$$

where $j \in \{1, 2\}$, $j \neq i$, $\mathbf{h}_{RS_i}$ is the $N_R \times 1$ channel coefficient vector between the transmit antenna of relay and the selected receive antennas of S_i , $\mathbf{h}_{S_i}$ is the $N_R \times 1$ residual LI channel coefficient vector between the active transmit and receive antennas at S_i and $\mathbf{n}_{S_i} [t]$ is the complex Gaussian noise vector at S_i . The entries of $\mathbf{h}_{RS_i}$, $\mathbf{h}_{S_i}$ and $\mathbf{n}_{S_i} [t]$ are respectively distributed according to $\mathcal{CN}(0, \Omega_i)$, $\mathcal{CN}(0, \Omega_S)$ and $\mathcal{CN}(0, \sigma_S^2)$. Similar to the relay, we model power of the residual LI channel at S_i as $\Omega_S = \beta P_S^{\lambda-1}$. After applying the PLNC technique, the received signal can be written as follows

$$\hat{\mathbf{y}}_i[t] = \mathbf{y}_i[t] - \mathbf{h}_{RS_i} G \sqrt{P_S \rho} h_{S_i R}^{k_i[t-1]} x_i^{q_i[t-1]}. \quad (7.10)$$

For given $\mathbf{h}_{S_j R}$ and $\mathbf{h}_{RS_i}$, the expected value and variance of $\hat{\mathbf{y}}_i[t]$ can be given by

$$\begin{aligned} \mathbb{E} \{ \hat{\mathbf{y}}_i[t] | \mathbf{h}_{S_j R}, \mathbf{h}_{RS_i} \} &= \mathbf{h}_{RS_i} G \sqrt{P_S \rho} h_{S_j R}^{k_j[t-1]} x_j^{q_j[t-1]} \\ \text{Var} \{ \hat{\mathbf{y}}_i[t] | \mathbf{h}_{S_j R}, \mathbf{h}_{RS_i} \} &= \|\mathbf{h}_{RS_i}\|^2 G^2 \{ \rho \Omega_R P_R + \sigma_R^2 \} + P_S \Omega_S + \sigma_S^2 \end{aligned} \quad (7.11)$$

By omitting time index t , the optimal ML detection rule of the TWFD-SM-AF-PS and TWHD-SM-AF-PS systems can be written as

$$[\hat{k}_j, \hat{q}_j] = \arg \min_{\tilde{k}_j, \tilde{q}_j} \left\| \hat{\mathbf{y}}_i - \mathbf{h}_{RS_i} G \sqrt{P_S \rho} h_{S_j R}^{\tilde{k}_j} x_j^{\tilde{q}_j} \right\|^2 \quad (7.12)$$

where $\hat{k}_j$ and $\hat{q}_j$ are the determined transmit antenna and M -PSK symbol index.

7.2.2 BER analysis

The BER expression for the proposed systems is derived throughout this section by using the PEP approach. For given $\mathbf{h}_{S_j R}$ and $\mathbf{h}_{RS_i}$, the CPEP expression is given by

$$\begin{aligned} P(\mathbf{x}_j \rightarrow \hat{\mathbf{x}}_j | \mathbf{h}_{S_j R}, \mathbf{h}_{RS_i}) &= P \left\{ \left\| \hat{\mathbf{y}}_i - \mathbf{h}_{RS_i} G \sqrt{P_S \rho} h_{S_j R}^{k_j} x_j^{q_j} \right\|^2 \right. \\ &\quad \left. \geq \left\| \hat{\mathbf{y}}_i - \mathbf{h}_{RS_i} G \sqrt{P_S \rho} h_{S_j R}^{\hat{k}_j} x_j^{\hat{q}_j} \right\|^2 \right\} \\ &= P(\xi_i \geq 0) \end{aligned} \quad (7.13)$$

where $\hat{\mathbf{x}}_j$ is the decided symbol vector when $\mathbf{x}_j$ is transmitted. Afterwards, the expected value and variance of ξ_i for given $\mathbf{h}_{S_j R}$ and $\mathbf{h}_{RS_i}$ can be written as

$$\begin{aligned} \mathbb{E}[\xi_i | \mathbf{h}_{S_j R}, \mathbf{h}_{RS_i}] &= -G\sqrt{P_S \rho} \|\mathbf{h}_{RS_i}\|^2 \left| h_{S_j R}^{k_j} x_j^{q_j} - h_{S_j R}^{\hat{k}_j} x_j^{\hat{q}_j} \right|^2 \\ \text{Var}[\xi_i | \mathbf{h}_{S_j R}, \mathbf{h}_{RS_i}] &= 2\|\mathbf{h}_{RS_i}\|^2 \left| h_{S_j R}^{k_j} x_j^{q_j} - h_{S_j R}^{\hat{k}_j} x_j^{\hat{q}_j} \right|^2 \left(\frac{P_R \|\mathbf{h}_{RS_i}\|^2}{D} + P_S \Omega_S + \sigma_S^2 \right) \end{aligned} \quad (7.14)$$

where $D = \frac{P_R}{G^2(\rho P_R \Omega_R + \sigma_R^2)}$. By using the Gaussian distributed random variable ξ_i , the CPEP expression given in (7.13) can be reorganized as follows

$$P(\mathbf{x}_j \rightarrow \hat{\mathbf{x}}_j | \mathbf{h}_{S_j R}, \mathbf{h}_{RS_i}) = Q \left(\sqrt{\frac{\gamma_{S_j R}^{\mathbf{x}_j \rightarrow \hat{\mathbf{x}}_j} \gamma_{RS_i}}{\gamma_{RS_i} + D}} \right) \quad (7.15)$$

where $\gamma_{S_j R}^{\mathbf{x}_j \rightarrow \hat{\mathbf{x}}_j} = \frac{P_S |h_{S_j R}^{k_j} x_j^{q_j} - h_{S_j R}^{\hat{k}_j} x_j^{\hat{q}_j}|^2}{2(P_R \Omega_R + \sigma_R^2 / \rho)}$ and $\gamma_{RS_i} = \frac{P_R \|\mathbf{h}_{RS_i}\|^2}{P_S \Omega_S + \sigma_S^2}$. One can easily show that $\gamma_{S_j R}^{\mathbf{x}_j \rightarrow \hat{\mathbf{x}}_j}$ and γ_{RS_i} are respectively exponentially and chi-squared distributed random variables. Thus, their PDF expressions are given as $\gamma_{S_j R}^{\mathbf{x}_j \rightarrow \hat{\mathbf{x}}_j}(x) = \frac{\exp(-x/\bar{\gamma}_{S_j R})}{\bar{\gamma}_{S_j R}}$ and $f_{\gamma_{RS_i}}(x) = \frac{x^{N_R-1} \exp(-x/\bar{\gamma}_{RS_i})}{\Gamma(N_R) \bar{\gamma}_{RS_i}^{N_R}}$, where $\bar{\gamma}_{S_j R} = \frac{\sigma_{q_j}^2 P_S \Omega_j}{2(P_R \Omega_R + \sigma_R^2 / \rho)}$, $\bar{\gamma}_{RS_i} = \frac{P_R \Omega_i}{P_S \Omega_S + \sigma_S^2}$ and

$$\sigma_{q_j}^2 = \begin{cases} |x_j^{q_j} - x_j^{\hat{q}_j}|^2, & k_j = \hat{k}_j \\ 2, & k_j \neq \hat{k}_j. \end{cases} \quad (7.16)$$

By defining $\kappa_i = \frac{\gamma_{S_j R}^{\mathbf{x}_j \rightarrow \hat{\mathbf{x}}_j} \gamma_{RS_i}}{\gamma_{RS_i} + D}$ and integrating the CPEP over the PDF of κ_i , the PEP can be written as

$$P(\mathbf{x}_j \rightarrow \hat{\mathbf{x}}_j) = \int_{y=0}^{\infty} Q(\sqrt{y}) f_{\kappa_i}(y) dy = \frac{1}{2\sqrt{2\pi}} \int_{y=0}^{\infty} \frac{\exp(-y/2)}{\sqrt{y}} F_{\kappa_i}(y) dy. \quad (7.17)$$

By applying [65, Eq. (3.471.9)], the CDF of κ_i can be found as

$$\begin{aligned} F_{\kappa_i}(y) &= 1 - \frac{e^{-y/\bar{\gamma}_{S_j R}}}{\Gamma(N_R) \bar{\gamma}_{RS_i}^{N_R}} \int_{z=0}^{\infty} z^{N_R-1} \exp\left(-\frac{z}{\bar{\gamma}_{RS_i}} - \frac{yD}{z\bar{\gamma}_{S_j R}}\right) dz \\ &= 1 - \frac{2(y\mu_i)^{N_R/2}}{\Gamma(N_R) e^{y/\bar{\gamma}_{S_j R}}} K_{N_R}(2\sqrt{y\mu_i}) \end{aligned} \quad (7.18)$$

where $\mu_i = D/(\bar{\gamma}_{S_j R} \bar{\gamma}_{RS_i})$. By substituting (7.18) into (7.17) and using [65, Eq. (6.643.3)], the PEP expression of the proposed systems can be derived as

$$P(\mathbf{x}_j \rightarrow \hat{\mathbf{x}}_j) = \frac{1}{2} - \frac{e^{\frac{\mu_i}{2\omega_j}} \Gamma(N_R + \frac{1}{2})}{2\sqrt{2\mu_i} \Gamma(N_R)} \left(\frac{\mu_i}{\omega_j}\right)^{N_R/2} W_{-\frac{N_R}{2}, \frac{N_R}{2}}\left(\frac{\mu_i}{\omega_j}\right) \quad (7.19)$$

where $\omega_j = 0.5 + 1/\bar{\gamma}_{S_j R}$. Finally, the BER expression at S_i for the TWFD-SM-AF-PS and TWHD-SM-AF-PS systems is obtained in a closed-form by using the union-bound technique [42] and given by

$$P_{b,i} \leq \sum_{k_j=1}^{N_T} \sum_{q_j=1}^M \sum_{\hat{k}_j=1}^{N_T} \sum_{\hat{q}_j=1}^M \frac{N(\mathbf{x}_j \rightarrow \hat{\mathbf{x}}_j) P(\mathbf{x}_j \rightarrow \hat{\mathbf{x}}_j)}{\log_2(N_T M) N_T M} \quad (7.20)$$

where $N(\mathbf{x}_j \rightarrow \hat{\mathbf{x}}_j)$ represents the number of bit errors associated with the corresponding pairwise error event. The derived expression provides the exact BER of the proposed systems when $N_T = 2$ and $M = 1$, i.e., SSK technique is employed at the sources having two transmit antennas. For the higher values of N_T and M , it is an upper-bound BER expression. It is worthwhile to remark that we need to take $\beta = 0$ for TWHD-SM-AF-PS, where two-way transmission accomplished in two time-slots and the LI channels do not exist. On the other hand, the value of β varies according to the quality of LI cancellation for the TWFD-SM-AF-PS systems.

7.2.3 Spectral efficiency of TWFD-SM-AF-PS and TWHD-SM-AF-PS

In the TWFD-SM-AF-PS systems, both source nodes simultaneously exchange $\log_2(N_T M)$ bits in a single time-slots over the same frequency band. Therefore, a total of $2 \log_2(N_T M)$ bits are conveyed in each time-slot when the FD transmission adopted. However, the same amount of bits are conveyed in two time-slots when all nodes operate in HD mode. Hence, $\log_2(N_T M)$ bits are conveyed on average in each time-slot for the TWHD-SM-AF-PS systems. So, the spectral efficiency expressions of the FD and HD cases can be respectively given by

$$\begin{aligned} \eta_{FD} &= 2 \log_2(N_T M), \\ \eta_{HD} &= \log_2(N_T M). \end{aligned} \quad (7.21)$$

As understood from the above expressions, when HD transmission scheme is employed, either M -PSK constellation size or N_T should be increased in order to obtain the same spectral efficiency.

7.2.4 Numerical results

In this section, the theoretical and Monte Carlo type computer simulation results for the BER performance of the proposed TWFD-SM-AF-PS and TWHD-SM-AF-PS systems over Rayleigh fading channels are represented for various scenarios. The BER curves

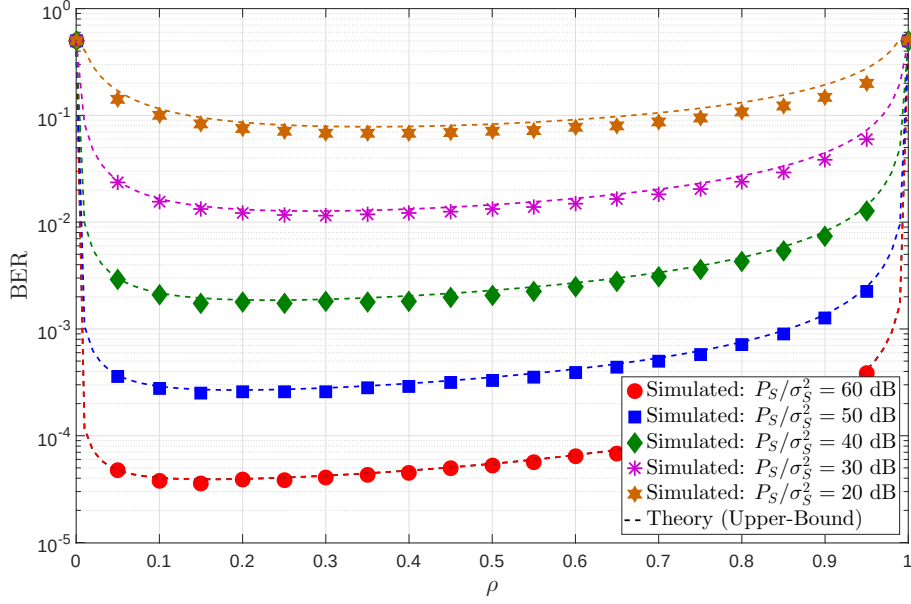


Figure 7.5 : BER performance of TWFD-SM-AF-PS versus ρ for different values of P_S/σ_S^2 , where $\eta = 4$ [bits/s/Hz], $N_T = 2$ and $N_R = 1$.

are generated according to the received signal at S_1 (i.e., $P_{b,1}$). In order to make a fair comparison between the FD and HD transmission, the total number of available RF chains are taken as the same. During the numerical analyses, we assume that the power of the source-relay links is equal to the unity as $\Omega_1 = \Omega_2 = 1$, all the noise terms have the same variance $\sigma_R^2 = \sigma_S^2$ and the energy conversion operation is handled without any loss ($\varphi = 1$) as in [53]. Additionally, the LI cancellation parameters are chosen as $\beta = 0.1$ and $\lambda = 0.2$, unless otherwise stated.

In Figure 7.5, the BER performance of the TWFD-SM-AF-PS systems with BPSK is plotted versus different values of ρ for $P_S/\sigma_S^2 \in \{20, 30, 40, 50, 60\}$ dB where $\eta = 4$ [bits/s/Hz], $N_T = 2$ and $N_R = 1$. It is demonstrated that the computer simulation results are tightly upper-bounded by the theoretical results, especially for the higher values of P_S/σ_S^2 , which proves the accuracy of the derived BER expression for TWFD-SM-AF-PS. We observe that the optimum value for the power splitting ratio providing the lowest BER result changes for the different value of P_S/σ_S^2 . To illustrate, when the power splitting ratio equals $\rho \in \{0.35, 0.28, 0.23, 0.19, 0.16\}$, the minimum BER results are obtained for $P_S/\sigma_S^2 \in \{20, 30, 40, 50, 60\}$ dB, respectively.

The BER performance of the TWHD-SM-AF-PS systems with BPSK versus different values of ρ for $P_S/\sigma_S^2 \in \{20, 30, 40, 50, 60\}$ dB is demonstrated in Figure 7.6 . Here, we assume that $\eta = 3$ [bits/s/Hz], $N_T = 4$ and $N_R = 3$. It is seen that the computer

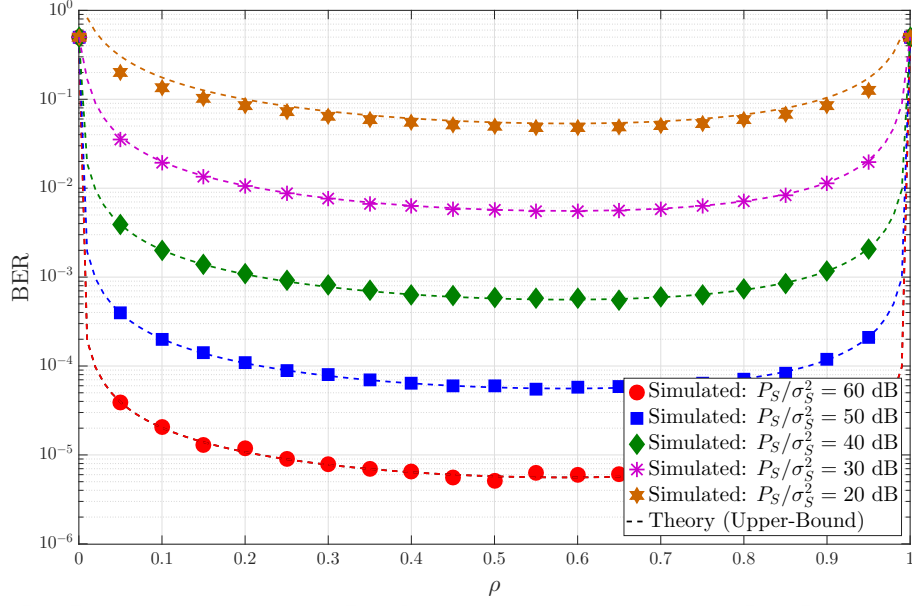


Figure 7.6 : BER performance of TWHD-SM-AF-PS versus ρ for different values of P_S/σ_S^2 , where $\eta = 3$ [bits/s/Hz], $N_T = 4$ and $N_R = 3$.

results for TWHD-SM-AF-PS are also tightly upper-bounded by theoretical curves validating the accuracy of analysis. In contrast to the FD transmission, the optimum power splitting ratio is constant for all P_S/σ_S^2 when HD transmission is utilized in this scenario. For example, $\rho = 0.59$ provides the lowest BER results independent from the value of P_S/σ_S^2 .

Figure 7.7 represents the BER performance comparison of the TWFD-SM-AF-PS and TWHD-SM-AF-PS systems versus P_S/σ_S^2 , where $\eta = 2$ [bits/s/Hz], $N_T = 2$, $N_R = 1$ and $\rho = 0.22$. For the FD transmission scheme, the BER curves are produced for the different values of λ and the required spectral efficiency is ensured by setting $M = 1$, i.e., a constant known signal (such as +1) is transmitted from the sources. On the other hand, BPSK signals are used for the HD transmission scheme. It is shown that TWFD-SM-AF-PS systems can provide better performance as long as the quality of LI cancellation increases. For instance, at a BER value of 10^{-3} , 2.2 dB and 0.9 dB performance improvement can be achieved with respect to TWHD-SM-AF-PS when $\lambda = 0$ and $\lambda = 0.2$, respectively. However, the HD transmission scheme performs better when λ and/or P_S/σ_S^2 increases.

The BER performance comparison between the TWFD-SM-AF-PS and TWHD-SM-AF-PS systems is shown in Figure 7.8, where $\eta = 8$ [bits/s/Hz], $N_T = 4$, $N_R = 2$ and $\rho = 0.5$. Here, QPSK and 64-PSK signals are respectively

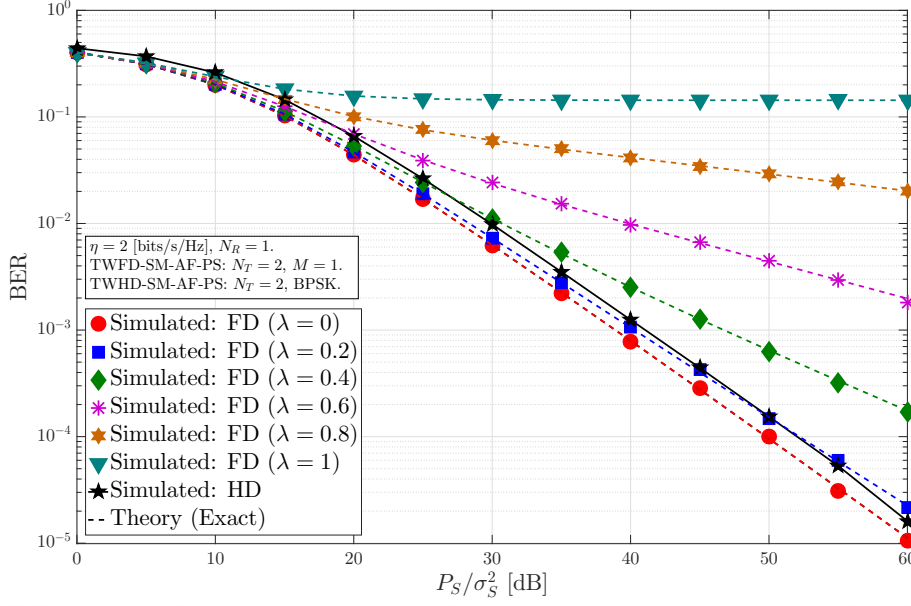


Figure 7.7 : BER performance of TWFD-SM-AF-PS and TWHD-SM-AF-PS versus P_S/σ_S^2 , where $\eta = 2$ [bits/s/Hz], $N_T = 2$, $N_R = 1$ and $\rho = 0.22$.

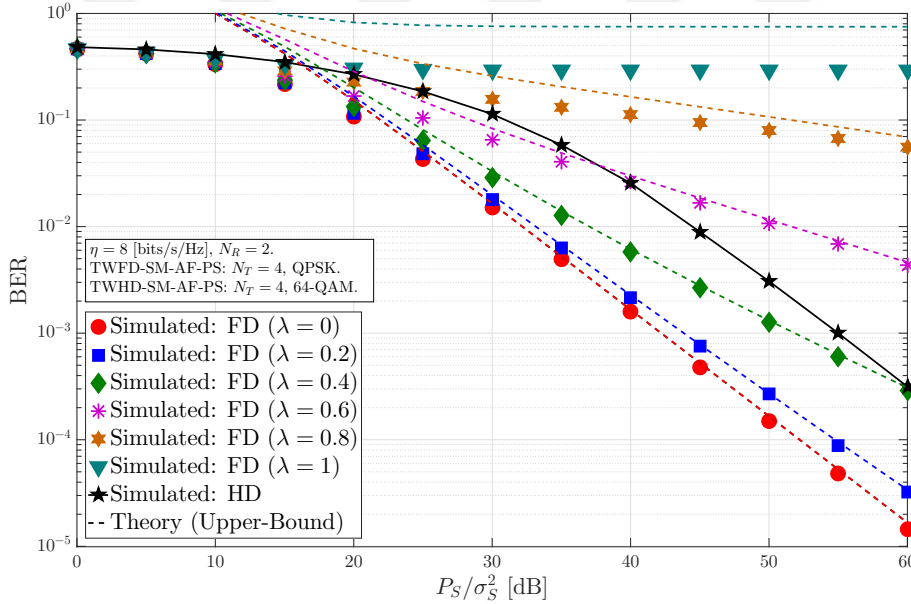


Figure 7.8 : BER performance of TWFD-SM-AF-PS and TWHD-SM-AF-PS versus P_S/σ_S^2 , where $\eta = 8$ [bits/s/Hz], $N_T = 4$, $N_R = 2$ and $\rho = 0.5$.

employed for the FD and HD transmission schemes for the desired spectral efficiency. In comparison to the previous simulation scenario, the performance improvement of TWFD-SM-AF-PS increases. For example, when $\lambda \in \{0, 0.2, 0.4\}$, the performance gain at a BER value of 10^{-3} rises to 12.8 dB, 11.2 dB and 3.1 dB, respectively. Thus, TWFD-SM-AF-PS can perform better than its HD counterpart as long as the requirement for the spectral efficiency increases.

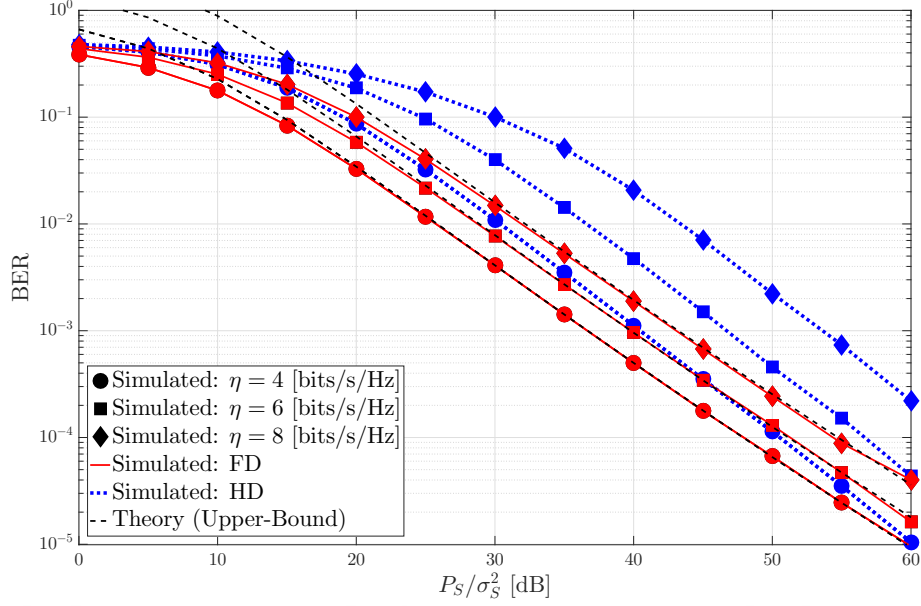


Figure 7.9 : BER performance of TWFD-SM-AF-PS and TWHD-SM-AF-PS versus P_S/σ_S^2 for $\eta \in \{4, 6, 8\}$ [bits/s/Hz], where $N_T = 4$, $N_R = 3$ and $\rho = 0.5$.

In Figure 7.9, the BER performance comparison of the TWFD-SM-AF-PS and TWHD-SM-AF-PS systems versus P_S/σ_S^2 is shown for $\eta \in \{4, 6, 8\}$ [bits/s/Hz], where $N_T = 4$, $N_R = 3$ and $\rho = 0.5$. Moreover, we assume that $\lambda = 0.25$. When two-way communication accomplished in a single time-slot with the FD transmission, the sources use a constant known, BPSK and QPSK signals for the required spectral efficiency values. Meanwhile, QPSK, 16-PSK and 64-PSK signals are utilized for the HD transmission. As seen, TWFD-SM-AF-PS provides quite better BER performance, especially for the higher values of η . For instance, the BER performance improvements of the FD case for the increasing values of spectral efficiency are respectively 3.8 dB, 6.9 dB and 10.5 dB at a BER value of 10^{-3} . Therefore, it is convenient to use the TWFD-SM-AF-PS systems for the higher spectral efficiency requirements.

The BER performance of the TWFD-SM-AF-PS and TWHD-SM-AF-PS systems has been investigated throughout this chapter. We have derived a unified upper-bound BER expression in a closed-form. The accuracy of the theoretical analysis has been validated via various simulation scenarios for both of the proposed transmission schemes. It has been shown that the BER performance of the TWFD-SM-AF-PS systems is highly relevant to the efficiency of LI cancellation operation. Furthermore, we have demonstrated that TWFD-SM-AF-PS can perform better as long as the LI cancellation improves and/or the spectral efficiency requirement increases.



8. CONCLUSIONS

In this thesis, new FD wireless communication systems allowing simultaneous transmission and reception operations over the same frequency band have been designed and their performance under the residual LI has been investigated for different wireless environments, e.g., Rayleigh and Nakagami- m fading channels. During the performance analyses, the outage probability or BER expressions have been derived for the proposed systems. The accuracy of the theoretical analyses has been verified separately via extensive Monte Carlo type computer simulations for various scenarios.

At the beginning, the outage performance of the OWFD-AF systems has been investigated, where only the relay node operates in FD mode and one-way transmission is completed in a single time-slot. We have first proposed the fixed-gain OWFD-AF systems and the exact outage probability expression for Rayleigh fading channels has been derived in a single-integral form. The numerical results have revealed that it can provide quite better performance with respect to the variable-gain case. Then, the variable-gain and fixed-gain OWFD-AF systems have been analyzed over Nakagami- m fading channels. For each system, the exact, lower-bound and asymptotic outage probability expressions have been obtained. Afterwards, five relay selection policies have been proposed for the variable-gain OWFD-AF-RS systems over Nakagami- m fading channels in order to improve the outage performance of the variable-gain OWFD-AF systems. The exact, lower-bound and asymptotic outage probability expressions have been also derived for each relay selection policy. It has been shown that the outage performance of OWFD-AF-RS can be improved as long as the number of available relays for the relay selection increases.

Secondly, we have analyzed the outage performance of both the variable-gain TWFD-AF and TWFD-AF-RS systems over Nakagami- m fading channels, where PLNC technique is employed, all nodes operate in FD mode and two-way transmission is accomplished in a single time-slot. For each system, the exact outage probability expressions have been derived in a single-integral form. Furthermore, the tight

lower-bound and asymptotic outage probability expressions have been found in the closed-forms. It has been illustrated that outage performance of the TWFD-AF-RS systems with max-min relay selection policy can be enhanced by either increasing the number of available relay nodes or improving the quality of LI cancellation.

Thirdly, the FD communications has been combined with the SM technique for the non-cooperative and cooperative networks. We have first investigated the error performance of the TWFD-SM systems under both P-CSI and I-CSI over Rayleigh fading channels. The numerical results generated via the closed-form upper-bound BER expression have shown that TWFD-SM can outperform the conventional TWHD-SM systems. Then, the BER performance of the OWFD-SSK-AF systems has been analyzed over Rayleigh fading channels and the upper-bound BER expression has been derived in a closed-form. The performance improvement of OWFD-SSK-AF has been represented with respect to the conventional OWHD-SSK-AF systems.

For the final work of this thesis, new wireless-powered cooperative systems named as TWFD-SM-AF-PS and TWHD-SM-AF-PS have been proposed, where the SM and PLNC techniques are employed at the sources and the relay node utilizes the AF relaying and the PS protocol for SWIPT. We have found a unified upper-bound BER expression in a closed-form for the FD and HD transmission schemes. By means of the numerical analyses, it has been shown that TWFD-SM-AF-PS can provide better BER performance when the quality of LI cancellation process improves and/or the spectral efficiency requirement increases.

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PUBLICATIONS, PRESENTATIONS AND PATENTS ON THE THESIS:

International Journal Papers

- **Koç, A.,** Altunbaş, İ. and Yongaçoğlu A. (2016) Outage performance of fixed-gain and variable-gain AF full-duplex relaying in non-identical Nakagami- m fading channels, accepted for publication in *EURASIP J. Wireless Commun. Netw.*
- **Koç, A.,** Altunbaş, İ. and Yongaçoğlu A. (2016) Relay selection in two-way full-duplex relay networks over Nakagami- m fading channels, Submitted to *Ann. Telecommun.*

International Conference Papers

- **Koç, A.,** Altunbaş, İ. and Başar, E. (2017) Space shift keying with full duplex amplify and forward relaying, *IEEE Int. Black Sea Conf. Commun. Netw.*
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- **Koç, A.,** Altunbaş, İ. and Yongaçoğlu A. (2016) Outage probability of two-way full-duplex AF relay systems over Nakagami- m fading channels, *IEEE 84th Veh. Technol. Conf. (VTC Fall)*, pp.1-6.
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National Journal Papers

- **Koç, A.** and Altunbaş, İ. (2016) Performance of full-duplex AF relay selection systems over Nakagami- m channels (In Turkish), Submitted to *Dokuz Eylül Univ. Eng. Faculty J. Sci. Eng.*.

National Conference Papers

- **Koç, A.**, Altunbaş, İ. and Başar, E. (2017) Performance analysis of full-duplex spatial-modulated communication systems (In Turkish), *25th Signal Process. Commun. Applicat. Conf. (SIU 2017)*.
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OTHER PUBLICATIONS:

International Conference Papers

- **Koç, A.**, Altunbaş, İ., and Yaman, B. (2017) Unified outage performance analysis of two-way/one-way full-duplex/half-duplex fixed-gain AF relay systems, *24th Int. Conf. Telecommun. (ICT 2017)*.

PROJECTS:

- **2015–2018:** The Scientific and Technological Research Council of Turkey, 1001-114E607: *Cooperative Spatial Modulation Systems*.
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AWARDS AND SCHOLARSHIPS:

- **2017–2020:** McGill Engineering Doctoral Award (MEDA), McGill University.
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- **2015:** Ranked 3rd among all undergraduate students in the Faculty of Electrical and Electronics.
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CERTIFICATES:

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