

T.C.
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MÜHENDİSLİK VE FEN BİLİMLERİ ENSTİTÜSÜ

PERFORMANCE OF MULTI-ANTENNA
DIVERSITY SCHEMES IN FADING
CHANNELS IN THE PRESENCE OF
PRACTICAL IMPAIRMENTS

AHMET FARUK COŞKUN
DOKTORA TEZİ
ELEKTRONİK MÜHENDİSLİĞİ ANABİLİM DALI

GEBZE

2013

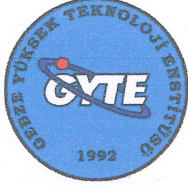
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DOÇ. DR. OĞUZ KUCUR

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2013



**GEBZE YÜKSEK
TEKNOLOJİ ENSTİTÜSÜ**

DOKTORA JÜRİ ONAY FORMU

GYTE Mühendislik ve Fen Bilimleri Enstitüsü Yönetim Kurulu'nun
...../...../..... tarih ve/...../..... sayılı kararıyla oluşturulan jüri
tarafından 04/11/2013 tarihinde tez savunma sınavı yapılan Ahmet Faruk
COŞKUN'un tez çalışması Elektronik Mühendisliği Anabilim Dalında DOKTORA
tezi olarak kabul edilmiştir.

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GYTE Mühendislik ve Fen Bilimleri Enstitüsü Yönetim Kurulu'nun
...../...../..... tarih ve/...../..... sayılı kararı.

İMZA/MÜHÜR

ÖZET

Yeni nesil telsiz haberleşme sistemleri, artan kullanıcı sayısı ile ortaya çıkan veri trafiği yükünü, makul devre karmaşıklığı, bant genişliği ve güç tüketimi kısıtları altında ve yüksek güvenilirlikle taşıyabilecek telsiz tasarımlara ihtiyaç duymaktadır. Sönümlemeli kanallarda, gerek sistem kapasitesi ve iletim hızı, gerekse haberleşme güvenirliliği açısından sağladıkları büyük iyileştirmeler, anten çeşitleme tekniklerini ilgili literatürde ve dolayısıyla yeni nesil telsiz haberleşme sistemlerinin tasarımında ön plana çıkartmıştır. Literatürde, birçok çeşitleme tekniği için sunulan performans analizleri, kanal kestiriminin sorunsuz şekilde yapıldığı, (tam veya kısmi) kanal durum bilgisinin (channel state information (CSI)) verici birime geri-besleme hattında hiçbir bozulma ve gecikme olmaksızın iletildiği ideal durumlarda geçerlidir. Çeşitleme tekniklerinin kanal kestirim hatası (channel estimation error (CEE)), geribesleme gecikmesi (feedback delay (FD)) ve geribesleme hatası (feedback error (FE)) gibi pratik uyumsuzluklar varlığındaki performans analizleri, bu sistemlerin gerçek uygulamalardaki performanslarının gözlenebilmesi, telsiz haberleşme standartları için öneminin ve geçerliliğinin incelenebilmesi açısından çok büyük önem taşımaktadır. Bu tez kapsamında, verici ve/veya alıcı biriminde hibrit anten çeşitleme tekniklerini kullanan telsiz haberleşme sistemlerinin, CEE, FD ve FE gibi pratik uyumsuzlukların etkisi altında, Nakagami- m sönümlemeli kanaldaki performanslarına ilişkin ölçütlerin analizleri yapılmış ve elde edilen teorik performans ifadelerin Monte Carlo simülasyonları kullanılarak doğrulanması sağlanmıştır. Tez kapsamında yapılan çalışma neticesinde, pratik uyumsuzluklar durumunda, çeşitleme tekniklerinin farklı karakteristikler sergileyen performans düşüşleri gösterdiği görülmüştür. Söz konusu performans düşüşleri her bir çeşitleme tekniği için etraflıca incelenerek, çeşitlemeli yapıların birbirleri ile karşılaştırılmaları sağlanmış ve sonrasında bazı çeşitlemeli yapılara ilişkin geliştirilen modifiye çeşitleme yapılarının pratik uyumsuzluklar durumunda, yaygın (klasik) sistemlere göre performans iyileşmeleri sağladığı gösterilmiştir.

Anahtar Kelimeler: Çok-yollu sönümleme, Anten çeşitlemesi, Kanal kestirim hatası, Geribesleme gecikmesi, Geribesleme hatası, Nakagami- m sönümlemeli kanallar.

SUMMARY

Next generation wireless communications systems have been requiring wireless system designs that can support the data traffic, which is brought along with the increasing number of clients, by enabling higher data rates with moderate hardware complexity, bandwidth and power consumption constraints. The substantial improvements in both the system capacity/transmission rates and the communications reliability that antenna diversity techniques provided in fading channels made them prominent in the related literature and therefore in the designs of wireless communications systems. The performance investigations on many diversity schemes that have been shown to be efficient at combating the fading effect are valid under the ideal conditions in which the channel estimation is perfect, the channel state information (CSI) (partially or fully) is sent to the transmitter with no delay and feedback error. Hence, the performance analyses of these diversity schemes in the presence of practical impairments such as channel estimation error (CEE), feedback delay (FD) and feedback error (FE) are of great importance for examining the practical performance of the diversity schemes that would show their suitability and feasibility to the real-world wireless communications applications and standards. Within the scope of this thesis, the investigation of many performance metrics related to the performances of wireless communications systems that employ hybrid diversity schemes in their transmit and/or receive ends in Nakagami- m fading channels is performed for practical impairments such as CEE, FD and FE, and the theoretical results have been validated by Monte Carlo simulations. As the result of the investigation provided in this thesis, the different performance degradation characteristics of diversity schemes are observed. By extensively examining the performance degradations of each hybrid diversity scheme, comparisons between diversity schemes are presented. Also, the modified versions of some hybrid diversity schemes which are proposed against practical conditions are shown to bring along with enhancements when compared to the conventional schemes in the presence of practical impairments.

Keywords: Multi-path fading, Multi-antenna diversity, Channel estimation error, Feedback delay, Feedback error, Nakagami- m fading channels.

ACKNOWLEDGMENT

I owe a debt of gratitude to my advisor Assoc. Prof. Oğuz Kucur, for sharing his experiences, knowledge, most-precious time and self-devoted leading through my undergraduate, graduate and doctorate education,

to Dr. Ahmet Yılmaz for his significant help at interpreting the problems,

to the Scientific and Technological Research Council of Turkey (TÜBİTAK) for supporting the work presented in this thesis in part under Grant EEEAG-112E010,

and to my family for their patience and their consistent support during my working and graduation period.

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LIST OF SYMBOLS AND ABBREVIATIONS

<u>Symbols and</u>	<u>Definitions</u>
<u>Abbreviations</u>	
ACG	: Asymptotic Coding Gain
ADO	: Asymptotic Diversity Order
AoF	: Amount of Fading
ARQ	: Automatic Repeat Request
AWGN	: Additive White Gaussian Noise
BER	: Bit Error Rate
BSC	: Binary Symmetric Channel
BPSK	: Binary Phase Shift Keying
CBFSK	: Coherent Binary Frequency Shift Keying
CCI	: Co-Channel Interference
CDF	: Cumulative Distribution Function
CEE	: Channel Estimation Error
CEP	: Conditional Error Probability
CF	: Correct Feedback
CSI	: Channel State Information
DBPSK	: Differential Binary Phase Shift Keying
EDO	: Effective Diversity Order
EF	: Erroneous Feedback
EGC	: Equal-Gain Combining
FD	: Feedback Delay
FE	: Feedback Error
FF	: Feedback Failure
FO	: Frequency Offset
FQE	: Feedback Quantization Error
GSC	: Generalized Selection Combining
HDS	: Hybrid Diversity Scheme
HSPA	: High-Speed Packet Access
ICEF	: Ideal Channel Estimation and Feedback
IEEE	: Institute of Electrical and Electronics Engineers
JTRAS	: Joint Transmit and Receive Antenna Selection

LoS	: Line-Of-Sight
LTE	: Long-Term Evolution
M-PAM	: M-ary Pulse Amplitude Modulation
M-PSK	: M-ary Phase Shift Keying
M-QAM	: M-ary Quadrature Amplitude Modulation
MGF	: Moment-Generating Function
MIMO	: Multi-Input Multi-Output
MISO	: Multi-Input Single-Output
MRC	: Maximal-Ratio Combining
MRT	: Maximal-Ratio Transmission
NCBFSK	: Non-Coherent Binary Frequency Shift Keying
OP	: Outage Probability
OSTBC	: Orthogonal Space-Time Block Coding
OSTTC	: Orthogonal Space-Time Trellis Coding
PA	: Power Allocation
PDF	: Probability Density Function
PEP	: Pair-wise Error Probability
QPSK	: Quadrature Phase Shift Keying
RAS	: Receive Antenna Selection
RF	: Radio-Frequency
SISO	: Single-Input Single-Output
SC	: Selection Combining
SER	: Symbol Error Rate
TAS	: Transmit Antenna Selection
TASC	: Transmit Antenna Subset Combination
TB	: Transmit Beamforming
W-CDMA	: Wide-Band Code Division Multiple Access
WCS	: Wireless Communications Systems
Wi-Fi	: Wireless Fidelity
WiMax	: Worldwide Interoperability for Microwave Access
3GPP	: Third Generation Partnership Project

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1. INTRODUCTION

Wireless communications systems (WCSs), especially those considered through the next generation communications standards, have been requiring wireless system designs that can support the data traffic, which is brought along with the increasing number of clients, by enabling higher data rates with moderate hardware complexity, bandwidth, and power consumption constraints. But nonetheless, the propagation environment that hosts the wireless communications faces many physical challenges such as shadowing, multi-path reflection, and scattering caused by the presence of high buildings, trees, hills, and other structures. Also, another degrading complication of wireless transmission is the relative motion of the transmitter and receiver. All these difficulties cause the time and frequency characteristics of the wireless propagation channel to vary in a continuous and random manner. These changes in the impulse response of the wireless channel result in deteriorations at the envelope of the received signals that might considerably degrade the overall system performance.

For the WCSs that face the multi-path fading effect caused by the aforementioned challenges, many transmission/reception techniques are developed and proposed in order to reduce and minimize the degrading effects of multi-path fading on the system performance. The main principle of the proposed techniques is to provide the replicas of the transmitted signal at the receiver decoder input in a way that all have faced independent fading effects. By virtue of these techniques, the deterioration on the total received signal envelope is minimized which brings along with considerable performance enhancements. The substantial improvements in both the system capacity/transmission rates and the communications reliability that antenna diversity techniques provided in fading channels made them prominent in the related literature and therefore in the designs of WCSs.

Antenna diversity schemes that are constructed by employing multiple antennas at the transmitter and/or the receiver ends are often called multi-input multi-output (MIMO) communications systems, and the diversity gathered by the usage of these schemes is called antenna (or spatial) diversity. Recent studies have been made on MIMO communications and antenna diversity and their performance-enhancements that encouraged the consideration of these schemes in wireless communications

standards and protocols such as 3GPP LTE [1], WiMax [2], HSPA+ [1] and IEEE 802.11n (Wi-Fi) [3], while also as the promising components of the forthcoming standards. Alternative to the receive diversity techniques that have been proposed recently, the transmit diversity schemes have also provided flexibility and freedom in the design of simpler mobile terminals of downlink transmissions in which the hardware complexity and power consumption of mobile ends are crucial. The most popular and efficient receive diversity techniques in the literature are maximal-ratio combining (MRC) [4], [5], equal-gain combining (EGC) [5], selection combining (SC) [5] and generalized selection combining (GSC) [6]. In addition to the receiver diversity techniques, the literature have consisted of many different transmit diversity schemes. One of the most popular transmit diversity technique is orthogonal space-time block coding (OSTBC) [6]-[9] that is shown to provide full-diversity orders by performing the communications via predefined transmission matrix without requiring the channel state information (CSI) partially or fully at the transmitter. This advantage has caused OSTBC to be a hot topic in the wireless communications literature and encouraged OSTBC schemes to be considered for wireless communications standards and protocols. Another full-diversity order transmit diversity scheme is transmit beamforming (TB) [10]. TB schemes perform signal transmission by using optimal transmit weights for each transmit branch that are constructed at the transmitter after a receiver estimation and feedback procedure. Hence, the optimality of this scheme is highly dependent on the perfectness of the channel estimation and feedback processes.

Another diversity technique that can be employed at the transmitter and/or receiver side is antenna selection which consists of receive antenna selection (RAS) [5], transmit antenna selection (TAS) [11], [12], and joint transmit and receive antenna selection (JTRAS) [13] schemes. For all three schemes, the main objective is to convey the information signal over the best available (transmit and/or receive) diversity branches which provide full-diversity orders with reduced number of radio-frequency (RF) chains (i.e., reduced number of power-amplifiers and signal processing units). Performing the transmission/reception over the transmitter/receiver antennas brings along with considerable enhancements in the overall system performance.

However, note that, the performance investigations on many diversity schemes that have been shown to be efficient at combating the fading effect are valid under the ideal conditions in which the channel estimation is perfect, the CSI (partially or fully) is

sent to the transmitter with no delay and feedback error. Unfortunately, in the presence of practical impairments such as channel estimation error (CEE), feedback delay and error (FD/FE), the aforementioned enhancements brought along with the usage of multi-antenna diversity schemes can not be obtained in the overall performance of WCSs.

Although the ideal channel estimation and feedback link assumptions are frequently used for and are efficient at first order examination of the WCSs, they lack the ability of interpreting the efficiency of diversity schemes and their feasibility for real-world wireless communications applications. Hence, the performance analyses of these diversity schemes in the presence of practical impairments such as CEE, FD, and FE are of great importance for examining the practical performance of the diversity schemes that would show their suitability and feasibility to the real-world wireless communications applications and standards.

1.1. The Aim of the Thesis

Through the investigation provided in this thesis, the performances of hybrid diversity schemes (HDSs) that are constructed by combining antenna selection at the transmitter and/or receiver with other combining and beamforming schemes are examined under the effects of practical impairments such as CEE, FD, and FE. By focusing on the derivation of performance metrics of HDSs in the presence of practical imperfections, the robustness and the fragility of different diversity schemes are exposed that results in a useful interpretation of the feasibility and suitability of the investigated WCSs to real-world communications applications. Thus, this thesis presents an extensive and a more realistic examination on the actual performance of many well-investigated and approved diversity schemes in erroneous channel estimation and feedback conditions. Moreover, this thesis, after analyzing the differing characteristics of several HDSs in different types of practical imperfections, proposes modified versions of some HDSs by showing more robustness and superiority than the conventional (not modified) ones in the presence of some type of practical impairments.

1.2. The Contributions of the Thesis

During the design, development, and implementation processes of wireless communications standards and WCSs, the main motto has always been to provide the best available level of communications reliability within the minimized levels of power consumption, bandwidth excess, and hardware and signal processing complexity. Therefore, the investigation of the practical-performance of multi-antenna WCSs that are shown in literature to perform well in fading channels are of great importance at the point of interpreting the feasibility of them for real-world wireless communications applications. Also, the proposal of modified HDSs that are shown to have higher immunity to the presence of practical impairments than the conventional ones constitute alternative schemes for real-world wireless communications applications and reveal great amount of information about the further WCS proposals.

Within the thesis' consideration, the exact and asymptotic expressions of performance metrics such as average outage probability (OP), average bit/symbol error rates (BERs/SERs) that provide useful insights for the performance interpretation of WCSs are derived and presented. Also, computer simulations based on the well-known Monte Carlo method are performed in order to validate the theoretical results. Moreover, the extensive work included in this thesis provides guidance and encouragement for the investigation of other practical impairments' effect on WCSs.

The detailed examinations related to the investigated HDSs and their outcomes that have presented throughout the thesis have also been published [33], [70], [74], [76], [86] or under consideration for publication [75], [85], [87] as journal papers [70], [74], [75], [85], [87] and conference proceedings [33], [76], [86].

1.3. The Contents of the Thesis

The performance examinations consisting of system models and derivations followed by the general definitions and descriptions related to the investigated diversity schemes and the practical impairments are given in detail in succeeding sections.

In the second chapter of the thesis, the background information needed to interpret the analyses' results are presented as the outcomes of the thesis. The characteristics of the wireless propagation channel, the fading effect and the

counter-measures that are proposed to combat with it, and the basic descriptions of the practical channel estimation and feedback impairments (i.e., CEE, FD and FE) are introduced briefly.

The third chapter of the thesis focuses on the performance investigation of HDSs in the presence of both CEE and FD. By considering the joint degrading effect of CEE and FD in Nakagami- m fading channels [18], the performance metrics related to hybrid maximal-ratio transmission (MRT)&RAS scheme [70] is provided in the first part of this chapter. Afterwards, by providing a realistic approach and revising the variance values that are used in the first part, the second part presents a unified framework that consists of four HDSs: TAS/OSTBC, TAS/MRT, MRT&RAS and JTRAS [75]. The exact and asymptotic performance metrics of the unified system model covering these four schemes are derived considering both CEE and FD effects.

In the fourth chapter of the thesis, the performance examination of HDSs in the presence of FE is presented. Under the effects of FEs in a feedback channel that has been modeled as binary-symmetric channel (BSC), the performance metrics related to hybrid TAS/OSTBC&MRC and JTRAS/OSTBC schemes are exposed in the first part of this chapter [33], [76], [85]. Sequentially, by using a more realistic feedback channel model that has fading characteristics as the main transmission link, the second and third parts make HDS proposals that are constructed by the modification of the transmission strategies of TAS/OSTBC&MRC schemes [74], [86], [87]. Both second and third parts of this chapter expose the robustness and superiority of the proposed schemes in the presence of FEs when compared to the conventional TAS/OSTBC&MRC schemes by also deriving their performance metrics related to the erroneous feedback conditions.

Finally, the last chapter of the thesis draws conclusions about the analytical and numerical results presented in the third and fourth chapters. The conclusion chapter not only provides final interpretation and insights about the outcomes of the extensive performance investigation presented throughout the thesis, but also points out the crucial need for further performance investigations by mentioning some open problems.

2. BACKGROUND ON WIRELESS COMMUNICATIONS

In order to provide basic definitions and background information about the WCSs that are examined within the scope of the thesis, this section introduces the dynamics, properties and challenges of the wireless propagation channel and the proposed diversity techniques to overcome these challenges. Besides, the practical imperfections that are on the focus of the performance investigation are briefly described.

2.1. Wireless Communication Channel and its Properties

In wireless communications environments, the transmitted radio signals face several complications such as reflection, scattering and diffraction while propagating in free space. Radio signals might be seriously affected by the atmospheric propagation characteristics, the presence of high buildings, bridges and hilly terrain existing in the wireless propagation environment. As also stated in the Introduction section, the most destructive contribution of the signaling process is the fading effect when compared to other complications including the additive receiver noise that is generally modeled as a Gaussian random process (i.e., additive white Gaussian noise (AWGN)).

2.1.1. Classification of Fading Effects

The fading effect might have effect on the performance of WCSs via three different forms; big-scale fading (i.e., path loss), medium-scale fading (i.e., shadowing) and small-scale fading (i.e., multi-path fading) according to the way of physical happening. This subsection briefly introduces these three different type of fading effects.

2.1.1.1. Big-Scale Fading (Path Loss)

Wireless communications can be characterized as the sequential combination of the transmission of electrical information packages after being transduced to electromagnetic waves, the propagation of the information signals via radio signals

to the receiver, and the receiver's detection and decoding processes [14].

Electromagnetic waves radiated from the transmit antennas propagate in free space by spreading the total energy on a wide surface (e.g., sphere for a isotropic radiation). Therefore, at a distant receiver point, the received signal power would be much lower than the transmit antenna neighborhood. This phenomenon which is defined as the attenuation in the received signal level because of the radial distance the transmitted signals cross is frequently called big-scale fading or namely path loss. The propagating radio waves should have reached the receiver antenna terminals with a power level above a certain threshold in order to be intercepted. To satisfy the received signal power levels above the receiver sensitivity (i.e., the threshold), the transmit signal power should be set properly. Besides, the degrading effect of big-scale fading on the performance of WCSs could be fixed by using signal repeaters that are placed at proper locations. Hence, big-scale fading does not play a critical role on the overall performance of WCSs.

2.1.1.2. Medium-Scale Fading (Shadowing)

Because of the presence of high-altitude obstacles especially near the main propagation link between the transmitter(s) and the receiver(s) of the WCSs, transmitted waveforms might face succeeding shadings and diffractions over knife-edges that cause fluctuations in the total signal power arrived at the receiving terminals. This change in the signal power level is defined as medium-scale fading or namely shadowing effect in the wireless communications literature [15]. The amount of medium-scale fading or shadowing highly depends on the electromagnetical, geometrical and mobility properties of the obstacles in the transmission/reception environment. Besides, the variation of all these parameters together with the number of obstacles results in a random shadowing characteristics which have frequently been modelled via Log-normal distribution [16].

2.1.1.3. Small-Scale Fading (Multi-Path Fading)

In WCSs, especially at mobile applications, the usage of low-altitude receiver antennas that are surrounded by so many obstacles of higher-altitudes generally hedges the existence of the line-of-sight (LoS) component between the transmitter and receiver

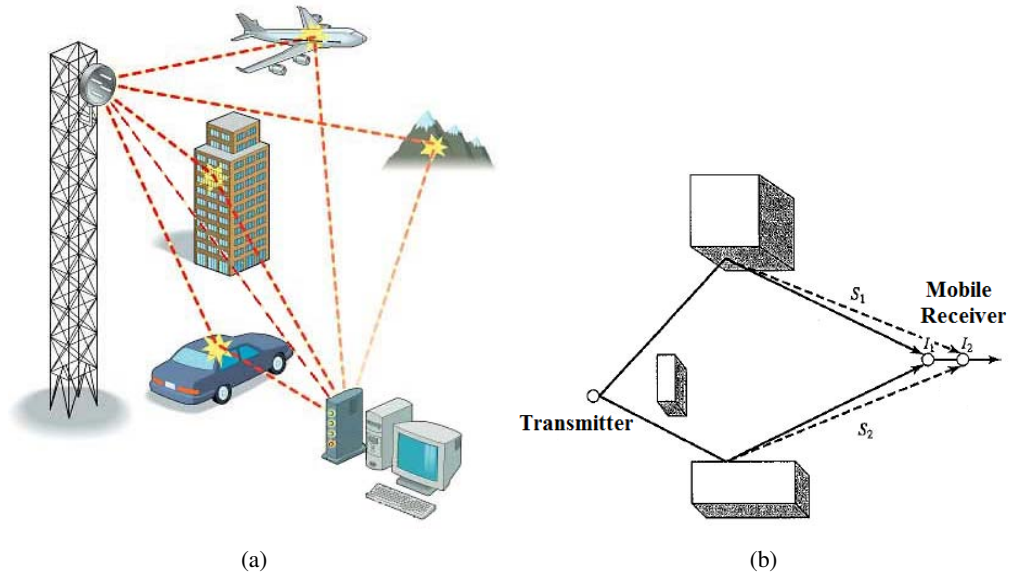


Figure 2.1: Schematics of the multi-path propagation phenomenon (a) in the presence of LoS component and (b) in the absence of LoS component.

ends. In such a situation, communications are performed through the reflected, scattered and diffracted signal replicas caused by the surrounding and blocking structures. Hence, these signal replicas would arrive at receiver antenna terminals with different amplitudes, delays and phases which result in small-scale fading or namely multi-path fading effect on the received signals. The schematic description of this phenomenon is given in Figure 2.1. As a consequence of the multi-path propagation, a bunch of signal replicas with different amplitudes, delays and phases whose superposition would probably bring along with deep fades in the received signal power level is obtained. The random nature of the superposed signal cluster would cause crucial challenges in channel estimation, equalization and tracking processes and consequently substantial degradations in overall performances of WCSs. Considering its potential of destructiveness, small-scale or multi-path fading constitutes the biggest challenge of the wireless propagation environments. Note that, since it is the most serious degradation item of the fading species, small-scale fading effect is simply called the fading effect throughout the following sections of the thesis as it has been done in many literature work. The time and frequency dependency of the fading process can be visualized with the help of the two-dimensional surface plot given in Figure 2.2 related to the time and frequency response of a typical wireless propagation channel. Beyond any doubt, the time and frequency instants corresponding to the deep fades shown in Figure 2.2 cause severe degradations in the communications

reliability. The time and frequency dependency of the fading characteristics has a great

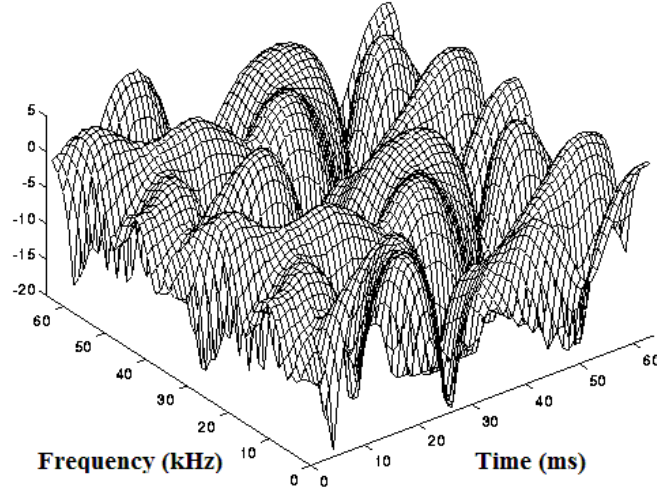


Figure 2.2: The time-frequency dependency of wireless communications channel.

importance for the classification of fading channels. When classified due to the time dependency, fading channels are categorized as slow fading or fast fading channels. Besides, when classified due to the frequency dependency, they are categorized as frequency-selective or frequency-non-selective (i.e., flat) fading channels [15].

2.1.2. Modelling of the Fading Channels

Since the fading effect encountered as the result of numerous reflections, diffractions, attenuations and scatterings that are of random nature and are not predictable, it cannot be expressed or analyzed via any deterministic approach. Alternatively, in the literature, there are several statistical models related to the fading processes. The most famous distributions that are often preferred to model the fading effect are Rayleigh, Ricean and Nakagami- m , in addition to rarely preferred fading models such as Weibull, Generalized Gamma and Generalized K [17].

2.1.2.1. Rayleigh Fading Model

In the case of no LoS existence between the transmitter and the receiver ends, the communications are performed via reflection and scattering-based signal replicas in the fading environment. Considering the in-phase and quadrature components of the total channel gains (i.e., the fading coefficients) follow Gaussian distribution with zero mean and the same variance, the envelope of the fading coefficients is distributed

according to Rayleigh distribution. The probability density function (PDF) related to Rayleigh fading process is given below [17]:

$$f_X(x) = \frac{2x}{\Omega} \exp\left(-\frac{x^2}{\Omega}\right), x \geq 0 \quad (2.1)$$

Here, $\Omega = E[x^2]$ denotes the mean power of the fading channel.

2.1.2.2. Ricean Fading Model

In addition to the scattering properties of Rayleigh fading case, the Ricean distributed fading coefficients might have non-zero mean values for both the inphase and the quadrature components which is brought along with the possible existence of LoS component or a more dominant signal component than the other signal branches. The in-phase and quadrature components of the Ricean fading coefficients follow Gaussian distribution with the same variance value and non-zero mean values. The Ricean fading model has a parameter that defines the ratio of the LoS component's power to the scattering components' total power, and the PDF expression of this model is given below [16]:

$$f_X(x; K) = \frac{2x(K+1)}{\Omega} e^{-K - \frac{(K+1)x^2}{\Omega}} I_0\left(\frac{K(K+1)}{\Omega}\right), x, K \geq 0 \quad (2.2)$$

In the PDF expression, $I_0(\cdot)$ denotes the zeroth order modified Bessel function of the first kind. Here, by the increase in the value of Ricean parameter K , the average power of the LoS (or another dominant signal) component increases which results in enhancements in the overall performance of WCSs. For asymptotic values of the parameter ($K \rightarrow \infty$), the fading channel characteristics converges to AWGN channel characteristics which meant the vanishing of the fading effect. Also, in the case of no LoS (or dominant signal) component existence, the fading channel's reduction to the Rayleigh fading channel can be easily observed by substituting $K = 0$ into the PDF expression in Equation (2.2).

2.1.2.3. Nakagami- m Fading Model

Nakagami- m , which provides a more generic and extensive fading model by modelling the fast fading effects at long distances, has been proposed in 1960 by Nakagami [18]. The PDF expression of a fading coefficient with Nakagami- m fading characteristics is given below:

$$f_X(x; m) = \frac{2m^m x^{2m-1}}{\Omega^m \Gamma(m)^m} \exp\left(-\frac{mx^2}{\Omega}\right), x \geq 0, m \geq \frac{1}{2} \quad (2.3)$$

At asymptotically higher values of the fading parameter ($m \rightarrow \infty$), the fading channel transforms into an AWGN channel as it was for the Ricean fading channel and $K \rightarrow \infty$. In addition, for some finite specific values of the fading parameter such as $m = 1$ and $m = 0.5$, the fading channel converges to Rayleigh fading channel and One-sided Gaussian fading channel, respectively.

2.2. Diversity Techniques

The reliability of WCSs are inversely proportional to the fading intensity of the transmission environment. Employing diversity techniques has been the most efficient way of combating the fading effect and preserving the reliability of communications at sufficient levels. Despite being applied in different ways and terminals, all diversity schemes are based on the same principle of providing the replicas of the message signal that have faced independent channel attenuations [14], [19]. The literature consists of many different techniques that are focused on providing diversity signals at the receiver end. Most famous diversity techniques have proposed to convey the message signal over independent time slots, frequency channels, antenna polarizations and multiple transmit and/or receive antennas. MIMO WCSs which are probably the most efficient and practical solutions against the fading channel deteriorations, have been proposed to gather multi-antenna (spatial) diversity by performing the signal transmission and/or reception with multiple antennas equipped at both sides. The most important advantage of MIMO WCSs have been their need for no extra bandwidth or transmit power when compared to a single-input single-output (SISO) system [19]. The spatial diversity provided by MIMO WCSs can be considered and gathered at the transmitter or the

receiver sides solely, or in corporation at both sides.

2.2.1. Receive Diversity Techniques

The multi-antenna schemes employed in the receiver side are mostly based on combining the independent signal replicas via some combining weights and/or selecting the best available one(s). Among these techniques that are called combining schemes, MRC, EGC, SC and GSC schemes exist [19].

2.2.1.1. Maximal-Ratio Combining

In this technique that has been firstly proposed by Kahn [4] in 1954, the signals from all of the receive branches are weighted according to their individual SNRs and summed. Figure 2.3 depicts the schematics of the MRC scheme where w_i , $i = 1, 2, \dots, n_R$ denotes the combining weight of the i^{th} receive branch. Here, co-phasing and weighting the signal replicas yield the instantaneous output SNR obtained at the receiver as

$$\gamma_{MRC} = \sum_{i=1}^{n_R} \gamma_i \quad (2.4)$$

where γ_i denotes the instantaneous SNR of each receive diversity branch. When the output SNR γ_{MRC} is examined, it is easily seen that MRC scheme produces an output SNR equal to the sum of the branch SNRs. Thus, this technique constitutes the optimal combining solution since it has the advantage of producing an output with an acceptable SNR even when none of the individual signals are themselves acceptable. Considering that each receive diversity branch has an equal average SNR value ($\forall i$, $E[\gamma_i] = \bar{\gamma}_i = \bar{\gamma}$), the average output SNR at the receiver would be obtained as [19, (7.1)]

$$\bar{\gamma}_{MRC} = n_R \bar{\gamma} \quad (2.5)$$

Note that, as clearly seen, the SNR at the combiner output would be n_R times the single receive branch SNR.

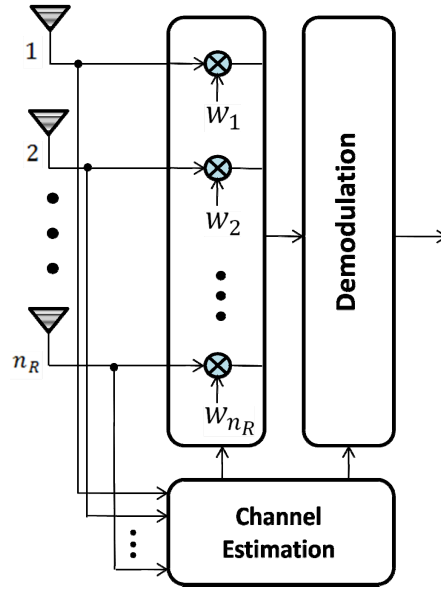


Figure 2.3: Block diagram of MRC scheme.

2.2.1.2. Equal-Gain Combining

When the cost that was brought along with the weighting process of MRC scheme has been thought to become a big challenge at the WCS design, the combining process has been considered with combining weights of unity. Thus, the diversity from all the received antennas are gathered by the coherent summation of all unit-weight signal replicas. This allows the receiver to exploit signals that are simultaneously received on each branch. The possibility of producing an acceptable signal from a number of unacceptable inputs is still retained, and performance is only marginally inferior to the optimal scheme (i.e., MRC). In order to show this clearly, the SNR expression at the EGC receiver can be examined [17, (9.89)]:

$$\gamma_{EGC} = \frac{(\sum_{i=1}^{n_R} \sqrt{\gamma_i})^2}{n_R} \quad (2.6)$$

Considering that each receive diversity branch has an equal average SNR value ($\forall i, E[\gamma_i] = \bar{\gamma}_i = \bar{\gamma}$), the average output SNR at the receiver would be obtained as [17, (9.49)]

$$\bar{\gamma}_{EGC} = \bar{\gamma} \left(1 + \frac{(n_R - 1)\pi}{4} \right) \quad (2.7)$$

2.2.1.3. Selection Combining

When compared to the aforementioned MRC and EGC schemes, this scheme is the simplest diversity technique that easily selects and activates only the best available receive antenna and uses its signal in demodulation process. SC determines the only receive branch by evaluating and comparing the SNR values of all branches. As the merit of this simplicity, SC scheme does not suffer from the hardware and signal processing burden of both MRC and EGC schemes. However, as expected, this flexibility obtained in the hardware design and the operating process brings along with a performance degradation when compared to MRC and EGC. The block diagram of SC scheme is given in Figure 2.4 below. Considering the instantaneous branch SNRs

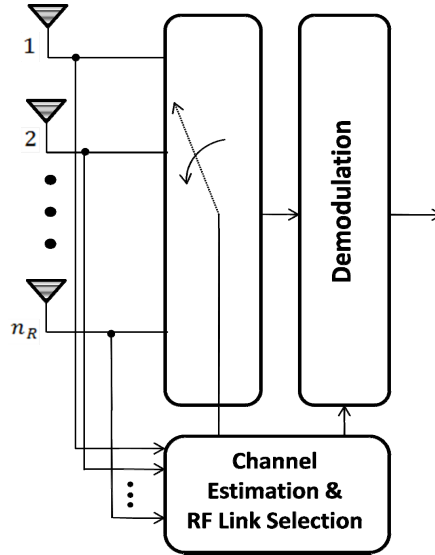


Figure 2.4: Block diagram of SC scheme.

$\gamma_i, i = 1, 2, \dots, n_R$, the output SNR at the SC receiver would be

$$\gamma_{SC} = \max_{1 \leq i \leq n_R} \{\gamma_i\} \quad (2.8)$$

Considering that each receive diversity branch has an equal average SNR value ($\forall i, E[\gamma_i] = \bar{\gamma}_i = \bar{\gamma}$), the average output SNR at the receiver would be obtained as [19, (7.10)]

$$\bar{\gamma}_{SC} = \bar{\gamma} \sum_{i=1}^{n_R} \frac{1}{i} \quad (2.9)$$

Also note that, SC scheme has been called the single-antenna-selecting version of receive antenna selection (RAS) technique in the related literature.

2.2.1.4. Generalized Selection Combining

In GSC, the signal reception is performed over an optimal receive antenna subset with n_S antennas that maximizes the output received SNR. GSC technique fills the performance and hardware complexity gap by performing the signal reception over a reduced antenna subset that results in flexibility and trade-off between system performance and hardware complexity. It is clearly seen that GSC scheme covers SC scheme by setting the number of active (selected) receive branch to 1, and similarly MRC scheme by activating all receive branches. The selection and combining structure of GSC scheme is given in Figure 2.5. The selection of the best available n_S receive

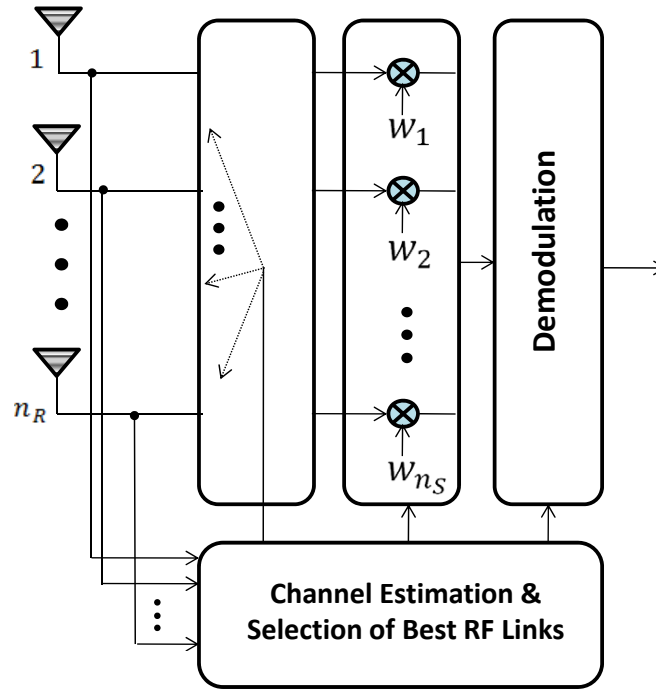


Figure 2.5: Block diagram of GSC scheme.

branches out of n_R ones results in the instantaneous SNR as [17, (9.330)]

$$\gamma_{GSC} = \sum_{i=1}^{n_S} \gamma^{(i)} \quad (2.10)$$

Here, instantaneous SNR values $\gamma^{(i)}$ denote the i^{th} ordered statistics that satisfy the inequalities $\gamma^{(1)} > \gamma^{(2)} > \dots > \gamma^{(n_R-1)} > \gamma^{(n_R)}$. Assuming that all receive branches have equal average values ($\forall i, E[\gamma_i] = \bar{\gamma}_i = \bar{\gamma}$), the combined average SNR of GSC scheme would be obtained as

$$\bar{\gamma}_{GSC} = n_S \bar{\gamma} \left(1 + \sum_{i=n_S+1}^{n_R} \frac{1}{i} \right) \quad (2.11)$$

2.2.2. Transmit Diversity Techniques

Receive antenna diversity techniques could be regardlessly implemented at wireless terminals that can manage and stand for the power consumption, hardware and signal complexity burdens of the multiple RF chains. For example, in mobile services, the quality of service and the communications reliability of the uplink can be easily enhanced by using multi-channel reception at the base station. However, again focusing on the case of mobile services, the deployment of these techniques at the mobile terminals for the downlink signaling would not be practical in terms of terminal dimensions and power consumption. Accordingly, in order to improve the downlink quality of service, transmit diversity techniques have been proposed. The transmit diversity can be gathered in the downlink transmission of base station-like terminals that have considerable flexibility to afford the power consumption, hardware and signal processing complexity caused by the multiple RF chains. Some of the transmit diversity schemes employ some kind of coding or signal processing method to provide an acceptable received SNR whereas some others operate with the continuous requirement of the partial or full CSI which meant a feedback link (and feedback burden) between the transmitter and the receiver.

The most important member of the schemes that have no need for CSI knowledge is the space-time coding (STC) scheme which consists of the orthogonal space-time trellis coding (OSTTC) and OSTBC schemes [7]-[9]. Since the implementation cost of them is lower when compared to OSTTC schemes, OSTBC schemes have become prominent in transmit diversity applications and researches. However, it is important to note that, despite their inefficiency at signal processing complexity, OSTTC schemes

are shown to provide higher coding gains especially in low SNR region when compared to the equivalent OSTBC case. The transmit diversity schemes that require the full and partial CSI knowledge are TB and TAS schemes, respectively.

2.2.2.1. Space-Time Block Coding

This technique has been based on the signal transmission via a predefined time-antenna schedule and the combining at the receiver via an MRC-like structure. The predefined time-antenna schedules which plan the distribution of the message signals on the transmit antennas and the time slots are called the transmission codeword matrices. Although OSTBC codeword matrices might be implemented as consisting of either real-defined signals or complex-defined signals, the complex codeword members are considered in this thesis since they provide full-diversity order with higher code rates. Here, the code rate of an OSTBC transmission matrix is defined as the ratio of the number of independent message signals in the codeword to the number of total time slots and it is denoted as R_s . The most popular OSTBC codewords are denoted as G_2 and G_3 that respectively employ two and three transmit antennas since the deployment of multi-antenna equipped schemes is the most probable. The first OSTBC transmission scheme was proposed in 1998 by Alamouti [7] who introduced the full-rate ($R_s = 1$) full-diversity OSTBC scheme with dual transmit antennas and dual transmission time slots. This scheme has also been called the Alamouti scheme. The transmission codeword matrix of Alamouti scheme is defined as given below.

$$G_2 = \begin{pmatrix} s_1 & s_2 \\ -s_2^* & s_1^* \end{pmatrix}$$

Here, each row of the matrix stands for a transmission period consisting of the versions of message signals which are assigned to two transmitter antennas. By virtue of the orthogonality of the codeword matrix, the symbol-wise detection and decoding processes are provided at the receiver. The dual-antenna equipped Alamouti scheme has been generalized in [8] for higher number of transmit antennas. Another popular member of OSTBC family has been the three-branch G_3 OSTBC scheme which employs three transmit antennas and eight time slots for the transmission of totally four independent message signals that results in a half-rate ($R_s = 0.5$) codeword. The

transmission codeword matrix of G_3 scheme has been defined in [8] as given below.

$$G_3 = \begin{pmatrix} s_1 & s_2 & s_3 \\ -s_2 & s_1 & -s_4 \\ -s_3 & s_4 & s_1 \\ -s_4 & -s_3 & s_2 \\ s_1^* & s_2^* & s_3^* \\ -s_2^* & s_1^* & -s_4^* \\ -s_3^* & s_4^* & s_1^* \\ -s_4^* & -s_3^* & s_2^* \end{pmatrix}$$

Note that the only OSTBC scheme that provides full-rate and full-diversity transmission is the Alamouti scheme among all available OSTBC schemes introduced in [8] and [9].

2.2.2.2. Antenna Beamforming

TB scheme performs signal transmissions with weighting the message signal's amplitude for all transmit antennas that means power allocation for each transmit branch. The transmit weights are adaptively computed due to the MIMO channel coefficients between the transmitter and the receiver antennas that are fed back after the channel estimation process at the receiver. The transmitter and receiver structure of the TB scheme is depicted in Figure 2.6. Similar to the combining schemes expressed in Section 2.2.1, TB scheme represents MRT scheme when the transmitter uses optimal transmit weights, the equal-gain transmission (EGT) scheme when it uses the transmit weights of equal absolute values, the selection-transmission (ST) scheme when it only weights and uses the best available transmit antenna and the generalized ST (GST) scheme when it weights and uses the best available transmit antenna subset.

2.2.2.3. Transmit Antenna Selection

In MIMO WCSs, it is possible to convey the information signals to the receiver end by using codewords as in OSTBC case and by using optimal weighting (i.e., optimal power allocation) as in TB case. However, both methods would suffer from the hardware and signal processing complexity caused by the joint deployment of multiple RF chains. The increase in the complexity parameters of the wireless designs hardens the implementation of these schemes in especially mobile terminals. However, TAS scheme could easily reduce the hardware complexity factors such as the number of

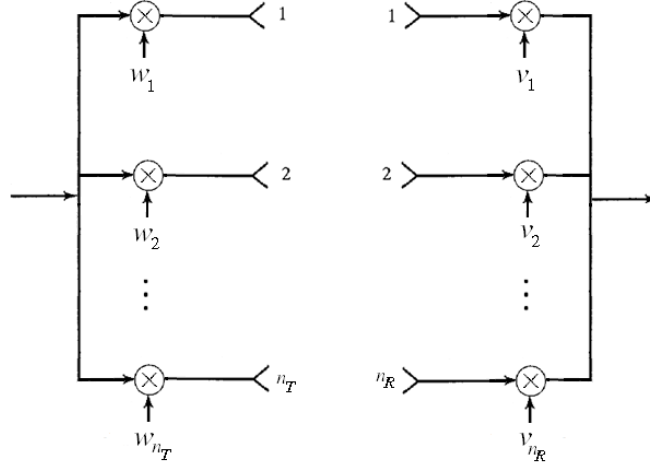


Figure 2.6: Block diagram of antenna beamforming scheme.

active RF chains and the corresponding power amplifiers by performing the signal transmission over a selected transmit antenna subset. The reduction provided in the hardware burden of WCSs increase the possibility of these schemes to be deployed in fixed or mobile terminals. Thus, TAS technique is of great importance at the point of deployment since it maintains the multi-antenna diversity gatherings of MIMO schemes with considerable reduction of the hardware and signal processing burden [9], [12]. An important advantage of TAS scheme is its flexibility at being solely deployed and being integrated to other transmit diversity schemes such as OSTBC and TB. A hybrid implementation scheme consisting TAS is depicted in Figure 2.7 below. In the system given in Figure 2.7, the transmit antenna subset consisting of n_S antennas

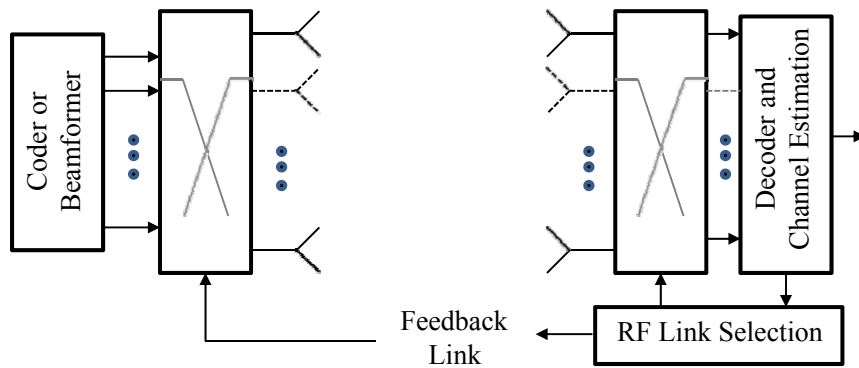


Figure 2.7: Block diagram of an antenna selection scheme.

that maximize the overall output SNR at the receiver is selected among n_T ones. The activation of the optimal transmit antenna subset is made after the channel estimation process performed at the receiver and the feedback of the transmit antenna indices that

are determined via TAS process. The asymptotic diversity order obtained by using TAS is shown to be the same as that was gathered by using all transmit antennas together.

2.3. Literature Survey for Multi-Antenna Hybrid Diversity Schemes

The performance enhancements that multi-antenna systems provided have focused the researchers' attention on these systems. For their spectral efficiency [1] and the fading-resistant structure [2], [3], they have been considered to be deployed in wireless communications systems as proposed within 3GPP LTE [1], WiMax [2], HSPA+ [1] and IEEE 802.11n (Wi-Fi) [3], while also being the promising components of the forthcoming standards. Several work have been presented in the literature examining full-diversity reception and transmission techniques such as MRC, OSTBC and TB which function at the expense of hardware complexity of multiple active RF chains. Despite providing substantial performance improvements, due to multiple RF chains they employ, multi-antenna systems suffer from hardware and signal processing complexities which are always bottlenecks especially for small mobile terminals. Subsequently, researches have been concentrated on antenna selection for reducing signal processing and hardware complexity which yields more feasible and efficient wireless systems while maintaining the benefits of multi-antenna diversity. The aim of antenna selection schemes is to maximize the received SNR (which brings along with full-diversity order transmission/reception) by performing the transmission/reception through a selected antenna subset. By employing antenna selection at either one side (TAS or RAS) or both sides (JTRAS) solely or in cooperation with other diversity techniques, HDSs are constructed to provide higher diversity gain and/or reduced hardware complexity.

The literature related to HDSs, which have been constructed by combining different diversity techniques at transmit and/or receive sides in order to obtain more diversity gain and/or reduce hardware complexity, includes many different wireless designs. Some of the previous works on HDSs have focused on TAS&MRC [11], [12], [20], [21], TAS/OSTBC [22]-[26], TAS/OSTTC [27], TAS&GSC [28], JTRAS [13], [29], [30], JTRAS/OSTBC [31]-[33], MRT&RAS [34]-[36] and TAS/MRT [37]-[41].

In [12], authors have provided exact BER results of binary phase shift keying

(BPSK) signals for TAS&MRC in Rayleigh fading channels. For independent and identically distributed (i.i.d.) Nakagami- m fading, moment-generating function (MGF)-based SER analysis of TAS&MRC scheme using M -ary modulations has been well investigated in [20]. A more extensive study on the TAS&MRC scheme for independent but non-identically distributed Nakagami- m fading case has been reported in [21]. Also, in [28], exact SER performances of combined TAS and GSC scheme have been examined for M -ary modulations.

Recent studies which concern both TAS and OSTBC have revealed the advantages of these schemes in the presence of severe channel and system impairments that might even end up with wrong antenna selection. The fact, that the diversity order of combined TAS/OSTBC schemes would be lower bounded by the diversity order achieved by OSTBC as discussed in [26], [27] have made these schemes prominent especially in practical conditions. A pairwise-error probability (PEP)-based analysis of TAS with OSTBC has been given in [22]. MGF-based BER/SER analyses of TAS/OSTBC systems have been given in [23] and [24]. In [23], the exact SER expressions have been derived for BPSK, coherent binary frequency shift keying (CBFSK), M -ary phase shift keying (M-PSK) and M -ary quadrature amplitude modulation (M-QAM) signals with Alamouti (G_2) coded transmission. [24] has analyzed the exact BER of M -ary pulse amplitude modulation (M-PAM) and M -QAM, and approximate BER of M-PSK signals for any number of selected transmit antennas. In [25] and [26], TAS/OSTBC scheme has been examined in Nakagami- m fading channels by using the MGF-based method. Also in both work, upper bounds have been derived which have shown that this scheme achieves full-diversity orders at high SNRs.

Authors of [13] have also derived upper bounded PEP expression for JTRAS case (i.e., joint TAS&RAS) which offers using only a single transmit-receive link with the highest branch SNR. The error performance of joint single TRAS scheme in i.i.d. Nakagami- m fading channels has been derived in [29] for binary and M -ary signals. In [29], it is shown that for a constant value of the product of numbers of transmit and receive antennas, JTRAS scheme provides the same diversity order and exactly the same BER/SER performance. Besides, the same problem is investigated in [30] in the presence of independent not-identically distributed fading channels.

TAS/OSTBC&RAS (i.e., JTRAS/OSTBC) scheme, which consists of combined

TAS/OSTBC at transmit end and single RAS at receive end, has been considered in [13], [31]-[33]. In [13], JTRAS with OSTBC has been analyzed asymptotically for Rayleigh fading channels by using upper bounds and PEP-based approach. The error performance of a special case, that is based on performing the OSTBC transmission by selecting two transmit antennas (i.e., Alamouti OSTBC [7]), has been examined in [31] for only BPSK modulation and Rayleigh fading channels. Although authors of [32] have considered the same problem with arbitrary number of selected antennas, they have given results for quadrature phase-shift keying (QPSK) modulations and Rayleigh fading channels by using numerical integrations and not provided any closed-form expressions for the performance metrics. The OP performance of JTRAS/OSTBC scheme has been examined for perfect feedback conditions in Nakagami- m fading channels in [33].

MRT&RAS scheme (i.e., combined MRT at transmit end and RAS at receive end) can be considered as well for wireless system designs that may tolerate some hardware complexity at transmit end. Besides, RAS provides simplicity at receive end since it uses only one RF chain with the highest SNR. In [34], authors have considered MRT&RAS scheme in Rayleigh fading channels and for only differential BPSK (DBPSK) signals by using PDF-based analysis approach. Author of [35] has examined MRT&RAS scheme in Rayleigh fading channels by using the PDF-based analysis method for ideal channel estimation and feedback (ICEF) conditions. The exact error performance of MRT&RAS scheme in Nakagami- m fading channels has been examined in [36] for only binary modulations and ICEF conditions.

The literature related to the combined TAS and TB techniques include the investigation of TAS/MRT, JTRAS/MRT and TAS/EGT schemes. In [37], the average SNR of the TAS/EGT scheme was derived for Rayleigh fading channels. Error performance analysis of the hybrid TAS/MRT scheme with the special case of two selected transmit antennas has been given in [38] for BPSK signals in Rayleigh fading channels by using the PDF-based analysis approach. The JTRAS/MRT scheme has been examined by doing computer simulations in [39] from the ergodic capacity perspective and in [40] from the error performance perspective. Also, the error performance of the hybrid TAS/MRT scheme has been examined in [41] for M -ary modulations by using the MGF-based analysis method over Nakagami- m fading channels.

2.4. Practical Impairments of Diversity Techniques

The performance enhancements provided by MIMO diversity schemes in fading channels are shown to gather full-diversity transmission/reception and considerable enhancements in fading channels as long as the channel estimation and the feedback processes are perfectly performed. Unfortunately, studies on the effects of practical impairments have shown the different degradation characteristics of many system and channel complications which mention the necessity of the investigation of diversity schemes under the system and channel imperfections. In order to examine the feasibility of these schemes in the real-world communications applications, and to expose the robustness of these schemes to erroneous conditions, researches have been being canalized on these practical issues. The literature includes some elementary work dealing with the performance analysis of simple (non-hybrid) diversity schemes for practical issues such as CEE [42], [43], FD [44], FE [45], frequency offset (FO) [46], [47], co-channel interference (CCI) [48], etc. The following subsections have presented the brief introduction of the practical imperfections CEE, FD and FE that are the focus of the thesis' examination.

2.4.1. Channel Estimation Error

The erroneous estimation of complex channel coefficients just before the demodulation process at the receiver causes impairments on the signal weighting, combining and selection efficiency gathered by the diversity schemes. Hence, the transmission rate and the communications reliability of WCSs might face severe degradations.

By denoting the channel coefficient matrix of a MIMO WCS with dimensions $n_R \times n_T$ as $\mathbf{H}$, the erroneous estimate of $\mathbf{H}$ (i.e., $\hat{\mathbf{H}}$) would be expressed as

$$\mathbf{H} = \hat{\mathbf{H}} + \mathbf{E}_{est} \quad (2.12)$$

where n_R and n_T respectively denote the number of receive and transmit antennas, $\mathbf{E}_{est}$ is the CEE matrix with entries that are frequently assumed to follow zero-mean Gaussian distribution. CEEs might occur as a result of the tracking deficiency of the

channel estimators employed at the receiver since the hardware and signal processing units used for channel estimation process are designed with some constraints on system complexity.

2.4.2. Feedback Delay

In closed-loop WCSs, the transmission of message signals is directed adaptively by the feedback information provided by the receiver to the transmitter through a feedback link. The partial or full CSI information that would cause the transmitter to construct the transmission weights (for TB scheme) or to determine the indices of the active transmit antenna(s) (for TAS schemes), is extracted from the (erroneous or perfect) estimates of complex channel coefficients. Unfortunately, in the case of channel coefficients' variation rate that is higher than the signaling period of transmitted waveforms (i.e., fast fading phenomenon), the CSI information might become outdated to be used for the following signaling period. The obligatory usage of old-fashioned CSI information in fast fading channels results in deteriorations in the system performance. This phenomenon which causes impairments in the communications reliability is called FD. The outdated channel coefficient matrix that has been obtained at the transmitter after a time-lag of τ can be represented as

$$\hat{\mathbf{H}} = \rho \hat{\mathbf{H}}^{(\tau)} + \mathbf{E}_{fd} \quad (2.13)$$

where $\hat{\mathbf{H}}$ denotes the receiver's estimate of the channel coefficient matrix. Here, ρ and $\mathbf{E}_{fd}$ respectively denote the correlation coefficient between two matrices $\hat{\mathbf{H}}$ and $\hat{\mathbf{H}}^{(\tau)}$, and the Gaussian distributed error matrix. In the related literature, ρ , which is a useful quantity to identify the time-correlation of channel coefficients, is generally defined in terms of the zeroth order modified Bessel function of the first kind: $\rho = J_0(2\pi f_D \tau)$. Here, f_D denotes the maximum Doppler frequency shift related to the time-varying characteristics of the wireless communications channel.

2.4.3. Feedback Error

In closed-loop WCSs, different from the CEE and FD complications, the distorting nature of the feedback link that might have fading channel characteristics and/or AWGN channel characteristics (at the transmitter of the main transmission link) is worth being analyzed. The quality of the feedback information would be degraded by the overall degradations caused by the fading and/or AWGN characteristics (i.e., FE). FEs whose existence are independent from the fading and/or AWGN effects of the main transmission link, cause severe degradations on the diversity schemes that are shown in the literature to be robust against the usual suspect (i.e., fading). Hence, the investigation of the system's robustness against FEs is of great importance when examining the feasibility of the proposed schemes for real-world wireless communications applications.

In almost all work in the literature, the feedback signal has been assumed to face only additive noise-type deteriorations rather than the fading effect which has been a *sine qua non* complication of the main transmission link [45], [49]. Under this assumption, the feedback channel is modelled as a binary symmetric channel (BSC) that is represented by the bit-wise transition probabilities given in Figure 2.8. Within this channel model, the transmission environment is considered to be

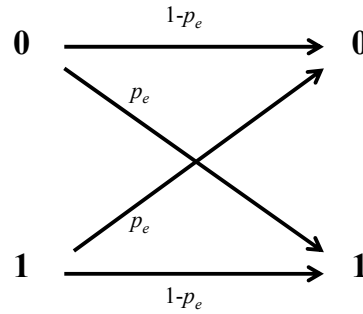


Figure 2.8: Transition probabilities of binary symmetric channel.

defined in bit-level and the bits representing the feedback message face a bit error probability of p_e . Although the BSC assumption on the feedback channel provides the opportunity of a simple and quick examination of the feedback impairments, it has lack of ability to examine the dependency of the feedback impairments' deteriorations on the modulation type of the feedback message or a physical phenomenon that

could exist on the feedback channel. Since then, there have been some work in the literature which claim that feedback channel modelling by including the fading and AWGN effects (similar to the main transmission link) would provide a more realistic perspective [19], [50].

3. PERFORMANCE ANALYSES IN THE PRESENCE OF CEE AND FD

This chapter focuses on the performance analyses of some HDSs, which are introduced in the previous chapter of the thesis, in the presence of CEEs and FD. Within the performance investigations, exact and approximate error probability expressions, exact OP expressions and Monte Carlo-based computer simulations are provided which are the most useful performance metrics. Besides, for some cases, other performance metrics such as ergodic capacity and SNR moments are included. Since all the analyses are performed for the generic Nakagami- m fading model, the numerical results have exposed the performance examination in various channel conditions. Also, the derivations are kept generic for the modulation level such that the performance of many frequently-used digital modulation schemes can be obtained by simple substitutions.

This chapter mainly consists of two sections that respectively focus on the practical-performance examinations of only the MRT&RAS scheme and a unified system model covering four HDSs: TAS/OSTBC, TAS/MRT, MRT&RAS and JTRAS. However, both sections seem to be dealing with the MRT&RAS scheme, the hybrid CEE and FD model used in two sections are different from each other. The model used in the second section provides a more realistic viewpoint by discarding some approximations on the variance value of the total CEE and FD effect that are made at the model given in the performance analysis in the first section of this chapter.

Before giving the analytical derivations of the performance metrics of the investigated HDSs, the current literature related to the performance investigations of diversity schemes under the effects of CEEs and/or FD is introduced.

3.1. Literature Survey About the CEE and FD

The literature related to wireless communications over fading channels, especially for the last fifteen years, has exposed the efficiency of multi-antenna systems in achieving enhanced communications reliability and/or channel capacity. Several diversity schemes, either employed solely or in cooperation with other diversity schemes (i.e., HDSs), have been shown to provide substantial improvements assuming

that the transmission channel is estimated perfectly and available at the transmitter partially or exactly. Unfortunately, real-world WCSs are likely to face various practical impairments such as CEE, feedback quantization error (FQE), FD and FE. Hence, some researchers have focused on practical channel estimation and feedback validity in order to examine the investigated diversity schemes from a more realistic viewpoint [41]-[44], [53]-[60].

The effect of the Gaussian-modeled CEE on the OP performance of the optimal receive diversity (i.e., MRC) has been investigated for Rayleigh fading channels in [42]. The performance of OSTBC schemes in the presence of mobility, multipath propagation and CEE has been examined for Rayleigh fading channels in [43]. The average BER of Alamouti-OSTBC [7] transmitter combined with antenna-selective receiver is derived in [53] through the effects of CEEs only for Rayleigh fading case and BPSK signals. By carrying out an upper bounded PEP analysis, [54] has provided the approximate BER performance of space-time-frequency coded transmission combined with RAS in Rayleigh fading channels in the presence of CEEs. The analysis provided in [44] has focused on the error performance of MRT employed with wide-band code division multiple access (W-CDMA) systems using BPSK and QPSK signals in the presence of FD in time-selective Rayleigh fading channels. The effect of FD on the outage and approximate error probability performances of transmit beamforming scheme has been presented in [55] and [56], respectively. By performing numerical computations for the error performance of MRT scheme using BPSK signals under erroneous channel estimation and outdated feedback cases, [57] has used the relationship between the variance of CEE and the number of total symbols and pilot symbols used in each frame as consistent with the WCS proposals provided in 3GPP [58]. The approximate OP analysis of TAS&MRC scheme under erroneous channel estimation and outdated feedback cases has been provided in [59]. The effect of FD on the performance of hybrid TAS/MRT scheme in Rayleigh fading channels has been analyzed in [60] for only BPSK signals. The same diversity scheme has been investigated by [41] for M -ary signals and Nakagami- m fading channels via simulating the CEE and FD effects.

The main motivation of the performance investigations provided in the following sections of this chapter is to focus on the practical performances of some HDSs under the effects of CEEs and FD that have not been considered in the literature

in an extensive and insightful examination. Therefore, this chapter aims to make contributions to the literature for the sake of exposing the real-world characteristics and the feasibility of HDSs to the wireless communications standards and protocols.

Firstly, in Section 3.2, the examination of the hybrid MRT&RAS scheme is given by providing exact and approximate expressions for some performance metrics such as average BERs/SERs of binary and M -ary signals, average OP, ergodic capacity and SNR moments that help evaluating other useful metrics such as average output SNR and amount of fading (AoF). Secondly, in Section 3.3, a unified framework is presented which investigates the effect of CEEs and FD on four HDSs; three of which (i.e., TAS/OSTBC, TAS/MRT and JTRAS) have not been analyzed in the literature for the presence of such practical imperfections. The unified system model presented in Section 3.3 also covers the hybrid MRT&RAS scheme that is investigated in Section 3.2 by using a more realistic CEE and FD model that helps the accurate and realistic observation of the deterioration caused by CEEs and FD.

3.2. Performance Analysis of MRT&RAS in Nakagami- m Fading Channels with CEEs and FD

This section focuses on the performance investigation of MRT&RAS scheme in i.i.d. flat Nakagami- m fading channels by considering CEEs and FD. Useful statistics of the post-processing SNR such as PDF, cumulative distribution function (CDF), MGF and n^{th} order SNR moments are presented. In order to examine the capacity and the error performances of MRT&RAS scheme, ergodic capacity, OP and average BER/SER for binary and M -ary modulations are analyzed. Exact analytical expressions for all the performance metrics are derived separately for practical situations consisting of CEE and FD, and ideal estimation and feedback conditions. By considering approximations for the statistics of the post-processing SNR, it is shown that, at high SNRs, the investigated system achieves full-diversity orders in the ideal case whereas it keeps maintaining receive diversity at practical impairments. The performance superiority of MRT&RAS scheme to other diversity schemes is also shown by numerical comparisons. Also, analytical performance results related to the OP and BER/SERs are validated by Monte Carlo simulations.

In the literature, there is no error performance analysis of even MRT scheme

specifically for binary and M -ary modulations in i.i.d. Nakagami- m fading channels. This section of this thesis, focusing on the hybrid MRT&RAS scheme in the presence of CEEs and FD in i.i.d. and flat Nakagami- m fading channels, makes the following specific contributions:

- We derive the PDF, CDF and MGF of the output SNR, also n^{th} order moments of the output SNR.
- The system capacity is examined from the ergodic capacity and OP perspectives.
- By using the conventional PDF-based analysis method, exact error probability expressions of binary and M -ary modulations are derived for the erroneous conditions and also by using the MGF-based analysis method, this section of the thesis also presents the exact average BER/SER expressions in terms of a unified integral expression for different modulation techniques such as DBPSK, CBFSK, M -PSK and M -QAM signals, for ICEF conditions.
- Besides, the asymptotic error performance analysis is carried out in order to derive the asymptotic diversity order (ADO) of the investigated HDS.
- Numerical comparisons that show the superiority of MRT&RAS scheme to other transmit diversity based techniques are also provided.
- Also, the analytical performance results in the presence of the CEE and FD are validated with the help of Monte Carlo simulations.

Note that, the analysis related to the ICEF conditions is valid for the fading parameters of half-integer values $m = N/2$, $N \in \mathbb{Z}^+$. Whereas, the analysis related to the erroneous conditions (CEE and FD) (except the ergodic capacity expressions which are given for integer values of the fading parameter $m \in \mathbb{Z}^+$) is valid for arbitrary values of $m \geq 1/2$.

The remainder of this section is organized as follows. We present the system and channel model in Section 3.2.1, briefly define the statistics of the output SNR in Section 3.2.2 and derive the MGF in Section 3.2.3, and n^{th} order SNR moments in Section 3.2.4. In Section 3.2.5, ergodic capacity and OP analyses have been done in order to examine the system capacity. Section 3.2.6 presents exact and approximate BER/SER performance results. Section 3.2.7 presents some numerical

results supported by simulations for system capacity and average error rates of CBFSK, M -PSK and M -QAM signals. Finally, Section 3.2.8 draws conclusions about the analytical and numerical results.

3.2.1. System Model and SNR Statistics for MRT&RAS Scheme

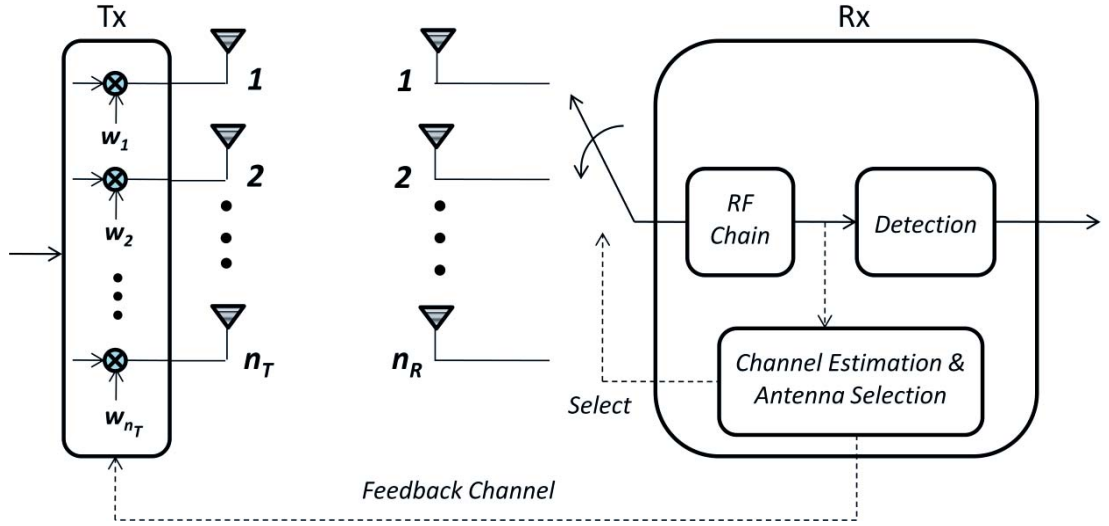


Figure 3.1: Block diagram of MRT&RAS scheme.

Hybrid MRT&RAS scheme, which is a MIMO scheme that employs n_T and n_R antennas at the transmit and receive ends, respectively as depicted in Figure 3.1, provides enhancements in the diversity order of a MRT system by using RAS or achieves the same diversity order as an MRT scheme with reduced number of RF chains. MRT&RAS scheme with higher diversity order further improves error performance of MRT and also outperforms existing transmit diversity based techniques such as OSTBC, TAS/OSTBC, JTRAS and TAS/MRT under perfect estimation and feedback assumptions. With this superiority, MRT&RAS scheme which is more complex at the transmitter and simpler at the receiver is preferable for systems requiring simple receivers, e.g., downlink.

In order to help the reader to follow the flow of the derivations and the analysis, the notations and assumptions that are valid throughout the examination have been given:

- T_s is the duration of each transmitted symbol,
- E_s is the average energy per symbol,

- N_d and N_{fr} are respectively the number of pilot symbols and total symbols used in each transmitted frame,
- N_o denotes the one-sided power spectrum of AWGN at each receiver,
- Bold-weight letters given in lowercase (e.g., ' $\mathbf{v}$ ') represent vectors,
- The mathematical notations, $(\cdot)^*$, $(\cdot)^H$ and $(\cdot)^T$ denote the conjugate, the Hermitian and the transpose of a matrix or a vector, and the statistical notation $E[\cdot]$ denotes the expectation operator of a random variable,
- $\beta \sim \mathcal{CN}(\mu, \sigma^2)$ denotes the random variable β that follows a complex Gaussian distribution with mean μ and variance $0.5\sigma^2$ per each dimension,
- $h_{i,j}^{(k)}$, $i = 1, 2, \dots, n_R$, $j = 1, 2, \dots, n_T$ denote the complex channel gains between each transmit-receive antenna pair for the k^{th} transmission interval. Their envelopes are assumed to follow i.i.d. Nakagami- m fading with the parameter m and squared mean $\Omega = E \left[|h_{i,j}^{(k)}|^2 \right]$ with frequency non-selective (flat) and quasi-static characteristics,
- $\mathbf{h}_p^{(k)}$ denotes the complex channel gain vector of the p^{th} (transmit or receive) diversity branch during the k^{th} transmission interval,
- $\hat{\mathbf{h}}_p^{(k)}$ denotes the estimate of the channel vector $\mathbf{h}_p^{(k)} = \hat{\mathbf{h}}_p^{(k)} + \mathbf{e}_{est}$ where $\mathbf{e}_{est}$ is the estimation error vector whose entries are zero-mean Gaussian distributed random variables with variance of σ_{est}^2 ,
- $\|\mathbf{v}\|_F = \sqrt{\sum_p |v_p|^2}$ denotes the Frobenius norm of the vector $\mathbf{v}$.

After briefly giving the assumptions and the notation that are valid throughout the analyses, the investigation focuses on the derivation of the basic statistical metrics required for the performance evaluation.

At every transmission period, all transmit antennas are used for MRT and among n_R receive antennas, a single antenna which maximizes the post-processing SNR is selected and used for reception. Before signal transmission, channel estimation and the selection of the receive antenna are performed by using pilot symbols. Pilot symbols are sent by the transmit antennas, and the received signals of all receive antennas are monitored in turn via single RF circuit and only the receive antenna

having the largest SNR is selected. The signal of the selected single receive antenna is fed to the single RF chain of the receiver for processing. But the time-varying characteristics of the fading channel cause impairments in the channel estimation. By considering CEE or FQE, which can be modeled with the same error model as shown in [61], the $1 \times n_T$ dimensional channel vector through the transmit antennas to the i^{th} receive antenna at the k^{th} signaling period (i.e., $\mathbf{h}_i^{(k)}$) is defined by means of its erroneous estimate (i.e., $\hat{\mathbf{h}}_i^{(k)}$) as $\mathbf{h}_i^{(k)} = \hat{\mathbf{h}}_i^{(k)} + \mathbf{e}_{est}$ where $\mathbf{e}_{est}$ represents the zero-mean complex Gaussian estimation error vector with the variance of $\sigma_{est}^2 = \frac{1}{N_d(E_s/N_0)} + 2\left\{\pi f_D T_s \left(\frac{2N_{fr}-N_d-1}{2}\right)\right\}^2$ [57]. Here, f_D denote the maximum Doppler frequency of the time-varying channel. Although the analysis has focused basically on CEE, the effect of FQE that has the same statistical properties as CEE (with zero mean and the variance of σ_{est}^2) on the system performance would be exactly the same. The receiver feeds the estimated channel gains back to the transmit end in order to maximize the instantaneous SNR at the receive end. But the delay, caused by the feedback link, forces the transmitter to use the delayed channel vector $\hat{\mathbf{h}}_i^{(k-1)}$ instead of the erroneously-estimated channel vector $\hat{\mathbf{h}}_i$. The relation between $\hat{\mathbf{h}}_i^{(k-1)}$ and $\hat{\mathbf{h}}_i^{(k)}$ is given in [57] as $\hat{\mathbf{h}}_i^{(k)} = \rho \hat{\mathbf{h}}_i^{(k-1)} + \mathbf{e}_{fd}$, where $\mathbf{e}_{fd}$ represents the zero-mean complex Gaussian error vector with the variance of $\sigma_{fd}^2 = (1 - \rho^2)E[|\hat{h}_{i,j}^{(k-1)}|^2]$. Here $\rho = J_0(2\pi f_D \tau)$, $0 < \rho < 1$, is the time correlation coefficient which is defined to qualify the feedback information and $J_0(\cdot)$ and τ denote the zero-order Bessel function of the first kind [62, (8.411-1)] and the time delay related to the feedback information, respectively. The transmitter uses the optimal transmit weight vector which is constructed as $\mathbf{w} = \frac{(\hat{\mathbf{h}}_i^{(k-1)})^H}{\|\hat{\mathbf{h}}_i^{(k-1)}\|_F}$ (satisfying $\|\mathbf{w}\|_F = 1$).

3.2.2. PDF and CDF of the Output SNR

After defining the channel estimation and the feedback process as shown above, at the i^{th} receive antenna, the received time-domain discrete signal at time instant $t = kT_s$ can be written as

$$y_i[k] = s[k]\sqrt{E_s}\mathbf{h}_i^{(k)}\mathbf{w}[k] + n_i[k] \quad (3.1)$$

where $s(t)$ is the transmit symbol with unit energy (i.e., $\int_0^{T_s} |s(t)|^2 dt = 1$) and $n_i(t)$ is wide-sense stationary complex AWGN with the variance of N_0 at the i^{th} receive antenna. Using (3.1) and the ratio of the square of the signal and the noise component, the instantaneous SNR related to the i^{th} receive branch can be defined as $\gamma_i = \frac{E_s}{N_0} \left| \left\{ \left(\rho \hat{\mathbf{h}}_i^{(k-1)} + \mathbf{e}_{fd} + \mathbf{e}_{est} \right) \left(\hat{\mathbf{h}}_i^{(k-1)} \right)^H / \|\hat{\mathbf{h}}_i^{(k-1)}\|_F \right\} \right|^2 = \frac{E_s}{N_0} |H_{e,i}|^2$ where $H_{e,i}$ (the equivalent channel gain of the i^{th} receive branch) can be shown to follow complex Gaussian distribution with the mean $\mu_{e,i} = \rho \|\hat{\mathbf{h}}_i^{(k-1)}\|_F$ and the variance of $\sigma_T^2 = \sigma_{est}^2 + \sigma_{fd}^2$, ($H_{e,i} \sim \mathcal{CN}(\mu_{e,i}, \sigma_T^2)$). For a given $\hat{\mathbf{h}}_i^{(k-1)}$ (assumed as a constant vector obtained from the feedback as in [57]), $|H_{e,i}|^2$ (the equivalent channel power gain of the i^{th} receive branch) follows a non-central Chi-square distribution with two degrees of freedom (since $H_{e,i}$ consists of two independent Gaussian components) and the non-centrality parameter $\nu = \mu_{e,i}^2$. The PDF of the non-central Chi-square distribution with two degrees of freedom and the non-centrality parameter ν can be easily obtained as

$$f_i(x; \nu) = \frac{e^{-(x+\nu)/\sigma_T^2}}{\sigma_T^2} I_0 \left(\frac{2\sqrt{\nu x}}{\sigma_T^2} \right), x \geq 0 \quad (3.2)$$

by using [16, (2.1-118)]. With the help of the relation ${}_0F_1(; 1; z) = I_0(2\sqrt{z})$ given in [63, (27.01.0003)] for the zero-order modified Bessel function of the first kind $I_0(\cdot)$ and the confluent hypergeometric limit function ${}_0F_1(; b; z) = \sum_{k=0}^{\infty} z^k / ((b)_k k!)$ [63, (02)], the PDF in (3.2) can be rewritten in terms of ${}_0F_1(; 1; z)$ as

$$f_i(x; \nu) = \frac{e^{-(x+\nu)/\sigma_T^2}}{\sigma_T^2} {}_0F_1 \left(; 1; \frac{\nu x}{\sigma_T^4} \right), x \geq 0 \quad (3.3)$$

Because the parameter ν depends on the channel realization $\hat{\mathbf{h}}_i^{(k-1)}$, the PDF related to the equivalent channel power gain of the i^{th} receive branch can be derived by averaging (3.3) over the distribution of the parameter ν . Using [36, (1)] (which can also be obtained by proper substitutions into [20, (2)] or [21, (2)]), one can easily show that the parameter ν follows Gamma distribution with the PDF given as

$$f_\nu(x; \Omega_\nu) = \left(\frac{m}{\Omega_\nu} \right)^{mn_T} \frac{x^{mn_T-1} e^{-xm/\Omega_\nu}}{\Gamma(mn_T)}, x \geq 0 \quad (3.4)$$

where $\Omega_\nu = \rho^2 \Omega$ and $\Gamma(s) = \int_0^\infty t^{s-1} e^{-t} dt$, $Re(s) > 0$, denotes the Gamma function [62, (8.310.1)]. Thus, averaging the conditional PDF in (3.3) over (3.4) will result in the PDF expression related to the equivalent channel power gain of the i^{th} receive branch $|H_{e,i}|^2$ as $f_i(x; \Omega_\nu) = \int_0^\infty f_i(x; \nu) f_\nu(\nu; \Omega_\nu) d\nu$. Performing the integral by using [62, (7.522-9)], $f_i(x; \Omega_\nu)$ can be obtained as

$$f_i(x; \Omega_\nu) = \frac{e^{-x/\sigma_T^2}}{\mathcal{A}^{mn_T} \sigma_T^2} {}_1F_1 \left(mn_T; 1; \frac{x}{\sigma_T^2} \mathcal{B}^{-1} \right), x \geq 0 \quad (3.5)$$

where $\mathcal{A} = 1 + \frac{\Omega_\nu}{m\sigma_T^2}$, $\mathcal{B} = 1 + \frac{m\sigma_T^2}{\Omega_\nu}$ and ${}_1F_1(a; b; z) = \sum_{k=0}^\infty \frac{(a)_k z^k}{(b)_k k!}$ [62, (210.1)] denotes the confluent hypergeometric function. Here, the term $(u)_n = \Gamma(u+n)/\Gamma(u)$ denotes the Pochhammer symbol [62, (9.749)]. Having derived the PDF expression, the CDF of $|H_{e,i}|^2$ is straightforward to obtain by substituting the series expansion of ${}_1F_1(a; b; z)$ into (3.5) and directly integrating. Rearranging the integral in terms of the incomplete Gamma function $\psi(s, x) = \int_0^x t^{s-1} e^{-t} dt$, $Re(s) > 0$, [62, (8.350.1)], results in the CDF of $|H_{e,i}|^2$:

$$F_i(x; \Omega_\nu) = \sum_{k=0}^\infty \frac{\mathcal{A}^{mn_T} (mn_T)_k}{\mathcal{B}^k \Gamma(k+1)^2} \frac{\psi \left(k+1, \frac{x}{\sigma_T^2} \right)}{\sigma_T^{-2k}}, x \geq 0 \quad (3.6)$$

Also, by using the relation given in [62, (8.351.2)] as $\psi(c_1, c_2 t) = \frac{{}_1F_1(1; 1+c_1; c_2 t)}{c_1 e^{c_2 t} (c_2 t)^{-c_1}}$, the CDF in (3.6) can be rewritten in terms of the confluent hypergeometric function as

$$F_i(x; \Omega_\nu) = \sum_{k=0}^\infty \frac{\sigma_T^{-2} (mn_T)_k}{{\mathcal{B}}^k \Gamma(k+1)} \frac{{}_1F_1 \left(1; k+2, \frac{x}{\sigma_T^2} \right)}{\Gamma(k+2) \mathcal{A}^{-mn_T}}, x \geq 0 \quad (3.7)$$

Since $\hat{\mathbf{h}}_i^{(k-1)}$ will be identical to $\mathbf{h}_i^{(k)}$ for ICEF conditions ($\sigma_T^2 \rightarrow 0$, $\rho \rightarrow 0$), the equivalent channel power gain of the i^{th} receive branch can be obtained as $|H_{e,i}|^2 = \|\mathbf{h}_i^{(k)}\|_F^2 = \sum_{j=1}^{n_T} |h_{i,j}^{(k)}|^2$. Thus, with the assumption that the channel estimation and feedback processes are error-free, the PDF of $|H_{e,i}|^2$ would be identical to (3.4) except a simple parameter change:

$$f_i(x; \Omega) = \left(\frac{m}{\Omega} \right)^{mn_T} \frac{x^{mn_T-1} e^{-xm/\Omega}}{\Gamma(mn_T)}, x \geq 0 \quad (3.8)$$

By simply integrating (3.8) and using the integral definition of incomplete Gamma function [62, (8.350.1)], the CDF related to $|H_{e,i}|^2$ can be obtained as [36, (2)] (which can also be obtained by proper substitutions into [20, (3)])

$$F_i(x; \Omega) = \frac{\psi(mn_T, xm/\Omega)}{\Gamma(mn_T)}, x \geq 0 \quad (3.9)$$

which can also be expressed as

$$F_i(x; \Omega) = 1 - e^{-xm/\Omega} \sum_{k=0}^{mn_T-1} \left(\frac{xm}{\Omega}\right)^k \frac{1}{k!}, x \geq 0 \quad (3.10)$$

for integer values of the product mn_T .

Note that, since the PDF and CDF expressions related to the erroneous and error-free conditions have been derived for the equivalent channel power gain $|H_{e,i}|^2$, the same statistics can be easily derived for the instantaneous SNR of the i^{th} receive branch, $\gamma_i = \bar{\gamma}_s |H_{e,i}|^2$, by simply interchanging Ω_ν with $\Omega_\gamma = \Omega_\nu \bar{\gamma}_s$ in (3.5), (3.6) (also in (3.7)) and Ω with $\Omega_\gamma = \Omega \bar{\gamma}_s$ in (3.8), (3.9) (also in (3.10)). Here, $\bar{\gamma}_s \triangleq E_s/N_0$ denotes the average SNR per symbol.

Before every transmission period, antenna selection process determines the receive antenna which will be used for reception by comparing instantaneous SNR values of all receive branches. Therefore, the instantaneous output SNR per symbol γ will be the maximum of γ_i : $\gamma = \max \{\gamma_i\}$. With the help of the highest order statistics [64], the CDF of the output SNR can be easily obtained by the product of the identical marginal CDF expressions $F_i(x; \Omega_\gamma)$ given in (3.6) related to the instantaneous branch SNRs γ_i :

$$F_\gamma(x; \Omega_\gamma) = [F_i(x; \Omega_\gamma)]^{n_R} = \mathcal{K}_0 \sum_{k_1=0}^{\infty} \sum_{k_2=0}^{\infty} \dots \sum_{k_{n_R}=0}^{\infty} \mathcal{K}_1 \prod_{i=1}^{n_R} \psi\left(k_i + 1, \frac{x}{\sigma_T^2}\right) \quad (3.11)$$

where $\mathcal{K}_0 = \frac{\mathcal{A}^{-mn_R n_T}}{\Gamma(mn_T)^{n_R}}$ and $\mathcal{K}_1 = \prod_{i=1}^{n_R} \mathcal{B}^{-k_i} \frac{\Gamma(k_i + mn_T)}{\Gamma(k_i + 1)^2}$. By using (3.7), the CDF of the output SNR can also be expressed in terms of the confluent hypergeometric function as

$$F_\gamma(x; \Omega_\gamma) = \mathcal{K}_0 \sum_{k_1=0}^{\infty} \sum_{k_2=0}^{\infty} \dots \sum_{k_{n_R}=0}^{\infty} \frac{\prod_{i=1}^{n_R} {}_1F_1\left(1; k_i + 2, \frac{x}{\sigma_T^2}\right)}{\mathcal{K}_2^{-1} e^{xn_R/\sigma_T^2} x^{-\mathcal{C}}} \quad (3.12)$$

where $\mathcal{C} = n_R + \sum_{i=1}^{n_R} k_i$, $\mathcal{K}_2 = \prod_{i=1}^{n_R} \frac{\Gamma(k_i + mn_T) \sigma_T^{-2}}{\mathcal{B}^{k_i} \Gamma(k_i + 1) \Gamma(k_i + 2)}$. By simply differentiating $F_\gamma(x; \Omega_\gamma) = [F_i(x; \Omega_\gamma)]^{n_R}$ with respect to x , using (3.5) and (3.7), the PDF of the output SNR is obtained as $f_\gamma(x; \Omega_\gamma) = n_R f_i(x; \Omega_\gamma) [F_i(x; \Omega_\gamma)]^{n_R-1}$ which yields

$$f_\gamma(x; \Omega_\gamma) = \mathcal{K}_3 \sum_{k_1=0}^{\infty} \sum_{k_2=0}^{\infty} \dots \sum_{k_{n_R-1}=0}^{\infty} \mathcal{K}_4 x^{\mathcal{C}'} e^{-xn_R/\sigma_T^2} \times {}_1F_1\left(mn_T; 1; \frac{x}{\sigma_T^2} \mathcal{B}^{-1}\right) \prod_{i=1}^{n_R} {}_1F_1\left(1; k_i + 2, \frac{x}{\sigma_T^2}\right) \quad (3.13)$$

where $\mathcal{C}' = n_R - 1 + \sum_{i=1}^{n_R-1} k_i$, $\mathcal{K}_3 = \frac{n_R \mathcal{A}^{-mn_R n_T}}{\sigma_T^{2n_R} \Gamma(mn_T)^{n_R-1}}$ and $\mathcal{K}_4 = \prod_{i=1}^{n_R-1} \frac{\Gamma(k_i + mn_T) \mathcal{B}^{-k_i}}{\Gamma(k_i + 1) \Gamma(k_i + 2)}$. Also, using the series expansion of ${}_1F_1(a; b; z)$ [62, (210.1)] in (3.5) and using the double-fold series expansion of the incomplete Gamma function given in [26, (5)] as $\psi(a_k, xc_k) = \Gamma(a_k) \sum_{p=0}^1 \sum_{q=0}^{p(a_k-1)} (-1)^p e^{-pxc_k} (xc_k)^q / q!$, in (3.6), the PDF of the output SNR can be expressed in an alternative way as:

$$f_\gamma(x; \Omega_\gamma) = \mathcal{K}_5 \sum_{k_1=0}^{\infty} \sum_{k_2=0}^{\infty} \dots \sum_{k_{n_R}=0}^{\infty} \sum_{p_1=0}^1 \sum_{p_2=0}^1 \dots \sum_{p_{n_R-1}=0}^1 \times \sum_{q_1=0}^{p_1 k_1} \sum_{q_2=0}^{p_2 k_2} \dots \sum_{q_{n_R-1}=0}^{p_{n_R-1} k_{n_R-1}} \mathcal{K}_6 e^{-xn_R \mathcal{D} / \sigma_T^2} x^{\mathcal{E}} \quad (3.14)$$

where $\mathcal{K}_5 = \frac{n_R \mathcal{A}^{-mn_R n_T}}{\sigma_T^{2n_R} \Gamma(mn_T)^{n_R}}$, $\mathcal{K}_6 = \mathcal{B}^{-\mathcal{C} + n_R} \sigma_T^{-2\mathcal{E}} (-1)^{\mathcal{D}-1} \frac{\Gamma(k_{n_R} + mn_T)}{\Gamma(k_{n_R} + 1)^2} \prod_{i=1}^{n_R-1} \frac{\Gamma(k_i + mn_T)}{\Gamma(k_i + 1) \Gamma(q_i + 1)}$, $\mathcal{D} = 1 + \sum_{i=1}^{n_R-1} p_i$ and $\mathcal{E} = k_{n_R} + \sum_{i=1}^{n_R-1} q_i$.

For ICEF conditions, the CDF expression related to the output SNR can be obtained by a similar approach using (3.10):

$$F_\gamma(x; \Omega_\gamma) = \left(1 - e^{-x \frac{m}{\Omega_\gamma}} \sum_{k=0}^{mn_T-1} \frac{(xm/\Omega_\gamma)^k}{k!}\right)^{n_R} = \sum_{p=0}^{n_R} \sum_{k=0}^{p(mn_T-1)} \binom{n_R}{p} \frac{\beta_k(p, mn_T) x^k}{(-1)^p e^{xmp/\Omega_\gamma}} \quad (3.15)$$

by using binomial expansion. Here, $\beta_k(p, mn_T)$ denotes the multinomial coefficients given in [62, (0.314)] and [36, (8)]. By simply differentiating $F_\gamma(x; \Omega_\gamma) = [F_i(x; \Omega_\gamma)]^{n_R}$ with respect to x , the PDF of the output SNR is obtained as $f_\gamma(x; \Omega_\gamma) = n_R f_i(x; \Omega_\gamma) [F_i(x; \Omega_\gamma)]^{n_R-1}$. After proper manipulations as in (3.15), the PDF results in

$$f_\gamma(x; \Omega_\gamma) = n_R \frac{(m/\Omega_\gamma)^{mn_T}}{\Gamma(mn_T)} \sum_{p=0}^{n_R-1} \sum_{k=0}^{p(mn_T-1)} \binom{n_R-1}{p} \times (-1)^p \beta_k(p, mn_T) x^{k+mn_T-1} e^{-xm(p+1)/\Omega_\gamma} \quad (3.16)$$

3.2.3. MGF of the Output SNR

Based on the statistics of the output SNR which are examined up until this point, the MGF expression, which is the most effective tool for examining the error performance of a communication system, is obtained for the MRT&RAS scheme in this subsection. One can derive the MGF of the output SNR by using the conventional definition $\mathcal{M}_\gamma(s) = \int_0^\infty e^{-sx} f_\gamma(x; \Omega_\gamma) dx$, whereas it is also possible to obtain the MGF expression by using the CDF of the output SNR as

$$\mathcal{M}_\gamma(s) = s \int_0^\infty e^{-sx} F_\gamma(x; \Omega_\gamma) dx \quad (3.17)$$

Using the CDF expressions given in (3.11) and (3.15) for erroneous and ICEF conditions, respectively, the MGF of the output SNR can be obtained for both cases. By substituting (3.11) into (3.17) and rearranging the resulting integral in terms of the Lauricella function $F_A^{(n)}$ described in [65, (2.4.2)] as $F_A^{(n)}(a; b_1, \dots, b_n; c_1, \dots, c_n; x_1, \dots, x_n) = \frac{1}{\Gamma(a)} \int_0^\infty e^{-st} t^{a-1} \{\prod_{i=1}^n {}_1F_1(b_i; c_i; x_i t)\} dt$, the MGF of the output SNR can be expressed as

$$\mathcal{M}_\gamma(s) = s\mathcal{K}_0 \sum_K \frac{\mathcal{K}_2 \Gamma(1+\mathcal{C})}{[a(s)/\sigma_T^2]^{1+\mathcal{C}}} F_A^{(n_R)} \left(1+\mathcal{C}; 1, \dots, 1; \right. \\ \left. k_1+2, \dots, k_{n_R}+2; \frac{1}{a(s)}, \dots, \frac{1}{a(s)} \right) \quad (3.18)$$

where $a(s) = s\sigma_T^2 + n_R$ and K denotes the n_R -fold summations. Similarly, substituting (3.15) into (3.17) and expressing the integral in terms of Gamma function results in the MGF expression for the ICEF conditions:

$$\mathcal{M}_\gamma(s) = \sum_{p=0}^{n_R-1} \sum_{k=0}^{p(mn_T-1)} \frac{\binom{n_R}{p} s (-1)^p \beta_k(p, mn_T)}{\Gamma(k+1) (s + \frac{mp}{\Omega_\gamma})^{k+1}} \quad (3.19)$$

3.2.4. SNR Moments

The moments of output SNR which are useful performance measures especially for evaluating the Average SNR and AoF metrics of WCSs are derived in this subsection. By substituting the PDF in (3.13) into the well-known definition $\mu_n = E[\gamma^n] = \int_0^\infty x^n f_\gamma(x; \Omega_\gamma) dx$ [17, (3.1.2)] and expressing the resulting integral in terms of $F_A^{(n)}$, the n^{th} order moment of the output SNR in the presence of CEE and FD can be derived as

$$\mu_n = \mathcal{K}_3 \sum_{K'} \mathcal{K}_4 \frac{\Gamma(2+n+\mathcal{C}')}{[a(0)]^{2+n+\mathcal{C}'}} F_A^{(n_R)} \left(2+n+\mathcal{C}'; 1, \dots, 1; \right. \\ \left. k_1+2, \dots, k_{n_R-1}+2; \frac{1}{a(\frac{mn_R}{\Omega_\gamma})}, \frac{1}{a(0)}, \dots, \frac{1}{a(0)} \right) \quad (3.20)$$

where K' denotes the $(n_R - 1)$ -fold summations. Similarly, considering (3.16) and expressing the integral in terms of Gamma function results in the n^{th} order moment of the output SNR for the ICEF conditions as

$$\mu_n = n_R \frac{(m/\Omega_\gamma)^{mn_T}}{\Gamma(mn_T)} \sum_{p=0}^{n_R-1} \sum_{k=0}^{p(mn_T-1)} \binom{n_R-1}{p} \\ \times (-1)^p \beta_k(p, mn_T) \frac{\Gamma(n+k+mn_T)}{(m(p+1)/\Omega_\gamma)^{n+k+mn_T}} \quad (3.21)$$

Proper substitutions in (3.20) and (23.1) enable deriving the average SNR (i.e., $n = 1$) and the amount of fading by the ratio $(\mu_2 - \mu_1^2)/\mu_1^2$ for both the erroneous and the ideal cases.

3.2.5. Ergodic Capacity and Outage Probability Analyses

Ergodic capacity can be defined as $C_{ERG} = E[C_\gamma]$, where $C_\gamma = \log_2(1 + \gamma)$ denotes the instantaneous capacity. Thus, averaging C_γ over the PDF of the output SNR gives us $C_{ERG} = \int_0^\infty \log_2(1 + x) f_\gamma(x; \Omega_\gamma) dx$. Substituting the PDF expressions given in (3.15) and (3.17) into this integral of the logarithmic argument, and using the identity given in [67, (78)], the ergodic capacity can be expressed as

$$C_{ERG} = \mathcal{K}_5 \sum_K \sum_P \sum_Q \frac{\mathcal{K}_6 \Gamma(1 + \mathcal{E})}{e^{-\mathcal{D}/\sigma_T^2}} \sum_{j=0}^{1+\mathcal{E}} \frac{\Gamma\left(j - 1 - \mathcal{E}, \frac{\mathcal{D}}{\sigma_T^2}\right)}{(\log_2(e))^{-1} \left\{ \frac{\mathcal{D}}{\sigma_T^2} \right\}^j} \quad (3.22)$$

for erroneous channel estimation and feedback conditions, where K , P and Q denote the multi-fold summations with indices k_i , p_i and q_i , respectively, described in (3.14), and

$$C_{ERG} = \frac{n_R \log_2(e)}{\left(\frac{\Omega_\gamma}{m}\right)^{-mn_T}} \sum_{p=0}^{n_R-1} \sum_{k=0}^{p(mn_T-1)} \sum_{j=1}^{k+mn_T} \frac{\binom{n_R-1}{p}}{(-1)^p} (mn_T)_k \times \frac{\beta_k(p, mn_T) \Gamma\left(j - k - mn_T, \frac{m(p+1)}{\Omega_\gamma}\right)}{e^{-m(p+1)/\Omega_\gamma} [m(p+1)/\Omega_\gamma]^{-j}} \quad (3.23)$$

for ICEF conditions. Since ergodic capacity gives ensemble average over all possible channel realizations, the OP which is defined as the probability that the instantaneous capacity is less than a given capacity (bandwidth efficiency) $\mathcal{R}$ (bit/s/Hertz) [51] is a more useful metric than ergodic capacity. Therefore, OP comprises a more realistic view on the system capacity. Since the instantaneous capacity is defined as $C_\gamma = \log_2(1 + \gamma)$, OP can be written as $P_{out} = Pr\{\log_2(1 + \gamma) \leq \mathcal{R}\} = Pr\{\gamma \leq 2^\mathcal{R} - 1\} = F_\gamma(2^\mathcal{R} - 1; \Omega_\gamma)$. As a result, the OP of MRT&RAS scheme can be easily obtained for the erroneous and ideal estimation and feedback conditions by simply substituting $2^\mathcal{R} - 1$ into the CDF expressions in (3.11) and (3.15), respectively.

3.2.6. Exact Error Performance Analysis

Recent studies on error performance analysis of WCSs in fading channels have used both the conventional PDF-based and the MGF-based approaches. Since the PDF/CDF and the MGF expressions related to the output SNR are already derived in Section 3.2.1-4, the exact error performance expressions can be obtained by using these approaches. In this section, in order to provide analytical tractability, the error probability of MRT&RAS scheme has examined by using the conventional PDF-based method for CEE and FD existence and by using the MGF-based method for the ICEF conditions.

Although the MGF expressions for both the erroneous and ideal conditions are derived, the error probability analysis for the CEE and FD cases is carried out by using the definition given in [66, (3)] as

$$P_s = - \int_0^\infty P'(e|x) F_\gamma(x; \Omega_\gamma) dx \quad (3.24)$$

which is obtained by integrating the conventional PDF-based SER expression by parts. Here, $P(e|x)$ denotes the conditional error probability (CEP) of the investigated modulation scheme. For binary modulations, CEP can be expressed as [17, (8.100)] $P(e|x) = \frac{\Gamma(\lambda_1, \lambda_2 x)}{2\Gamma(\lambda_1)}$, where the parameters are defined as $(\lambda_1, \lambda_2) = (1, 0.5)$ for NCBFSK, $(\lambda_1, \lambda_2) = (1, 1)$ for DBPSK, $(\lambda_1, \lambda_2) = (0.5, 0.5)$ for CBFSK and $(\lambda_1, \lambda_2) = (0.5, 1)$ for BPSK. By substituting (12) and the derivative of $P(e|x)$ which can be obtained as $P'(e|x) = -\lambda_2^{\lambda_1} x^{\lambda_1-1} e^{-\lambda_2 x} / 2\Gamma(\lambda_1)$ into (3.24), and using the definition of the Lauricella function $F_A^{(n)}$ [65, (2.4.2)], the exact average error probability expressions are obtained as [70]

$$P_b = \frac{\lambda_2^{\lambda_1} \mathcal{K}_0}{2\Gamma(\lambda_1)} \sum_K \frac{\mathcal{K}_2 \Gamma(\lambda_1 + \mathcal{C})}{[a(\lambda_2)/\sigma_T^2]^{\lambda_1 + \mathcal{C}}} F_A^{(n_R)} \left(\lambda_1 + \mathcal{C}; 1, \dots, 1; \right. \\ \left. k_1 + 2, \dots, k_{n_R} + 2; \frac{1}{a(\lambda_2)}, \dots, \frac{1}{a(\lambda_2)} \right) \quad (3.25)$$

For M -ary modulations such as QPSK and rectangular M -QAM, the CEP is defined as [17, (8.10)] $P(e|x) = \lambda_2 \mathcal{Q}(\sqrt{\lambda_1 x}) - \lambda_3 \mathcal{Q}^2(\sqrt{\lambda_1 x})$, where $\mathcal{Q}(\cdot)$ denotes the Gaussian error function and $(\lambda_1, \lambda_2, \lambda_3) = (1, 2, 1)$ for QPSK and $(\lambda_1, \lambda_2, \lambda_3) =$

$(3/(M-1), 4(1 - \sqrt{1/M}), 4(1 - \sqrt{1/M})^2)$ for M -QAM. Considering the definition of $\mathcal{Q}(\cdot)$ given in [68, (7.1.21)] as $\mathcal{Q}(x) = 0.5[1 - \frac{2x}{\sqrt{2\pi}} {}_1F_1(1, 1.5, 0.5x^2)]$, the derivative of $P(e|x)$ is obtained as $P'(e|x) = \frac{(\lambda_3 - \lambda_2)\sqrt{\lambda_1}}{e^{\lambda_1 x/2} \sqrt{8\pi x}} - \frac{\lambda_1 \lambda_3}{e^{\lambda_1 x/2} 2\pi} {}_1F_1(1, 1.5, \frac{\lambda_1 x}{2})$. Therefore, substituting $P'(e|x)$ and (3.11) into (3.24), the exact average SER expressions of QPSK and M -QAM signals can be derived as [70]

$$\begin{aligned}
P_s = & \sqrt{\frac{\lambda_1}{8\pi}} \mathcal{K}_0 \sum_K \mathcal{K}_2 \frac{(\lambda_2 - \lambda_3) \Gamma(0.5 + \mathcal{C})}{[a(0.5\lambda_1)/\sigma_T^2]^{0.5+\mathcal{C}}} F_A^{(n_R)} \left(0.5 + \mathcal{C}; \right. \\
& \left. 1, \dots, 1; k_1 + 2, \dots, k_{n_R} + 2; \frac{1}{a(0.5\lambda_1)}, \dots, \frac{1}{a(0.5\lambda_1)} \right) \\
& + \frac{\lambda_1 \lambda_3}{2\pi} \mathcal{K}_0 \sum_K \mathcal{K}_2 \frac{\Gamma(1 + \mathcal{C})}{a(\lambda_1)^{1+\mathcal{C}}} F_A^{(1+n_R)} \left(1 + \mathcal{C}; 1, \dots, 1; \right. \\
& \left. 1.5, k_1 + 2, \dots, k_{n_R} + 2; \frac{\lambda_1 \sigma_T^2}{a(2\lambda_1)}, \frac{1}{a(\lambda_1)}, \dots, \frac{1}{a(\lambda_1)} \right)
\end{aligned} \tag{3.26}$$

Note that Lauricella function $F_A^{(n)}(a; b_1, \dots, b_n; c_1, \dots, c_n; x_1, \dots, x_n)$ used in (3.18), (3.20), (3.25) and (3.26) satisfies the convergence criterion $\sum_{i=1}^n |x_i| < 1$ for possible values of its arguments and are evaluated by using the proper integral definition in [65, (2.3.3)].

For ICEF conditions, the error probability expressions are derived by using the MGF of the output SNR. In [17], the BER/SER expressions of DBPSK, CBFSK, M -PSK and M -QAM signals have been given in terms of MGF. The average BER of DBPSK signals is directly given by means of MGF as $P_{b,DBPSK} = 0.5\mathcal{M}_\gamma(1)$ [17] which can be easily obtained using (3.18) and (3.19) whereas those of other modulations need integration over the MGF. In [29], SER expressions of M -PSK and M -QAM signals have been given in terms of a common function $\mathfrak{T}_\gamma(\epsilon; \lambda) = \frac{1}{\pi} \int_0^\epsilon \mathcal{M}\left(\frac{\lambda}{\sin^2 \Phi}\right) d\Phi$ as $P_{s,MPSK} = 2\mathfrak{T}_\gamma(\pi/2; \lambda) - \mathfrak{T}_\gamma(\pi/M; \lambda)$ and $P_{s,MQAM} = 4q\mathfrak{T}_\gamma(\pi/2; \lambda) - 4q^2\mathfrak{T}_\gamma(\pi/4; \lambda)$, where $\lambda = \sin^2(\pi/M)$ for M -PSK and $\lambda = 1.5/(M-1)$, $q = 1 - \sqrt{(1/M)}$ for M -QAM. Note that setting $M = 2$ in $P_{s,MPSK}$ results in binary modulation case, with the parameter $\lambda = 1$ for BPSK and $\lambda = 0.5$ for CBFSK. Substituting the MGF in (3.19) into the common function $\mathfrak{T}_\gamma(\epsilon; \lambda)$ results in

$$\mathfrak{T}_\gamma(\epsilon; \lambda) = \sum_{p=0}^{n_R} \sum_{k=0}^{p(mn_T-1)} \int_0^\epsilon \frac{\mathcal{K}_7(\lambda/\sin^2 \Phi) d\Phi}{(\lambda/\sin^2 \Phi + mp/\Omega_\gamma)^{k+1}} \tag{3.27}$$

where $\mathcal{K}_7 = \binom{n_R}{p} \frac{(-1)^p}{\pi} \beta_k(p, mn_T) \Gamma(k+1)$. By making the change of variable $t = \sin^2 \Phi / \sin^2 \epsilon$ in the integral in (3.31) and using the definition of Appell hypergeometric function F_1 given in [62, (3.211)], $\mathfrak{T}_\gamma(\epsilon; \lambda)$ can be easily obtained as

$$\begin{aligned} \mathfrak{T}_\gamma(\epsilon; \lambda) &= \sum_{p=0}^{n_R} \sum_{k=0}^{p(mn_T-1)} \mathcal{K}_7 \frac{\Gamma(k+0.5) \lambda^{-k} \beta_k(p, mn_T)}{2\Gamma(k+1.5) (\sin \epsilon)^{-2k-1}} \\ &\times F_1(k+0.5, 0.5, k+1; k+1.5; \sin^2 \epsilon; -\alpha \sin^2 \epsilon) \end{aligned} \quad (3.28)$$

where the parameter is $\alpha = (mp)/(\lambda \Omega_\gamma)$. Note that F_1 can be easily calculated by using well known software programs such as MATHEMATICA and MAPLE. Thus, the exact average BER/SER of BFSK, M -PSK and M -QAM signals can be easily evaluated by using (3.28).

3.2.7. Asymptotic Error Performance Analysis

Since exact error probability expressions given in (3.25), (3.26) and (3.28) do not give much information about the diversity order, this subsection focuses on the asymptotic behavior of the error performance. For this purpose, approximate expressions for error probabilities of different modulation signals are derived by using the high-SNR approximation technique given in [69] which will enable us to observe the diversity order of the MRT&RAS scheme clearly. The relation between the asymptotic error performance of a system and the behavior of PDF of the output SNR has been well investigated in [69]. If it is possible to express the PDF of the total channel power gain $Z = \gamma/\bar{\gamma}_s$ (which was defined as $|H_{e,i}|^2$ in Section 3.2.2), as $f_Z(x; \Omega) = ax^t + o(x^t), a > 0$, one can express the error performance approximately as

$$P_e = \frac{2^t a \Gamma(t+1.5)}{\sqrt{\pi}(t+1)} (k \bar{\gamma}_s)^{-(t+1)} + o(\bar{\gamma}_s^{-(t+1)}) \quad (3.29)$$

for any modulation technique with a CEP $P(e|z)$ consisting of $\mathcal{Q}(\sqrt{zk\bar{\gamma}_s})$ where k denotes the modulation-dependent parameter. Here, the term $o(\cdot)$ denotes the omitted part of the asymptotic error probability expression for high SNR values by satisfying $f_1(x) = o(f_2(x))$ for $x \rightarrow x_0$, if $\lim_{x \rightarrow x_0} \frac{f_1(x)}{f_2(x)} = 0$, where f_1 and f_2

are polynomial-type functions. Substituting the approximate representation of the incomplete Gamma function given in [20, (15)] as

$$\psi(a_k, x_{C_k}) = \frac{(x_{C_k})^{a_k}}{a_k} + o(x^{a_k}), x \geq 0 \quad (3.30)$$

into the CDF expression in (3.11), selecting all the summation indices $k_i = 0, \forall i$ (in order to represent the summation in terms of the least power of the variable x as in [69]) and differentiating the resulting approximate CDF expression, the PDF of the output SNR can be expressed asymptotically as $f_\gamma(x; \Omega_\gamma) = \frac{n_R x^{n_R-1}}{\mathcal{A}^{mn_R n_T} \sigma_T^{2n_R}} + o(x^{n_R-1})$. By making some manipulations in this approximate PDF, the PDF of the total channel power gain Z can be easily obtained as $f_Z(x; \Omega_\nu) = \frac{n_R x^{n_R-1}}{\mathcal{A}^{mn_R n_T} \sigma_T^{2n_R}} + o(x^{n_R-1})$. As a result, for MRT&RAS scheme the parameters are found as $a = \frac{n_R}{\mathcal{A}^{mn_R n_T} \sigma_T^{2n_R}}$ and $t = n_R - 1$. Therefore, (3.29) can be rewritten as

$$P_e = \frac{n_R 2^{n_R-1} \Gamma(n_R + 0.5) (k \bar{\gamma}_s)^{-n_R}}{\mathcal{A}^{mn_R n_T} \sigma_T^{2n_R} \sqrt{\pi} \Gamma(n_R)} + o(\bar{\gamma}_s^{-n_R}) \quad (3.31)$$

for any modulation technique whose CEP can be described as the linear combination of $P(e|z) = \mathcal{Q}(\sqrt{z k \bar{\gamma}_s})$. Similarly, derivation of the approximate PDF of the total channel power gain by substituting the approximate representation of the incomplete Gamma function given in (3.30) into the CDF in the left hand side of (3.15); $F_Z(x; \Omega) = \left(\frac{\psi(mn_T, x m / \Omega)}{\Gamma(mn_T)} \right)^{n_R}$, results in the parameters as $a = (m^{mn_R n_T} mn_R n_T) / \Gamma(mn_T + 1)^{n_R}$ and $t = mn_R n_T - 1$ for the ICEF conditions. Thus, for a given CEP of $P(e|z) = \mathcal{Q}(\sqrt{z k \bar{\gamma}_s})$, the approximate error probability can be written as

$$P_e = \frac{\left(\frac{2m}{k \bar{\gamma}_s} \right)^{mn_R n_T} \Gamma(mn_R n_T + 0.5)}{2\sqrt{\pi} \Gamma(mn_T + 1)^{n_R}} + o(\bar{\gamma}_s^{-mn_R n_T}) \quad (3.32)$$

For example, considering the identity $k = 2$ for the binary case ($M = 2$), the asymptotic BER performances of BPSK and BFSK modulations can be obtained as [70] $P_{b,BPSK/BFSK} = \frac{n_R}{\sigma_T^{2n_R}} \left(1 + \frac{\Omega_\nu}{m \sigma_T^2} \right)^{-mn_R n_T} \frac{\Gamma(n_R + 0.5)}{2\sqrt{\pi} \Gamma(n_R)} (\lambda \bar{\gamma}_s)^{-n_R} + o(\bar{\gamma}_s^{-n_R})$ for the erroneous channel estimation and feedback conditions and $P_{b,BPSK/BFSK} =$

$\left(\frac{m}{\lambda\bar{\gamma}_s}\right)^{-mn_Rn_T} \frac{\Gamma(mn_Rn_T+0.5)}{2\sqrt{\pi}\Gamma(mn_T+1)^{n_R}} + o(\bar{\gamma}_s^{-mn_Rn_T})$ for the ICEF conditions. By using the CEP expressions approximately given as $P(e|z) \approx 2Q(\sqrt{2z\lambda\bar{\gamma}_s})$ and $P(e|z) \approx 4qQ(\sqrt{2z\lambda\bar{\gamma}_s})$ for M -PSK and M -QAM, respectively, the approximate error probability expressions for these modulations can be easily obtained for both erroneous and ideal channel estimation and feedback cases by simple substitutions into (3.31) and (3.32).

Since the CEP expression for DBPSK signals is not defined in terms of $Q(\cdot)$ function, the asymptotic error performance of this modulation cannot be expressed by using (3.31) and (3.32). By substituting the approximate expression of incomplete Gamma function given in (3.30) into the CDF expressions of both erroneous and ideal cases and deriving the approximate MGF expression in terms of the CDFs, the approximate error probability of DBPSK signals can be obtained via the simple identity $P_{b,DBPSK} = 0.5\mathcal{M}_\gamma(1)$.

Approximate error probability expressions which are derived for binary and M -ary modulations show that the MRT&RAS scheme achieves full diversity order of mn_Rn_T for ICEF conditions while it keeps maintaining the receive diversity (i.e., n_R) in the presence of CEE and FD.

3.2.8. Numerical Results for MRT&RAS Scheme in the Presence of CEEs and FD

In order to examine the performance of MRT&RAS scheme for the erroneous and ICEF conditions in Nakagami- m fading channels, some numerical results have been presented in this section consisting of theoretical and simulation results of miscellaneous system and channel conditions for unit channel power ($\Omega = 1$), total transmit energy of E_s (or E_b , which is average energy per bit) and feedback delay of one symbol ($\tau = T_s$) for the erroneous cases as in [44] and [57].

In Figures 3.2-6, some performance metrics such as OP and BER/SER are depicted versus average SNR. The theoretical and simulation results related to these figures have been given for some specific values of $T_s = 81.97 \mu s$ (which corresponds to the signaling rate of 12.2 kilobits per second for downlink communications) and the ratio $N_d/N_{fr} = 0.2$ which are compatible with the 3GPP specifications [58]. Also, note that, for the evaluation of the multi-fold infinite summations that appear in the

theoretical performance expressions related to the erroneous channel estimation and feedback conditions, considering only the first 50 terms per each summation has been good enough to provide consistency with the approximate and simulation results.

Since OP provides a more realistic view on the system capacity than ergodic capacity, theoretical and simulation results of the OP are presented in Figure 3.2 for two-branch MRT ($n_T = 2$) and several numbers of receive antennas in Rayleigh fading environment ($m = 1$). For constant values of $N_d = 4$ and $N_{fr} = 20$ and the bandwidth efficiency $\mathcal{R} = 1$ bit/s/Hz, the effect of the change in f_D on the OP performance (which is relevant to the system capacity performance) is depicted. Since the increment in the value of the Doppler frequency (i.e., f_D) corresponds to a proportional increment in the variances of both CEE and FD (σ_{est}^2 and σ_{fd}^2), it can be easily seen from the OP curves that the performance degradation due to increasing f_D becomes more severe. Note that theoretical results are in perfect agreement with simulation results. Figures 3.3-3.6 show the average error performances of

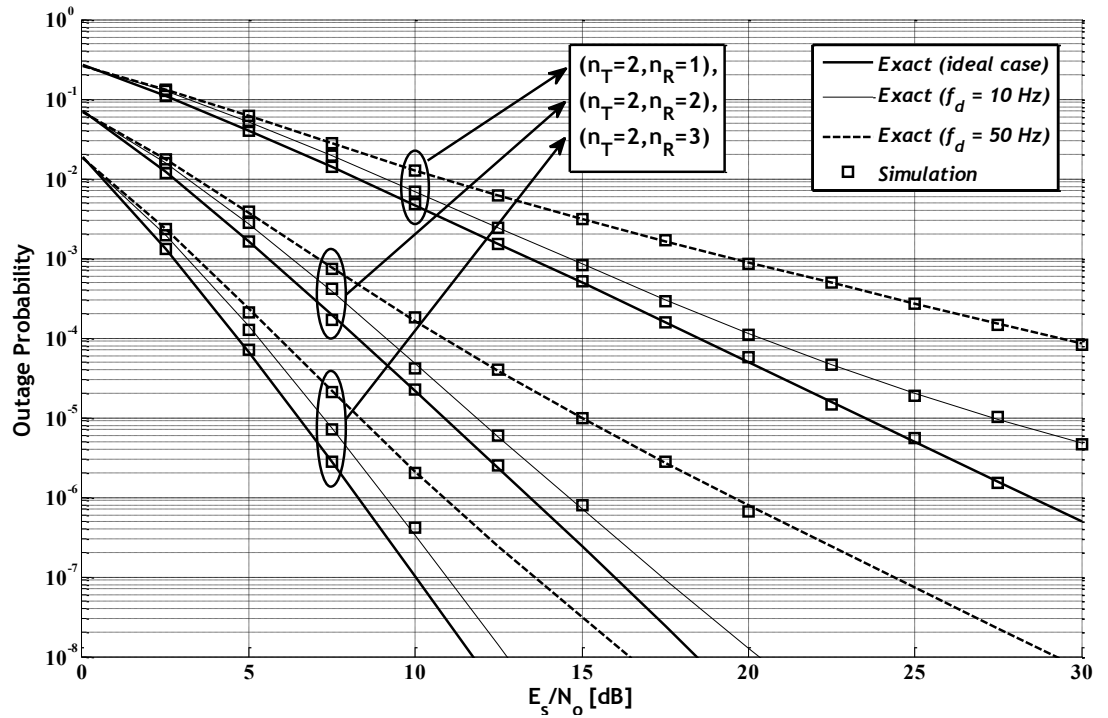


Figure 3.2: Average OP vs. average SNR per symbol for MRT&RAS scheme for several n_R and f_D ($n_T = 2$, $\mathcal{R} = 1$ bit/s/Hertz, $m = 1$ (Rayleigh Fading Channel), $N_d = 4$, $N_{fr} = 20$).

different modulation signals for different channel and system configurations. BER performances of MRT&RAS scheme with other diversity techniques in Rayleigh

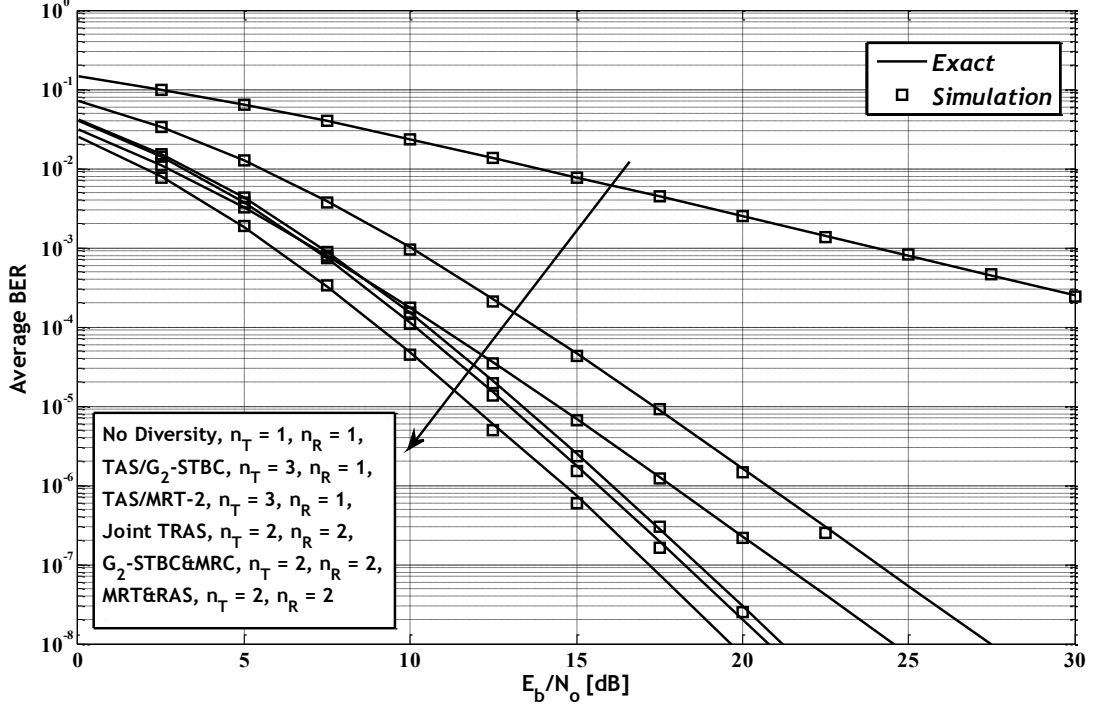


Figure 3.3: BER comparison of MRT&RAS scheme with other diversity techniques in Rayleigh fading channel ($m = 1$) for BPSK signals and $n_T + n_R = 4$ (with no CEE/FQE and FD).

fading channel ($m = 1$) are depicted in Figure 3.3 for BPSK and $n_T + n_R = 4$ in ICEF conditions (with no CEE/FQE and FD). Exact BER curves of G_2 -OSTBC&MRC (Alamouti) and TAS/ G_2 -OSTBC, JTRAS, TAS/MRT-2 are obtained by using [25, (11)], [29, (6)], [38, (17)], respectively. BER performance of TAS&MRC scheme is not depicted since it is the same as that of MRT&RAS scheme. As seen in Figure 3.3, MRT&RAS $\{n_T = 2, n_R = 2\}$ scheme (whose ADO is 4) outperforms the scenarios with the ADO of 4 (JTRAS $\{n_T = 2, n_R = 2\}$ and G_2 -OSTBC&MRC $\{n_T = 2, n_R = 2\}$) due to its optimum power allocation. Besides, when compared to the scenarios with the ADO of 3, MRT&RAS $\{n_T = 2, n_R = 2\}$ outperforms the scenario TAS/MRT-2 $\{n_T = 3, n_R = 1\}$ due to its higher diversity order and the scenario TAS/ G_2 -OSTBC $\{n_T = 3, n_R = 1\}$ due to its higher diversity order together with its optimum power allocation.

In Figure 3.4, the SER performances of QPSK signals are given for several n_T , n_R and f_D , $N_d = 4$, $N_{fr} = 20$, and for better transmission media with the fading parameter $m = 1.5$ in the presence of CEE and FD. The SER performances of 16-QAM signals in the presence of CEE and FD are given in Figure 3.5 for several

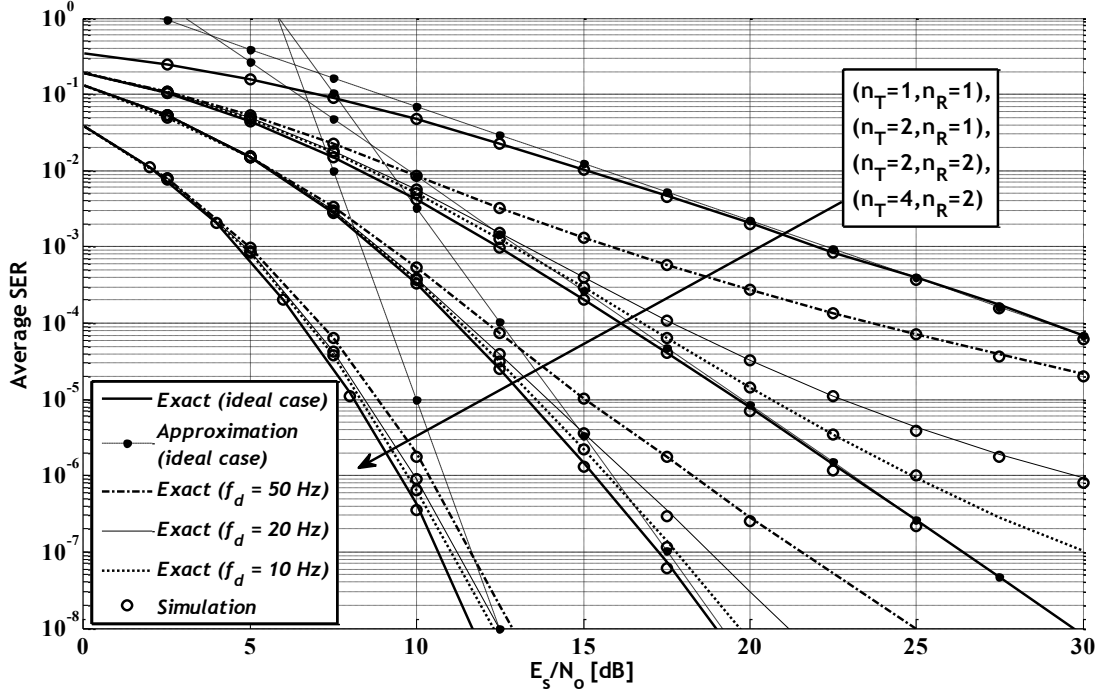


Figure 3.4: SER performance of MRT&RAS scheme in Nakagami- m fading channel ($m = 1.5$) for QPSK signals, several n_T , n_R and f_D in the presence of CEE and FD ($N_d = 4$, $N_{fr} = 20$).

n_T , n_R , N_d and N_{fr} , and for weak fading environments with the fading parameter $m = 2$ and $f_D = 20$ Hertz. In Figure 3.5, in order to provide a fair comparison, the ratio N_d/N_{fr} , which is significant for evaluating the effective signaling rate of WCSs, is kept equal to 0.2 [58] for all scenarios.

As seen in Figures 3.4 and 3.5, increasing the total number of antennas up to 6 leads to SNR gain values up to 17.5 dB and 14.5 dB, respectively at a SER of 10^{-3} for ICEF cases, simulation results perfectly match with the theoretical results, and approximate results for ICEF cases are compatible with exact ones at high SNRs. As the number of transmit antennas used for MRT and/or the number of receive antennas available for RAS increases, the diversity order and error performance of the MRT&RAS scheme improve. Moreover, the analytical and simulation results clearly show that for increasing variances of CEE and/or FD (σ_{est}^2 and/or σ_{fd}^2), the degradation on the SER performance becomes greater for MRT than that of MRT&RAS, which indicates that MRT&RAS scheme is more robust to the presence of CEE (or FQE) and FD than MRT scheme. The robustness can be seen from the SNR degradation (caused by CEE and FD) between the scenarios $\{n_T = 2, n_R = 1\}$ (which denotes the two-branch pure MRT scheme) and $\{n_T = 2, n_R = 2\}$ (which denotes the two-branch

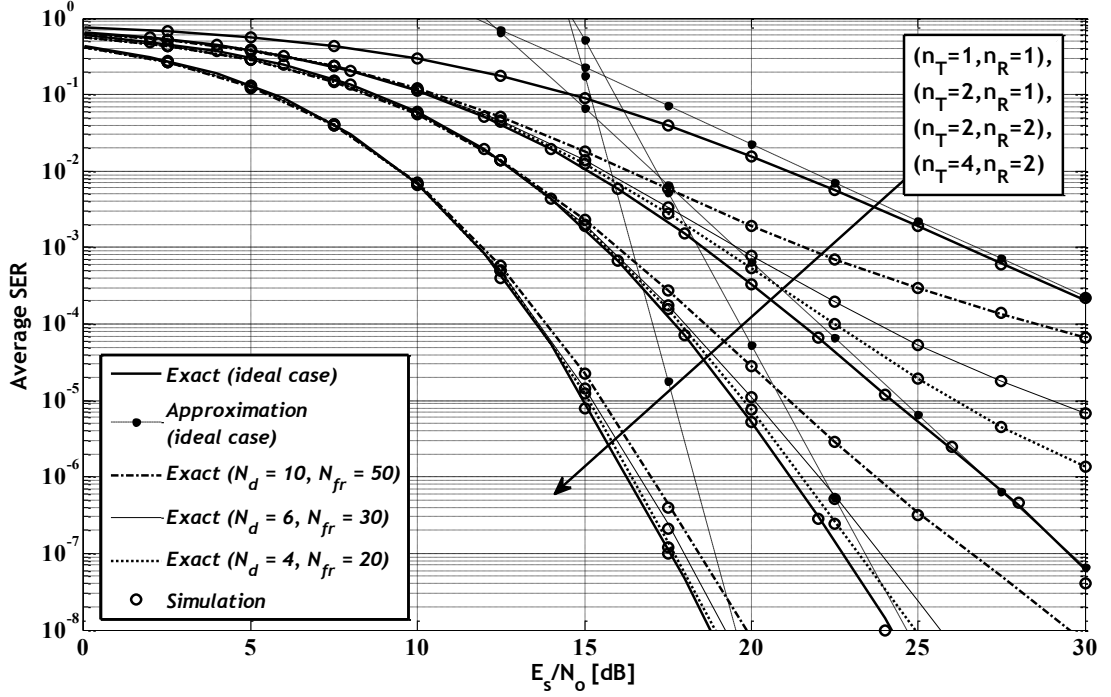


Figure 3.5: SER performance of MRT&RAS scheme in Nakagami- m fading channel ($m = 2$) for 16-QAM signals, several n_T , n_R , N_d and N_{fr} in the presence of CEE and FD ($f_D = 20$ Hertz).

MRT&RAS scheme) as shown in Figures 3.4 and 3.5 for a SER of 10^{-6} .

In Figure 3.4, for the SER curves related to the case $f_D = 20$ Hertz, the SNR degradation for the MRT&RAS scheme is almost 1 dB whereas this value is 7 dB for the pure MRT scheme. Similarly, in Figure 3.5, considering the case $N_d = 4$, $N_{fr} = 20$, the SNR degradation for the MRT&RAS scheme is less than 0.4 dB whereas this value is more than 3 dB for the pure MRT scheme.

Finally, in Figure 3.6, exact and approximate BER performances of MRT&RAS scheme using CBFSK signals are given for several n_T , n_R and m in the presence of CEE (or FQE) and FD ($f_D = 10$ Hertz, $N_d = 4$, $N_{fr} = 20$) in order to show that the ADO achieved in the erroneous channel estimation and feedback conditions is n_R . This result shows that, the diversity gain obtained by the optimal weighted transmission structure is extremely sensitive to CEE and FD, and vanishes for high SNR values. Whereas, selection diversity (RAS) structure keeps acquiring diversity gain for high SNR values despite facing driblet performance degradation.

Basically, the robustness of MRT&RAS scheme to pure MRT scheme can be easily realized by the sensitivity comparison of transmit beamforming structure (which is employed by both MRT&RAS and pure MRT) and the selection diversity

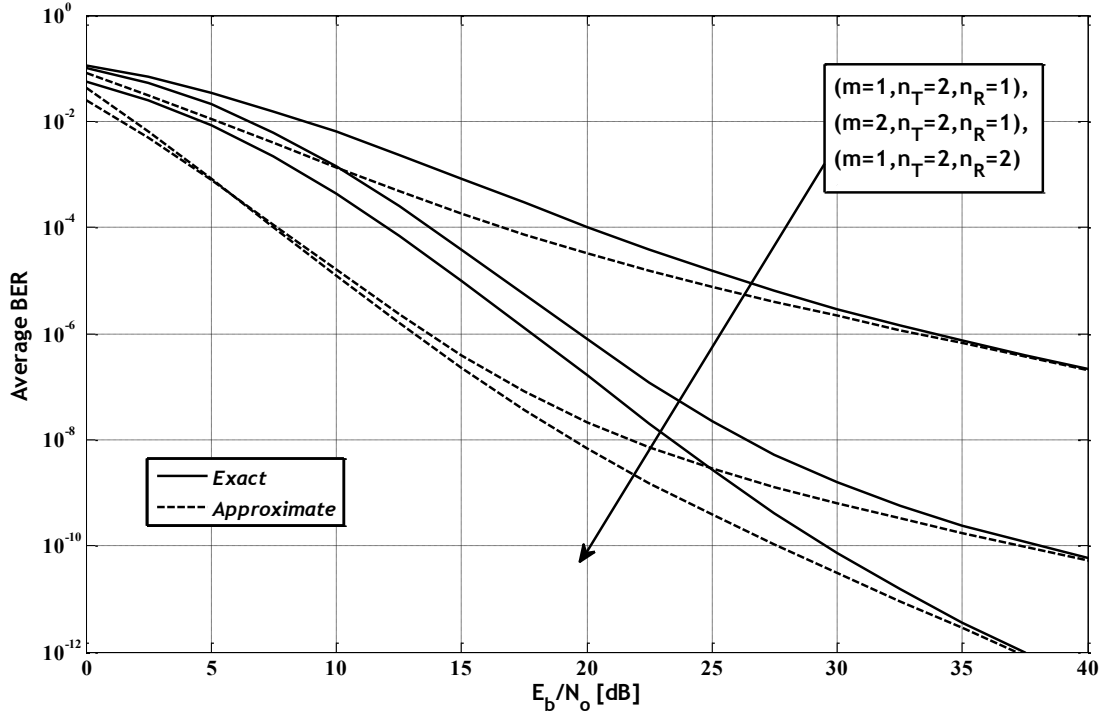


Figure 3.6: Exact and approximate BER performances of MRT&RAS scheme in Nakagami- m fading channels for CBFSK signals, several n_T , n_R and m in the presence of CEE and FD ($f_D = 10$ Hertz, $N_d = 4$, $N_{fr} = 20$).

structure. Simply, the mismatches between the channel fading coefficients $(h_{i,j}^{(k)})$ and the delayed estimates of them $(\hat{h}_{i,j}^{(k-1)})$ would cause the optimal weighted sum $(\sum_{j=1}^{n_T} h_{i,j}^{(k)} \hat{h}_{i,j}^{(k-1)*})$ to consist of extra noisy components that would seriously deteriorate the system performance for every channel realization. Whereas, the same mismatches would not always cause the selection of wrong antenna which occurs if $l \neq k$ where $l = \arg \max_{1 \leq i \leq n_R} \left\{ \sum_{j=1}^{n_T} h_{i,j}^{(k)} h_{i,j}^{(k)*} \right\}$ and $k = \arg \max_{1 \leq i \leq n_R} \left\{ \sum_{j=1}^{n_T} h_{i,j} h_{i,j}^{(k)} \hat{h}_{i,j}^{(k-1)*} \right\}$.

3.2.9. Conclusions for the Effects of CEE and FD on the Performance of MRT&RAS Scheme

Section 3.2 of the thesis focused on the performance evaluation of hybrid MRT&RAS scheme considering a generic fading model like Nakagami- m in the presence of practical impairments such as CEE (or FQE) and FD. Extensive analyses have been carried out in order to derive the exact expressions related to the statistics of the output SNR (PDF, CDF, MGF and n^{th} order SNR moments) and the capacity and error performances (ergodic capacity, OP and BER/SER) for both the ideal and the practical conditions. Also, by deriving approximate expressions for BER/SER

of several modulation techniques (e.g., DBPSK, CBFSK, M -PSK and M -QAM), it has been shown that MRT&RAS scheme achieves full diversity order (i.e., mn_Rn_T) for ICEF conditions while it keeps maintaining the receive diversity (i.e., n_R) in the presence of CEE (or FQE) and FD. As indicated in Figures 3.2, 3.4 and 3.5, the OP and BER/SER performances can easily achieve great improvements due to higher diversity orders yielded by the increase in the number of transmit antennas used for MRT and/or the number of receive antennas available for RAS in MRT&RAS scheme. Moreover, Monte Carlo simulation results have been obtained as in perfect agreement with the theoretical results.

The performance results clearly point out the superiority of MRT&RAS scheme to other diversity techniques, that is provided by the virtue of full-CSI feedback to the transmitter however this CSI requirement feature tends the MRT&RAS scheme to be accorded especially with downlink communications systems (since they can tolerate complexity at the transmitter side). Also, for reasonable values of the parameters of practical existence such as f_D , N_d and N_{fr} , MRT&RAS scheme still achieves considerable outage and error probability performances (with tolerable SNR degradation). Besides, the robustness brought along with the selection diversity structure employed at MRT&RAS scheme against practical impairments such as CEE and FD has been exposed by providing numerical results related to these cases and intuitive interpretation of different structures of transmit beamforming and selection diversity.

3.3. Analysis of Hybrid Diversity Schemes Employing Antenna Selection in the Presence of CEEs and FD¹

The main objective of the performance investigation provided in this section is to examine the effect of practical impairments such as CEE and FD on the performance of HDSs over Nakagami- m fading channels. As stated in Section 3.1, this section presents a unified framework which investigates the effect of CEEs and FD on four HDSs; three of which (i.e., TAS/OSTBC, TAS/MRT and JTRAS) have not been analyzed in the literature for the presence of such practical imperfections. The unified

¹This work is supported by the Scientific and Technological Research Council of Turkey (TÜBİTAK) under grant EEEAG-112E010

system model presented in this section also covers the hybrid MRT&RAS scheme, which has been examined in Section 3.2, by using a more realistic CEE and FD model that helps the accurate and realistic observation of the deterioration caused by CEEs and FD. Exact analytical expressions for OP and error rates of M -ary modulations are presented in order to provide an extensive examination on the capacity and error performance of HDSs that experience both CEE and FD. The ADO gathered by the investigated HDSs are derived via the well-known high-SNR approximation approach used also in Section 3.2.7. Besides, Monte Carlo simulations are performed in order to validate the theoretical results.

To the best of the authors' knowledge and based on the current literature provided in Section 3.1, the related work in the literature have considered a non-parametric fading channel model (i.e., Rayleigh fading case) [42]-[44], [53]-[57], [59], [60], only some specific modulations (e.g., BPSK/QPSK) [44], [53], [57], [60], and usually simple (non-hybrid) diversity techniques (e.g., MRC [42], OSTBC [43], MRT [55], [56]) and focused on only some specific performance metrics such as OP [42], [55], [59], error probability [43], [44], [56] via some approximations such as PEP approach [43], [54], upper bounding [54], [55], high SNR approximation [56] and numerical computations [57]. Also, the analyses provided in [57], [59] and [70] (or Section 3.2 of this thesis) are based on some approximations for the statistical characteristics and distributions of CEE and FD processes. Besides, almost all of the referred work above (except [57], [59] and [70]) have not focused on the joint examination of CEE and FD. Hence, when speaking for the HDSs TAS/OSTBC, TAS/MRT and JTRAS, there is no such examination of the exact performance considering both the CEE and FD cases for a generic fading model like Nakagami- m (even for Rayleigh fading channels). The joint model related to the deteriorating effect of CEE and FD considered throughout this section also provides an improved and accurate analysis of MRT&RAS scheme than [70] (or Section 3.2) does. Thus, investigation of HDSs facing both CEE and FD would bring along with a useful perspective for the performance examination of MIMO WCSs under realistic system and channel conditions. This part of the thesis has focused on the performance analysis of TAS/OSTBC, TAS/MRT, MRT&RAS and JTRAS schemes in the presence of both CEE and FD in i.i.d. and flat Nakagami- m fading channels. The detailed analyses provided in this presentation contribute the PDF, CDF, MGF of the output SNR, the OP and exact error probability expressions

of several digital modulations such as BPSK, QPSK, DBPSK, CBFSK, NCBFSK, M -PAM and M -QAM signals without using any approximations. The error probability expression derived through the exact error performance analysis also covers the approximate SER of high-order PSK (i.e., M -ary PSK for $M \geq 8$). The analytical performance results in the presence of CEE and FD are validated with the help of Monte Carlo simulations. Besides, the asymptotic analysis is carried out in order to derive the ADO of the investigated HDSs. Hence, this performance investigation contributes to a very important part of the related literature through providing the opportunity to extensively compare the real-world performances of different diversity schemes.

The remainder of the investigation is organized as follows. Section 3.3.1 presents the system and channel model, briefly define the statistics of the output SNR by deriving the PDF, CDF and OP expressions. Section 3.3.2 presents the derivation of the unified performance expression covering the MGF of the output SNR and exact (and approximate for M -PSK, $M \geq 8$) BER/SER performance expressions. The ADO analysis of the investigated HDSs is provided in Section 3.3.3. Section 3.3.4 presents some numerical results supported by simulations for OP and average error rates of binary and M -ary signals. Finally, Section 3.3.5 draws conclusions about the analytical and numerical results.

3.3.1. System Model and SNR Statistics for HDSs

By considering four different HDSs, i.e., TAS/OSTBC, TAS/MRT, MRT&RAS and JTRAS, through a unified system model, this part of the thesis focuses on the performance analysis over Nakagami- m fading channels in the presence of both CEEs and FD. As seen from Figure 3.7-a and 3.7-b, first two schemes have similar transmitter and receiver structures employing n_T transmitter antennas and a single receiver antenna. The detailed system descriptions are provided in [25], [32] (for TAS/OSTBC) and in [41], [60] (for TAS/MRT), respectively. The latter two schemes have similar receiver structures employing n_T transmitter and n_R receiver antennas as seen from the latter two block diagrams given in Figure 3.7-c and 3.7-d. The detailed system descriptions of these schemes are also given extensively in [70] (for MRT&RAS) and in [29] (for JTRAS) respectively. TAS/OSTBC and TAS/MRT schemes, both are

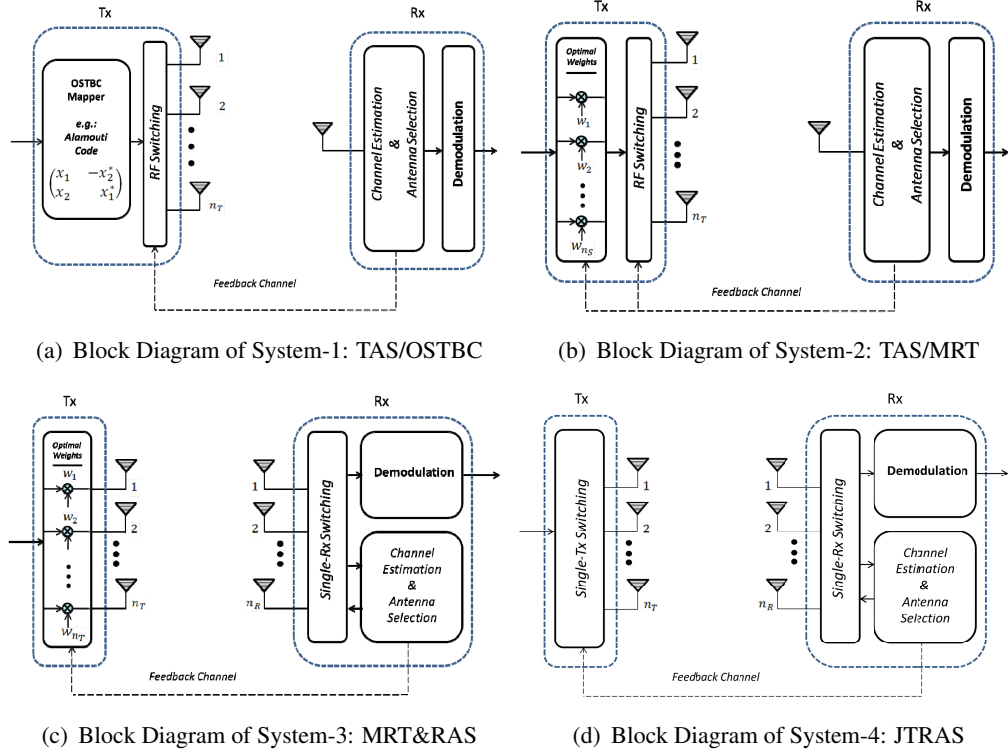


Figure 3.7: Block diagrams of the investigated diversity schemes.

examples of multi-input single-output (MISO) systems employing a single receiver antenna ($n_R = 1$), select the most efficient antenna subset with $n_S (\leq n_T)$ transmit antennas among all possible combinations by maximizing the post-processing SNR at the receiver. After selecting of the best transmit antenna subset, the transmitter performs n_S -branch OSTBC or MRT, respectively. Whereas, MRT&RAS employs n_T -branch MRT and single RAS at the transmit and receive ends, respectively, and JTRAS scheme selects and activates the optimal transmit-receive link that maximizes the output SNR. This subsection presents the output SNR statistics of each diversity scheme and defines the unified scheme model that covers all investigated schemes.

In order to help the reader to follow the flow of the derivations and the analysis, in addition to the notations and definitions provided in Section 3.2, some other notations and assumptions that are valid throughout the examination have been given:

- R_s denotes the OSTBC codeword transmission ratio (i.e., the ratio of the number of symbols to the number of transmission intervals),
- Bold-weight letters given in lowercase and uppercase (e.g., ' $\mathbf{v}$ ' and ' $\mathbf{M}$ ') represent vectors and matrices, respectively,

- $\mathbf{H}^{(k)}$ denotes the $n_R \times n_T$ complex channel gain matrix representing the MIMO channel, and $\mathbf{h}_p^{(k)}$ denotes the complex channel gain vector of the p^{th} (transmit or receive) diversity branch during the k^{th} transmission interval,
- $\hat{\mathbf{H}}^{(k)}$ denotes the estimate of the matrix $\mathbf{H}^{(k)} = \hat{\mathbf{H}}^{(k)} + \mathbf{E}_{est}$ where $\mathbf{E}_{est}$ is the estimation error matrix whose entries are zero-mean Gaussian distributed random variables with the variance of σ_{est}^2 ,
- $\|\mathbf{M}\|_F = \sqrt{\sum_i \sum_j |M_{i,j}|^2}$ denotes the Frobenius norm of the matrix $\mathbf{M}$,
- $\mathbf{H}_{(n_S:n_T)}^{(k)}$, ($n_S \leq n_T$) denotes the subset of the MIMO channel matrix for the k^{th} transmission interval that consists of the first n_S columns with the highest vector norms.

After briefly giving the assumptions and the notation that are valid throughout the analyses, the investigation focuses on the derivation of the basic statistical metrics required for the performance evaluation. By considering and modifying the output SNR expression given in [71, (11)], the instantaneous SNR at the receiver of TAS/OSTBC scheme (System-1) can be easily obtained as $\gamma_1^{(k+1)} = \left(\frac{E_s}{N_o n_S R_s} \right) \left| H_{o,(n_S:n_T)}^{(k+1)} \right|^2$ at the end of $(k+1)^{th}$ codeword transmission interval where

$$H_{o,(n_S:n_T)}^{(k+1)} = \frac{1}{\left\| \hat{\mathbf{H}}_{(n_S:n_T)}^{(k)} \right\|_F} \mathbf{H}_{(n_S:n_T)}^{(k+1)} \left(\hat{\mathbf{H}}_{(n_S:n_T)}^{(k)} \right)^H \quad (3.33)$$

Similarly, using TAS followed by MRT (i.e., System-2) would result in the instantaneous combined SNR at the receiver as shown in [60, (6)] as $\gamma_2^{(k+1)} = \left(\frac{E_s}{N_o} \right) \left| H_{o,(n_S:n_T)}^{(k+1)} \right|^2$. Thus, combining these output SNR definitions for System-1 and System-2 yields

$$\gamma_{1,2}^{(k+1)} = \left(\frac{E_s}{N_o} \right) C_{1,2} \left| H_{o,(n_S:n_T)}^{(k+1)} \right|^2 \quad (3.34)$$

while $C_1 = 1/(n_S R_s)$ and $C_2 = 1$. The total complex channel power gain $H_{o,(n_S:n_T)}^{(k+1)}$ has been shown in detail in [44] and [70] to follow a complex Gaussian distribution with the mean value of $\mu_{(n_S:n_T)} = \rho \left\| \hat{\mathbf{H}}_{(n_S:n_T)}^{(k)} \right\|_F$ and the variance of σ_T^2 :

$$H_{o,(n_S:n_T)}^{(k)} \sim \mathcal{CN}(\mu_{(n_S:n_T)}, \sigma_T^2) \quad (3.35)$$

where $\sigma_T^2 = \sigma_{est}^2 + \sigma_{fd}^2$ is the sum of the variances of complex Gaussian distributed CEE and FD processes, respectively. Here, $\sigma_{est}^2 = \frac{N_o}{N_d E_s} + 2\{\pi f_D T_s (2N_{fr} - N_d - 1)/2\}^2$ and $\sigma_{fd}^2 = (1 - |\rho|)^2 E \left[\left| \hat{h}_{i,j}^{(k)} \right|^2 \right]$, $\forall i, j, k$, as defined in [44], [57] and [70] where $\rho = J_0(2\pi f_D \tau)$ is the time correlation coefficient between the channel matrices of the successive signal transmission intervals (e.g., $\mathbf{H}^{(k)}$ and $\mathbf{H}^{(k+1)}$). The derivations presented in [57] and [70] have combined the effects of CEE and FD, and constructed the identity for the delayed estimate of the channel matrix as

$$\mathbf{H}_{(n_S:n_T)}^{(k+1)} = \rho \hat{\mathbf{H}}_{(n_S:n_T)}^{(k)} + \mathbf{E}_T \quad (3.36)$$

where $\mathbf{E}_T = \mathbf{E}_{est} + \mathbf{E}_{fd}$ is the total noisy contribution of CEE and FD processes. The analyses provided throughout the papers [57], [59] and [70] are based on the assumption that the variances $\Omega = E \left[\left| h_{i,j}^{(k)} \right|^2 \right]$ and $\hat{\Omega} = E \left[\left| \hat{h}_{i,j}^{(k)} \right|^2 \right]$ are exactly the same. However, by simply considering the identity given in (3.36), one can easily obtain the relationship between two variances as $\hat{\Omega} = \Omega + \sigma_{est}^2$. Using this value would provide to examine the practical performance of diversity schemes under erroneous conditions more accurately since using the former one (i.e., $\Omega = \hat{\Omega}$) would yield to underestimate the deterioration caused by CEE and FD phenomena.

From the viewpoint of the total channel power gain evaluation through a unified framework, MRT&RAS scheme (System-3) can be considered as if it activated all afore-selected n_T transmit antennas and performed transmission by using the outdated transmit weights despite not employing TAS in actual operation. Thus, for MRT&RAS scheme whose block diagram is given in Figure 3.7-c, the equivalent channel power gain (ECPG) of the i^{th} receive branch through all transmit antennas would be derived as $H_{3,i}^{(k+1)} = \left| H_{o,(n_T:n_T),i}^{(k+1)} \right|^2$. Here, $H_{o,(n_T:n_T),i}^{(k+1)}$ has the same statistical characteristics as (3.35) while the term i denotes the index of the receive branch. Then, the single receiver antenna, that maximizes the total channel power gain (also accordingly the output SNR) at the demodulator input, is selected and the instantaneous SNR at the end of the $(k+1)^{th}$ transmission interval becomes:

$$\gamma_3^{(k+1)} = \left(\frac{E_s}{N_o} \right) \max_{1 \leq i \leq n_R} \left\{ H_{3,i}^{(k+1)} \right\} \quad (3.37)$$

Similarly, for the fourth scheme included at this section's focus (i.e., JTRAS), the ECPG of the i^{th} receive branch could easily be obtained by modifying that of MRT&RAS. Since, JTRAS scheme activates a single afore-selected transmit antenna, the complex sub-channel that was introduced by the selected transmit antenna would be distorted according to (3.35) and would cause impairments at the received signal. Hence, the ECPG of the i^{th} receive branch can be expressed as $H_{4,i}^{(k+1)} = \left| H_{o,(1:n_T),i}^{(k+1)} \right|^2$ which yields the output SNR expression of the same receive branch as

$$\gamma_4^{(k+1)} = \left(\frac{E_s}{N_o} \right) \max_{1 \leq i \leq n_R} \left\{ H_{4,i}^{(k+1)} \right\} \quad (3.38)$$

Therefore, considering each output SNR expressions derived up to this point and representing them by a generic expression results in the unified output SNR covering four HDSs

$$\gamma_w^{(k+1)} = \frac{E_s}{N_o} C_w \max_{1 \leq p \leq N} \left| H_{o,(M:K),p}^{(k+1)} \right|^2, p = 1, 2, \dots, N \quad (3.39)$$

where $w \in \{1, 2, 3, 4\}$ denotes the system index for TAS/OSTBC, TAS/MRT, MRT&RAS and JTRAS schemes, respectively. The parameters in (3.39) are defined as $(N, M, K) = (1, n_S, n_T)$ for $w = 1, 2$, $(N, M, K) = (n_R, n_T, n_T)$ for $w = 3$ and $(N, M, K) = (n_R, 1, n_T)$ for $w = 4$. Also, $C_1 = 1/(n_S R_s)$, for $w = 1$ and $C_w = 1$ for $w \neq 1$. By using (3.39), the statistical characteristics of the output SNRs have been derived which yield to examine the OP and BER/SER performances. The squared norm of the p^{th} complex channel power gain $H_{o,(M:K),p}^{(k+1)} \sim \mathcal{CN}(\mu_{(M:K)}, \sigma_T^2)$, $\forall p = 1, 2, \dots, N$, has been shown in [59] and [70] to follow a non-central Chi-square distribution with a degree of freedom of 2 and the non-centrality parameter of $v_{(M:K)} = \mu_{(M:K)}^2 = \rho^2 \left\| \hat{\mathbf{H}}_{(M:K)}^{(k)} \right\|_F^2$. The Chi-square distributed norm has been represented by the PDF expression below:

$$f_p(x|v) = \frac{e^{-\frac{x+v}{\sigma_T^2}}}{\sigma_T^2} I_0 \left(\frac{2\sqrt{vx}}{\sigma_T^2} \right) = \frac{e^{-\frac{x+v}{\sigma_T^2}}}{\sigma_T^2} {}_0F_1 \left(; 1; \frac{vx}{\sigma_T^4} \right) \quad (3.40)$$

In (3.40), $I_0(\cdot)$ and ${}_0F_1(\cdot; \cdot; \cdot)$ denote the zeroth-order modified Bessel function of the first kind [62, (2.1-118)] and the confluent hypergeometric limit function [63, (02)], respectively. Integrating (3.40) would result in the CDF expression of the p^{th} complex channel power gain as

$$F_p(x|v) = e^{-\frac{v}{\sigma_T^2}} \sum_{q=0}^{\infty} \frac{(v/\sigma_T^2)^q}{(q!)^2} \psi\left(q+1, \frac{x}{\sigma_T^2}\right), x \geq 0 \quad (3.41)$$

where $\psi(\cdot, \cdot)$ is the incomplete Gamma function [62, (8.350.1)]. In order to resolve the condition of (3.40) and (3.41) on the parameter v , both expressions should be averaged over the distribution of v . The PDF of the non-centrality parameter can be obtained by performing the Inverse Laplacian operator over the MGF of this variable that is given as

$$\begin{aligned} \mathcal{M}_v^{(p)}(s) = & K! \left(\frac{m}{\rho^2 \hat{\Omega}}\right)^{mK} \sum_{\mathbf{U}^{(p)}} \mathcal{K}_1^{(p)} \mathcal{K}_2^{(p)} \sum_{j_p=M}^K \sum_{i_p=1}^{1+a_{j_p}} \beta_{j_p, i_p}^{(p)} \\ & \times \left(s + \frac{m}{\rho^2 \hat{\Omega}}\right)^{-\alpha_p} \left(s + \frac{m j_p}{\rho^2 \hat{\Omega} M}\right)^{-i_p} \end{aligned} \quad (3.42)$$

in [41, (15)]. The parameters $\mathcal{K}_1^{(p)}$, $\mathcal{K}_2^{(p)}$, $\beta_{j_p, i_p}^{(p)}$, $\mathbf{U}^{(p)}$, a_{j_p} and α_p given in (3.42) have been presented in detail in [41]. By using the Laplace transform pair given in [72, (2.1.4-8)], one can easily derive the PDF expression of the non-centrality parameter v as

$$\begin{aligned} f_v(x; \rho, \hat{\Omega}) = & K! \left(\frac{m}{\rho^2 \hat{\Omega}}\right)^{mK} \sum_{\mathbf{U}^{(p)}} \mathcal{K}_1^{(p)} \mathcal{K}_2^{(p)} \sum_{j_p=M}^K \sum_{i_p=1}^{1+a_{j_p}} \beta_{j_p, i_p}^{(p)} \\ & \times \frac{x^{\alpha_p + i_p - 1}}{\Gamma(\alpha_p + i_p)} {}_1F_1\left(i_p; \alpha_p + i_p; \frac{m x}{\rho^2 \hat{\Omega} M} (M - j_p)\right) \end{aligned} \quad (3.43)$$

where ${}_1F_1(a; b; z) = \sum_{r=0}^{\infty} \frac{\Gamma(a+r)\Gamma(b)}{\Gamma(a)\Gamma(b+r)} \frac{z^r}{r!}$ denotes the confluent hypergeometric function [62, (210.1)]. Substituting the series expansion of ${}_1F_1$ function into (3.43) and averaging the CDF expression given in (3.41) over $f_v(x; \rho, \hat{\Omega})$ will resolve the conditional dependency of (3.41) to v as

$$F_p(x; \rho, \hat{\Omega}) = \int_0^\infty F_p(x|v) f_v(x; \rho, \hat{\Omega}) dv = \mathcal{K}_0 \sum_{\mathbf{U}^{(p)}} \mathcal{K}_1^{(p)} \times \mathcal{K}_2^{(p)} \sum_{j_p} \sum_{i_p} \beta_{j_p, i_p}^{(p)} \sum_{q_p} \sum_{r_p} \mathcal{K}_3^{(p)} \psi\left(q_p + 1, \frac{x}{\sigma_T^2}\right) \quad (3.44)$$

where $\mathcal{K}_0 = K! \left(\frac{m}{\rho^2 \hat{\Omega}}\right)^{mK}$ and $\mathcal{K}_3^{(p)} = \left\{ \frac{m(M-j_p)}{\rho^2 \hat{\Omega} M} \right\}^{r_p} \frac{\Gamma(i_p + r_p)}{\Gamma(q_p + 1)^2} \frac{(\sigma_T^2)^{r_p + i_p + \alpha_p - q_p}}{\Gamma(i_p) \Gamma(1 + r_p)}$. Since the unified channel power gain would be obtained as the maximum of all receive branches according to (3.39), the CDF of the unified channel power gain is derived as [75]

$$F_w(x; \rho, \hat{\Omega}) = \prod_{p=1}^N F_p(x; \rho, \hat{\Omega}) = \mathcal{K}_0^N \sum_{\mathbf{U}^{(1)}} \sum_{\mathbf{U}^{(2)}} \cdots \sum_{\mathbf{U}^{(N)}} \left\{ \prod_{p=1}^N \mathcal{K}_1^{(p)} \mathcal{K}_2^{(p)} \right\} \sum_{j_1} \sum_{j_2} \cdots \sum_{j_N} \times \sum_{i_1} \cdots \sum_{i_N} \left\{ \prod_{p=1}^N \beta_{j_p, i_p}^{(p)} \right\} \sum_{q_1} \cdots \sum_{q_N} \sum_{r_1} \cdots \sum_{r_N} \left\{ \prod_{p=1}^N \mathcal{K}_3^{(p)} \psi\left(1 + q_p; \frac{x}{\sigma_T^2}\right) \right\} \quad (3.45)$$

$$= \bar{\mathcal{K}}_0 \sum_{\bar{\mathbf{U}}} \bar{\mathcal{K}}_1 \bar{\mathcal{K}}_2 \sum_{\bar{\mathcal{J}}} \sum_{\bar{\mathcal{I}}} \bar{\beta}_{\bar{\mathcal{J}}\bar{\mathcal{I}}} \sum_{\bar{\mathcal{Q}}} \sum_{\bar{\mathcal{R}}} \bar{\mathcal{K}}_3 \prod_p \psi\left(1 + q_p; \frac{x}{\sigma_T^2}\right)$$

In (3.45), the parameters with overbars represent the products (e.g., $\bar{\mathcal{T}}_d = \prod_{p=1}^N \mathcal{T}_d^{(p)}$, $d = 1, 2, 3$), and sets (e.g., $\sum_{\bar{\mathcal{S}}} = \sum_{s_1} \sum_{s_2} \cdots \sum_{s_N}$) over all values of $p = 1, 2, \dots, N$. Using the identity $\psi(c_1, c_2 t) = \frac{{}_1F_1(1; 1 + c_1; c_2 t)}{c_1 (c_2 t)^{-c_1}} e^{-c_2 t}$ given in [62, (8.351.2)], (3.45) can be expressed as [75];

$$F_w(x; \rho, \hat{\Omega}) = \bar{\mathcal{K}}_0 \sum_{\bar{\mathbf{U}}} \bar{\mathcal{K}}_1 \bar{\mathcal{K}}_2 \sum_{\bar{\mathcal{J}}} \sum_{\bar{\mathcal{I}}} \bar{\beta}_{\bar{\mathcal{J}}\bar{\mathcal{I}}} \sum_{\bar{\mathcal{Q}}} \sum_{\bar{\mathcal{R}}} \bar{\mathcal{K}}_4 \times x^{N + \sum_{p=1}^N q_p} e^{-xN/\sigma_T^2} \prod_{p=1}^N {}_1F_1\left(1; 2 + q_p; \frac{x}{\sigma_T^2}\right) \quad (3.46)$$

where the parameter is $\bar{\mathcal{K}}_4 = \bar{\mathcal{K}}_3 \prod_{p=1}^N \frac{(\sigma_T^2)^{-1 - q_p}}{1 + q_p}$. Note that, however, (3.45) and (3.46) define the statistical characteristics of the unified ECPG $G_w = \max_{1 \leq p \leq N} |H_{o, (M:K), p}^{(k+1)}|^2$, $\forall k, w \in \{1, 2, 3, 4\}$, these identities can be easily modified to express the output SNR characteristics by simply substituting $\hat{\Omega} \bar{\gamma}_w$ into (3.45) and (3.46) instead of the parameter $\hat{\Omega}$ where $\bar{\gamma}_w = \frac{E_s}{N_o} C_w$.

The output SNR characteristics would not only pave the way for the analytical derivation of the exact error performances but also provides the direct OP evaluation through the well-known identity $P_{w, out} = F_w(2^{\mathcal{R}} - 1; \rho, \hat{\Omega} \bar{\gamma}_w)$ as shown in [21], [29],

[55], [59], [70] and [41]. Here, $\mathcal{R}$ denotes the desired spectral efficiency in bit/s/Hertz. Note that, the incomplete Gamma function which was used in the CDF expression in (3.45) can be easily evaluated by using the built-in functions of the well-known software programs such as MATHEMATICA and MATLAB.

3.3.2. Exact Error Performance Analysis for HDSs

Based on the statistics of the output SNR related to the investigated HDSs, this section derives the exact performance expressions of the MGF and average BER/SER which are the most effective metrics for examining the performance of WCSs. For this purpose, the unified integral representations given in [73, (26)] and [74, (7)], respectively as

$$\mathcal{J}_{\gamma_w}(\theta, \epsilon, \Phi) = \theta \int_0^\infty x^\epsilon e^{-x\Phi} F_w\left(x; \rho, \hat{\Omega}\bar{\gamma}_w\right) dx \quad (3.47)$$

and

$$\hat{\mathcal{J}}_{\gamma_w}(\theta, \Phi) = \theta \int_0^\infty e^{-x\Phi} F_w\left(x; \rho, \hat{\Omega}\bar{\gamma}_w\right) {}_1F_1\left(1; \frac{3}{2}; \frac{x\Phi}{2}\right) dx \quad (3.48)$$

are used. Here, setting different values in the parameters of (3.47) and (3.48) (i.e., θ, ϵ and Φ) will result in the MGF and the BER/SER of some binary/ M -ary modulations as defined in [73], [74]. By considering both (3.37) and (3.48), it is possible to evaluate the MGF of the output SNR by using the identity $M_{\gamma_w}(s) = \mathcal{J}_{\gamma_w}(s, 0, s)$, the average BER of BPSK, CBFSK, NCBFSK and DBPSK signals and the approximate SER of M -PSK signals by using the identity $P_{w,s} = \mathcal{J}_{\gamma_w}\left(\frac{\lambda_3(\lambda_2^{\lambda_1})}{2\Gamma(\lambda_1)}, \lambda_1 - 1, \lambda_2\right)$, SER of M -PAM signals by using the identity $P_{w,s} = \mathcal{J}_{\gamma_w}\left(\sqrt{\frac{3(M-1)}{\pi M^2(M+1)}}, \frac{-1}{2}, \frac{3}{(M^2-1)}\right)$ and SER of QPSK and M -QAM signals by using the identity $P_{w,s} = \mathcal{J}_{\gamma_w}\left(\sqrt{\frac{\lambda_4}{8\pi}}(\lambda_5 - \lambda_6), \frac{-1}{2}, \frac{\lambda_4}{2}\right) + \hat{\mathcal{J}}_{\gamma_w}\left(\frac{\lambda_4\lambda_6}{2\pi}, \lambda_4\right)$. The modulation-dependent parameters for binary and M -ary cases are listed below in Table 3.1. Thus, deriving the exact expressions of (3.47) and (3.48)

will lead us to obtain the performance metrics of the investigated HDSs for several modulations [75]. By substituting (3.46) into (3.47) and (3.48), and rearranging the resulting integrals in terms of the Lauricella function of n variables described in [65,

Table 3.1: Parameters defined in the average BER/SER expressions of binary and M -ary modulations.

	BPSK	CBFSK	NCBFSK	DBPSK	QPSK	M -PSK (Approx.)	M -QAM
λ_1	0.5	0.5	1	1	-	0.5	-
λ_2	1	0.5	0.5	1	-	$\sin^2(\pi/M)$	-
λ_3	1	1	1	1	-	2	-
λ_4	-	-	-	-	1	-	$\frac{3}{M-1}$
λ_5	-	-	-	-	2	-	$4 - \frac{4}{\sqrt{M}}$
λ_6	-	-	-	-	1	-	$(2 - \frac{2}{\sqrt{M}})^2$

(2.4.2)], the unified integral expressions in (3.47) and (3.48) can be respectively derived as [75];

$$\begin{aligned} \mathcal{J}_{\gamma_w}(\theta, \epsilon, \Phi) &= \theta \bar{\mathcal{K}}_0 \sum_{\bar{\mathbf{U}}} \bar{\mathcal{K}}_1 \bar{\mathcal{K}}_2 \sum_{\bar{\mathcal{J}}} \sum_{\bar{\mathcal{I}}} \bar{\beta}_{\bar{\mathcal{J}}\bar{\mathcal{I}}} \sum_{\bar{\mathcal{Q}}} \sum_{\bar{\mathcal{R}}} \frac{\bar{\mathcal{K}}_4 \Gamma(\mathcal{B}(\epsilon))}{\left(\frac{\mathcal{D}(\Phi)}{\sigma_T^2}\right)^{\mathcal{B}(\epsilon)}} \\ &\times F_A^{(N)}\left(\mathcal{B}(\epsilon); 1, \dots, 1; b_1, \dots, b_N; \frac{1}{\mathcal{D}(\Phi)}, \dots, \frac{1}{\mathcal{D}(\Phi)}\right) \end{aligned} \quad (3.49)$$

and

$$\begin{aligned} \hat{\mathcal{J}}_{\gamma_w}(\theta, \Phi) &= \theta \bar{\mathcal{K}}_0 \sum_{\bar{\mathbf{U}}} \bar{\mathcal{K}}_1 \bar{\mathcal{K}}_2 \sum_{\bar{\mathcal{J}}} \sum_{\bar{\mathcal{I}}} \bar{\beta}_{\bar{\mathcal{J}}\bar{\mathcal{I}}} \sum_{\bar{\mathcal{Q}}} \sum_{\bar{\mathcal{R}}} \frac{\bar{\mathcal{K}}_4 \Gamma(\mathcal{B}(0))}{\left(\frac{\mathcal{D}(\Phi)}{\sigma_T^2}\right)^{\mathcal{B}(0)}} \\ &\times F_A^{(N+1)}\left(\mathcal{B}(0); 1, \dots, 1; \frac{3}{2}, b_1, \dots, b_N; \frac{0.5\Phi}{\mathcal{D}(\Phi)}, \frac{1}{\mathcal{D}(\Phi)}, \dots, \frac{1}{\mathcal{D}(\Phi)}\right) \end{aligned} \quad (3.50)$$

where $\mathcal{B}(\epsilon) = 1 + \epsilon + N + \sum_{p=1}^N q_p$, $\mathcal{D}(\Phi) = \Phi \sigma_T^2 + N$, $b_p = \alpha_p + i_p + r_p$, $p = 1, 2, \dots, N$. The Lauricella function $F_A^{(n)}(a; b_1, \dots, b_n; c_1, \dots, c_n; x_1, \dots, x_n)$ used in (3.49) and (3.50) satisfies the convergence criterion $\sum_{i=1}^n |x_i| < 1$ for possible values of its arguments and are evaluated by using the proper integral definition in [65, (2.3.3)]. Also, [21] provides an efficient way for the numerical evaluation of this function by using the Gauss-Laguerre integration method.

3.3.3. Asymptotic Diversity Order Analysis for HDSs

Since exact expressions of the performance metrics given in (3.45), (3.46), (3.47) and (3.48) do not give much information about the ADO, approximate expressions for the PDF/CDF of the output SNR are derived by using the high-SNR approximation technique given in [69] which will enable us to observe the ADO of the investigated HDSs clearly. This approximation technique requires the asymptotic representation of the PDF of the ECPG at the receiver-output G_w as $f_{G_w}(x) = ax^t + o(x^t)$, $a > 0$.

In order to obtain the asymptotic PDF expression $f_{G_w}(x)$, the PDF/CDF of each diversity branch has been derived asymptotically. Considering the series expansion of the incomplete Gamma function given in [62, (8.354.1)], one can easily obtain the approximate expression as

$$\psi(n, z) = \frac{z^n}{n} + o(z^n), z \rightarrow 0 \quad (3.51)$$

By substituting (3.51) into (3.41), the CDF expression related to the p^{th} diversity branch would be expressed as

$$F_p(x|v) = e^{-\frac{v}{\sigma_T^2}} \sum_{q=0}^{\infty} \frac{(v/\sigma_T^2)^q}{(q!)^2} \left\{ \frac{\frac{x}{\sigma_T^2}^{q+1}}{q+1} + o(x^{q+1}) \right\}, x \geq 0 \quad (3.52)$$

Also, using the series expansion of the term e^{-v/σ_T^2} and rewriting the CDF expression in (3.52) by considering the fact that the whole expression is almost dominated by the polynomial component with the smallest exponent yields

$$F_p(x|v) = \frac{x}{\sigma_T^2} + o\left(\frac{x}{\sigma_T^2}\right) = F_p(x), x \rightarrow 0 \quad (3.53)$$

where the dependency on the non-centrality parameter v has also vanished. Using the approximate CDF expression obtained in (3.53), the CDF of the overall channel power gain can be derived by using the product definition as $F_w(x) = \prod_{p=1}^N F_p(x) = x^N \sigma_T^{-2N} + o(x^N)$. Finally, derivative of $F_w(x)$ will result in the approximate PDF expression of the overall channel power gain [75]:

$$f_w(x) = N\sigma_T^{-2N}x^{N-1} + o(x^{N-1}), x \rightarrow 0 \quad (3.54)$$

As a result, for the unified diversity scheme the parameters are found as $a = N\sigma_T^{-2N}$ and $t = N - 1$. Thus, by using these parameters, the asymptotic BER/SER and ADO of all schemes can be easily obtained. Note that the first two schemes (i.e., TAS/OSTBC and TAS/MRT) achieve a signal transmission/reception with an ADO of $N = 1$ whereas, the third and fourth schemes (i.e., MRT&RAS and JTRAS) achieve a signal transmission/reception with an ADO of $N = n_R$.

3.3.4. Numerical Results for HDSs in the Presence of CEEs and FD

In order to examine the performance of each HDS for the erroneous conditions in Nakagami- m fading channels, this subsection presents some numerical results in this section consisting of theoretical and simulation results of miscellaneous system and channel conditions for unit channel power ($\Omega = 1$), total transmit energy of unity ($E_s = 1$) (or $E_b = 1$, which is average energy per bit). In Figures 3.8, 3.9, 3.10 and 3.12, the values of N_{fr} and N_d have been set to typical values such as $N_{fr} = 20$ and $N_d = 4$ that are consistent with the 3GPP proposals [58], whereas Figure 3.11 has provided the variation of the average SER performance due to varying values of $N_d \in [1, 10]$ while $N_{fr} = 20$.

For the performance consideration of the first two HDSs that have similar transmitter structures, the numerical results related to TAS/OSTBC scheme are exposed since there is a deterministic performance difference between TAS/OSTBC and TAS/MRT schemes depending on the number of active transmit antennas. When compared to any particular OP/BER/SER performance of TAS/OSTBC scheme, TAS/MRT scheme would be succeeding an SNR advantage of $10\log_{10}(n_S)$ dB in the high SNR region. Having this important remark made, the OP performance of TAS/OSTBC scheme ($w = 1$) has been depicted in Figure 3.8 for $m = 2$, $\mathcal{R} = 2$ bit/s/Hertz, $n_R = 2$, $N_d = 4$, $N_{fr} = 20$ in order provide a realistic view on the system capacity. The theoretical OP results of hybrid TAS/two-branch ($n_S = 2$) and three-branch ($n_S = 3$) OSTBC schemes are evaluated via (3.45) and validated by performing Monte Carlo simulations as seen in Figure 3.8. For $f_D = 20$ Hertz and $f_D = 50$ Hertz, it has been shown that the OP performances are worse than the ideal

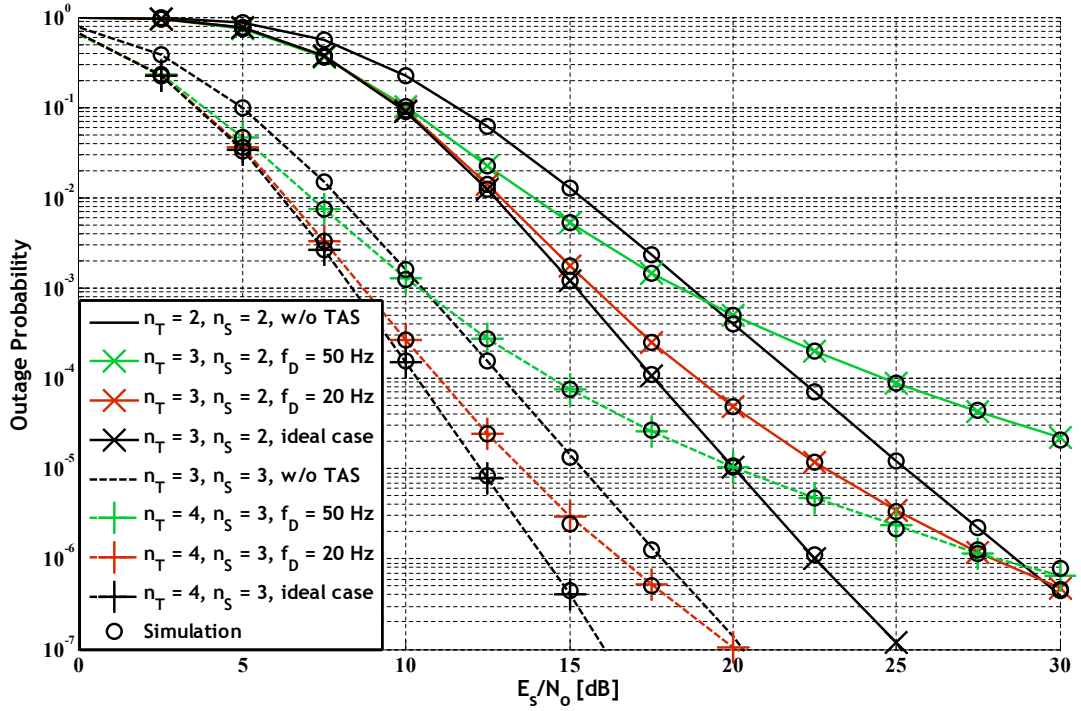


Figure 3.8: Average OP performance of System-1: TAS/OSTBC Scheme for the parameters $m = 2$, $\mathcal{R} = 2$ bit/s/Hertz, $n_R = 2$, $N_d = 4$, $N_{fr} = 20$.

channel estimation and feedback cases and the degradation on the OP performances becomes severe with the increasing values of f_D . As a consequence of the degradation caused by the wireless channel's Doppler frequency shift, the erroneous OP curves intersect with those corresponding to cases without TAS for an average SNR value. From the curves, one can easily see that beyond this critical SNR value, the practical cases face worse OP performances than those of ideal cases without TAS. The OP curves related to erroneous conditions $f_D = 20$ Hertz and $f_D = 50$ Hertz of the scenario $\{n_T = 3, n_S = 2\}$ intersect the ideal (TAS-free) scenario at the average SNR values of approximately 29 dB and 19 dB, respectively. These critical SNR values can be seen to be approximately 20 dB and 11 dB for the scenario $\{n_T = 4, n_S = 3\}$. For both scenarios $\{n_T = 3, n_S = 2\}$ and $\{n_T = 4, n_S = 3\}$, the investigated HDSs (TAS/ G_{n_S} -OSTBC and TAS/ n_S -branch MRT) would still achieve variable amount of SNR gain when compared to the pure OSTBC and MRT schemes (G_{n_S} -OSTBC and n_S -branch MRT) whereas for SNR values greater than the critical values stated above, the degradation caused by the CEE and FD effects would vanish the SNR and diversity gain advantage of the transmit diversity (yielded by TAS, OSTBC, MRT schemes and the scattering environment) and deteriorates the overall performance.

Figure 3.9 depicts the average SER performance of the fourth HDS (i.e., JTRAS scheme) for QPSK signals and two receive antennas ($n_R = 2$) in Rayleigh fading channels ($m = 1$). By examining the SER curves, it can be seen that for a given number of transmit antennas (n_T), the SER performances get worse with the increasing values of f_D . For the scenario $\{n_T = 3\}$ and a SER of 10^{-6} , the SNR values corresponding to the $f_D = 20$ Hertz and $f_D = 50$ Hertz cases are approximately 0.35 dB and 2 dB greater than those that correspond to the ideal (channel estimation and feedback) cases, respectively. The SNR degradation values are seen to be 0.5 dB and 1.8 dB, respectively for the scenario $\{n_T = 4\}$. Also, the slope of the SER curves for the erroneous cases especially in the high SNR region indicates that the ADO achieved in the presence of CEE and FD for both scenarios (i.e., $\{n_T = 3\}$ and $\{n_T = 4\}$) is equal to $n_R = 2$ as consistent with the ADO analysis.

The average SER performance of the third HDS (i.e., MRT&RAS scheme) that has a similar receiver structure to that of the fourth one, is exemplified in Figure 3.10 for One-Sided Gaussian fading channels ($m = 0.5$) and 16-QAM signals. When the SER curves related to the four-branch ($n_T = 4$) MRT scheme for one and two receive antennas are examined, the effect of increasing f_D on the SER performance can be seen to cause increasing performance degradation as it was in Figures 3.8 and 3.9. For the case of no RAS (i.e., pure MRT with $n_R = 1$) and a SER of 10^{-3} , the SNR degradations corresponding to $f_D = 20$ Hertz and $f_D = 50$ Hertz Doppler frequency shift cases are 1.05 dB and 4.55 dB, respectively when compared to the ideal SER performance. The degradation values become 0.07 dB and 0.46 dB, respectively when the RAS is employed at the receive side. This points out the performance improvement and robustness in the presence of CEE and FD achieved by employing RAS together with MRT as consistent with the outcomes of [41].

In order to examine and compare the system performances of the investigated HDSs in practical channel estimation and feedback conditions, the examination has focused on the variation of average SER performances of the first two schemes (i.e., TAS/OSTBC and TAS/MRT), which have similar transmitter structures, depending on the variation of N_d , and the variation of average SER performances of the third and fourth schemes (i.e., MRT&RAS and JTRAS), which have similar receiver structures, depending on the variation of f_D . Figure 3.11 depicts the SER performance comparison of TAS/OSTBC and TAS/MRT schemes in the presence of CEE and FD

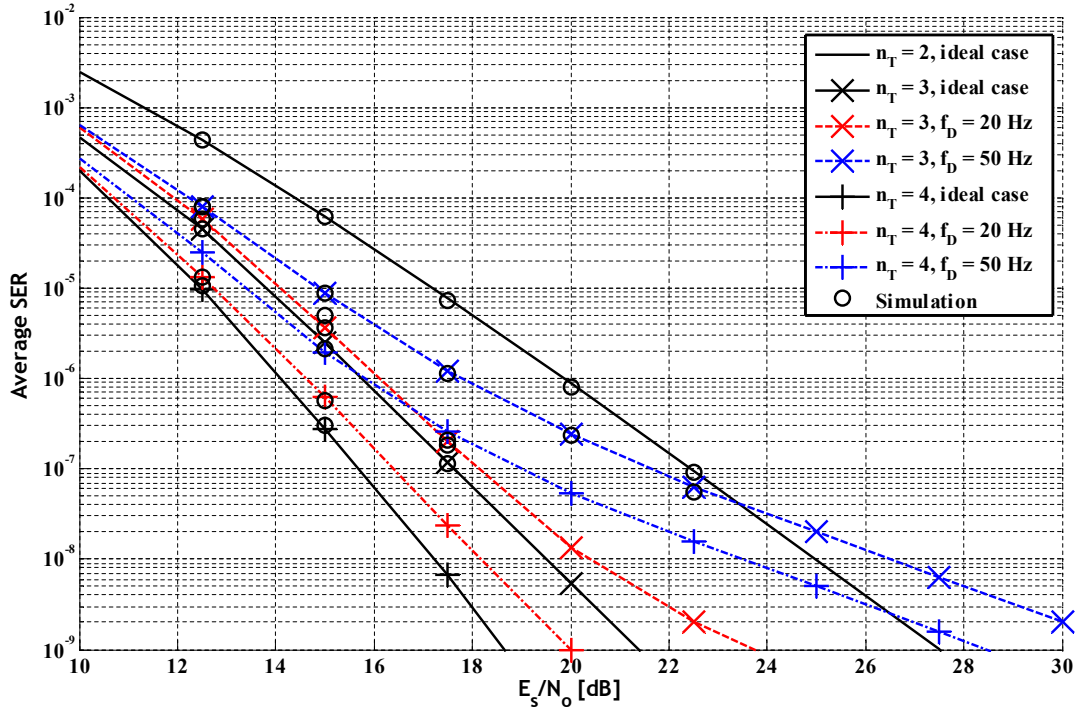


Figure 3.9: Average SER performance of QPSK Signals for System-4: JTRAS Scheme and for the parameters $m = 1$, $n_R = 2$, $N_d = 4$, $N_{fr} = 20$.

for Rayleigh fading channels ($m = 1$), BPSK signals, $n_T = 3$, $n_S = 2$, $f_D = 15$ Hertz, $N_{fr} = 20$ and several average SNR values $\left(\frac{E_s}{N_o}\right)_{dB} \in [8, 12]$, with respect to different values of N_d . SER curves not only confirm the important remark that was made about the deterministic performance difference of TAS/OSTBC and TAS/MRT schemes, but also show that both schemes perform better with the increasing values of N_d . As the most important result of the performance comparison, the SER variation of both schemes points out that TAS/MRT scheme achieves more SER performance improvement than TAS/OSTBC scheme does as the number of pilot symbols increases.

Similar to the comparison made in Figure 3.11, the SER performances of MRT&RAS and JTRAS schemes have been taken under consideration in Figure 3.12 with respect to several values of f_D for Nakagami- m fading channels with the parameter $m = 1.5$, QPSK signals, $n_T = 2$, $n_R = 2$ and average SNR values of $\left(\frac{E_s}{N_o}\right)_{dB} \in \{8, 12, 16\}$. By examining Figure 3.12, it can be seen that the SER degradation caused on the performance of both schemes due to the increase in the value of f_D becomes more severe as the average SNR value increases. In other words, both schemes become more delicate to the increment of f_D for higher average SNR values. Also, when comparing two schemes for the investigated system and channel conditions

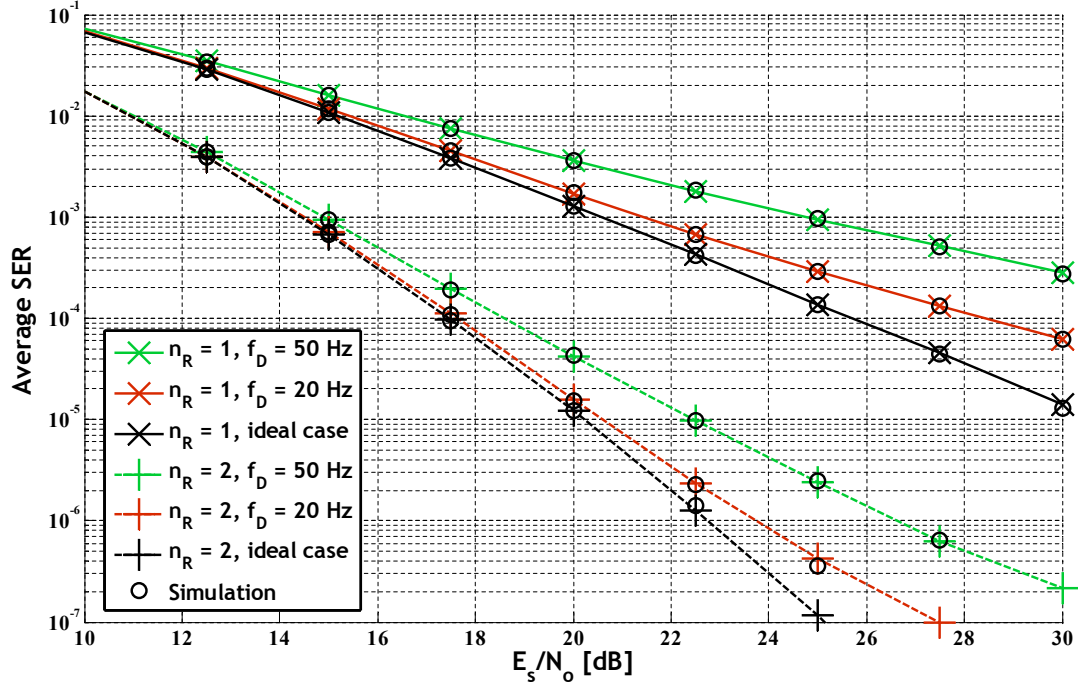


Figure 3.10: Average SER performance of 16-QAM Signals for System-3: MRT&RAS Scheme and for the parameters $m = 0.5$, $n_T = 4$, $N_d = 4$, $N_{fr} = 20$.

in Figure 3.12, the performance degradation of MRT&RAS scheme is clearly seen to be lower than that of JTRAS scheme which points out the robust structure of MRT&RAS scheme to increasing values of f_D than that of JTRAS scheme. The results in Figures 3.11 and 3.12 show the robustness of HDSs to CEE and FD while the same diversity systems may not be robust against other practical impairments such as FE [74].

3.3.5. Conclusions for the Effects of CEE and FD on the Performance of HDSs

This section has focused on the performance evaluation of four different HDSs (TAS/OSTBC, TAS/MRT, MRT&RAS and JTRAS schemes) considering a generic fading model like Nakagami- m in the presence of practical impairments such as CEE and FD. Extensive analyses have been carried out in order to derive the exact expressions related to the statistics of the output SNR (PDF, CDF and MGF) and the capacity and error performances (OP and BER/SER) for all schemes in erroneous conditions. Also, with the help of ADO analysis, it is shown that the unified scheme achieves an ADO of N which is shown to be unity for the first two (MISO) schemes

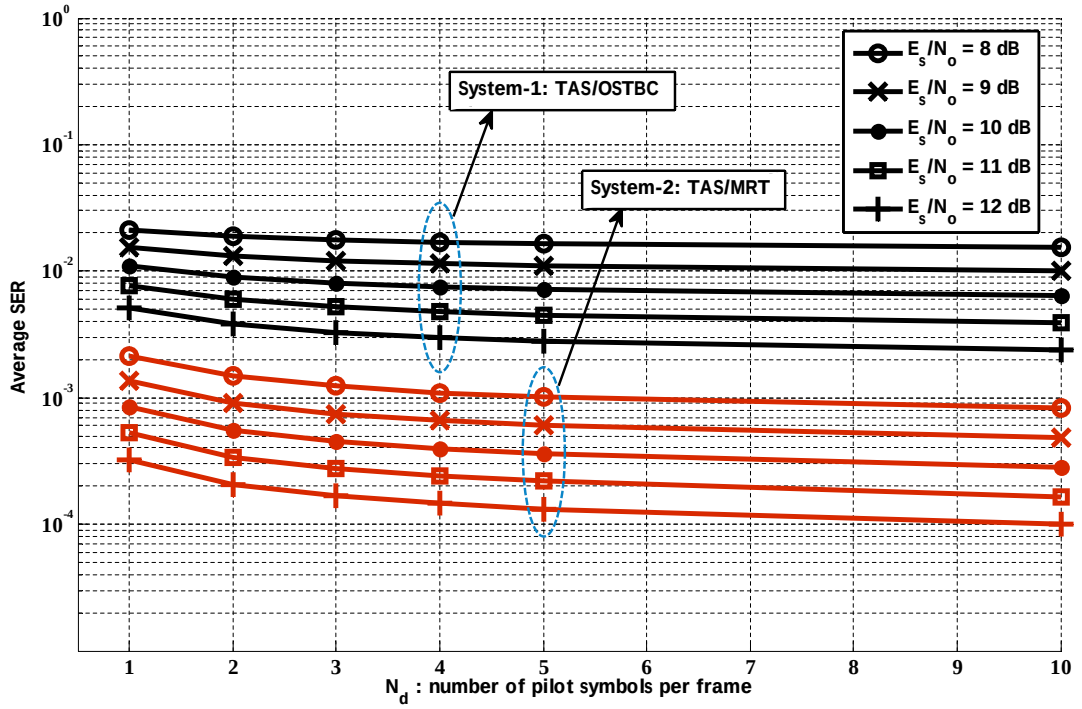


Figure 3.11: Variation of the SER performances of System-1: TAS/OSTBC and System-2: TAS/MRT depending on the number of pilot symbols per frame (N_d), for BPSK Signals and the parameters $m = 1$, $n_T = 3$, $n_S = 2$, $f_D = 15$ Hertz, $N_{fr} = 20$.

(i.e., TAS/OSTBC and TAS/MRT) and $N = n_R$ for the latter two (MIMO) schemes (i.e., MRT&RAS and JTRAS). As indicated in Figures 3.8 and 3.9, for the ideal conditions of all schemes, the OP and BER/SER performances can easily achieve great improvements due to higher diversity orders yielded by the increase in the number of transmit and/or receive antennas. Also, for the erroneous cases, the increase in the number of transmit and/or receive antennas provides SER performance enhancements in the low SNR region. All the performance results indicate that each HDS more or less faces performance degradations caused by CEE and FD phenomena, whereas it is important to state that each HDS still provides performance improvements when compared to its non-hybrid version in an SNR region restricted by a critical SNR value as mentioned in Numerical Results section. Beyond these critical SNR values, the benefits gathered by the transmit diversity of HDSs vanish in the presence of practical imperfections while only receive diversity persists as the ADO analysis has verified. Also, the performance comparisons that have been made between the first two HDSs (i.e., TAS/OSTBC and TAS/MRT schemes) and the latter two HDSs (i.e., MRT&RAS and JTRAS schemes) separately have provided insights about the robustness of two

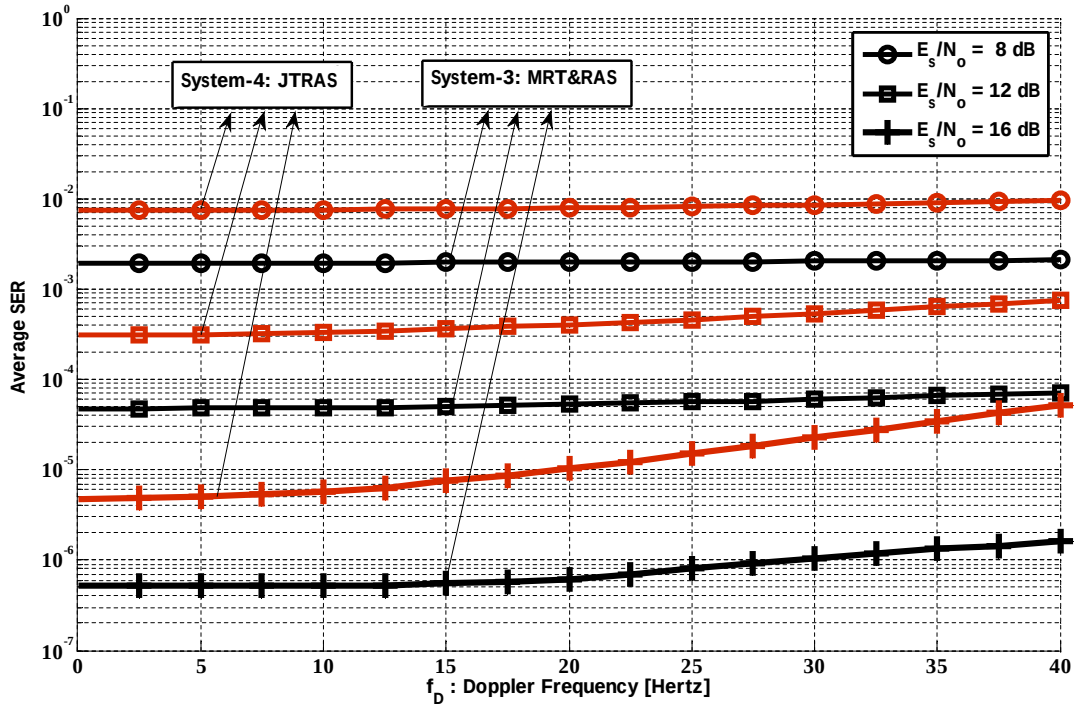


Figure 3.12: Variation of the SER performances of System-3: MRT&RAS and System-4: JTRAS depending on the Doppler frequency (f_D), for QPSK Signals and the parameters $m = 1.5$, $n_T = 2$, $n_R = 2$, $N_d = 4$, $N_{fr} = 20$.

groups of HDSs to the variation of the number of pilot symbols (N_d) and the maximum Doppler frequency shift (f_D), respectively. Consequently, the extensive examination of these four HDSs has provided a framework on the inspection of practical performance, superiority, robustness and importance in physical deployment scenarios of MIMO diversity schemes in real-world wireless communications.

4. PERFORMANCE ANALYSIS IN THE PRESENCE OF FE

This chapter focuses on the performance analyses of some HDSs, which are introduced in the previous chapter of the thesis, in the presence of FEs. Within the performance investigations, exact and approximate error probability expressions, exact OP expressions and Monte Carlo-based computer simulations are provided which are the most useful performance metrics. Since all the analyses are performed for the generic Nakagami- m fading model, the numerical results have exposed the performance examination in various channel conditions. Also, the derivations are kept generic for the modulation level such that the performance of many frequently-used digital modulation schemes can be obtained by simple substitutions.

This chapter mainly consists of three sections that respectively focus on the practical-performance examinations of the hybrid TAS/OSTBC and JTRAS/OSTBC schemes, a modified TAS/Alamouti-OSTBC, and a generalized modified TAS/OSTBC (M-TAS/OSTBC) scheme. However, all sections seem to be dealing with the hybrid TAS/OSTBC scheme, the feedback channel models used in the first section (Section 4.2) and the following two sections (Sections 4.3 and 4.4) are different from each other. The feedback channel is assumed to be a BSC which is modelled by the transition probabilities as defined in Section 2.4.3 in detail, and it enables researchers to perform a simple, quick but coarse examination of FEs. The model used in Sections 4.3 and 4.4 provides a more realistic viewpoint than that of Section 4.2 by considering a feedback link with fading and AWGN characteristics reciprocally to the main transmission link. Section 4.4 presents a framework on modified TAS/OSTBC schemes that are robust to FEs and require reduced amount of feedback burden when compared to conventional TAS/OSTBC (C-TAS/OSTBC) schemes. The framework in Section 4.4 also covers the work presented in Section 4.3 that has focused on the practical performance investigation of M-TAS/Alamouti-OSTBC scheme as the first results of performance analysis of M-TAS/OSTBC schemes under severe feedback conditions.

Before giving the analytical derivations of the performance metrics of the investigated HDSs, the current literature related to the performance investigations of diversity schemes under the effects of FEs is introduced.

4.1. Literature Survey About the FE

Researchers have been dealing with the examination of diversity schemes under the effect of CEEs, FD and FEs, in order to expose the feasibility of the diversity schemes for the real-world communications systems. Particularly, extensive and insightful work have been included in the literature on the effect of FE, since it provides a useful and efficient perspective for the assessment of diversity schemes in the presence of feedback impairments. MISO systems such as single TAS, TAS/G₂-OSTBC, single TAS/power allocation (PA) and TAS/PA/G₂-OSTBC have been analyzed considering FEs in [45] for Rayleigh fading channels and BPSK signals. The average BER expressions of TAS/G₂-OSTBC scheme and TAS/PA/G₂-OSTBC scheme with only four transmit antennas are derived for the selection of non-overlapping transmit antenna subsets which cannot guarantee the selection of the subset containing the best transmit antennas. The analysis provided in [45] deals with systems employing n_T transmit antennas where $n_T = 2^k$, $k \in \mathbb{Z}^+$ that provides simplicity in mapping/de-mapping of feedback bits and the evaluation of a priori probabilities of each transmit antenna subset combination (TASC). In [49], authors have examined the downlink opportunistic scheduling scheme by considering an erroneous feedback link case with arbitrary (not necessarily the power of 2) total numbers of feedback quantization levels. Also, the error performance analysis of JTRAS/OSTBC scheme in the presence of FEs has been investigated in [76] for QPSK signals and Rayleigh fading case. The performances of closed-loop diversity schemes such as single TAS and TB have been considered in [77] in the presence of FEs for Rayleigh fading channels by using upper bounds for BER of BPSK signals without providing any closed-form and exact BER expressions. The analyses provided in [45], [49], [76] and [77] have considered that the feedback channel has BSC characteristics.

As similar to the previous chapter about CEE and FD, the main motivation of the performance investigations provided in the following sections of this chapter is to focus on the practical performances of some HDSs under the effects of FEs that have not been considered in the literature in an extensive and insightful examination. Therefore, this chapter aims to make contributions to the literature for the sake of exposing the real-world characteristics and the feasibility of HDSs to the wireless communications standards and protocols.

Firstly, in Section 4.2, the examination of the hybrid TAS/OSTBC and JTRAS/OSTBC schemes is given by providing exact and approximate expressions for some performance metrics such as average BERs/SERs of binary and M -ary signals, average OP that help interpreting practical performance of WCSs. Secondly, by proposing a feedback-rate efficient and FE-resistant TAS/Alamouti-OSTBC scheme, Section 4.3 focuses on the practical performance of the proposed scheme in the presence of FEs in a feedback link that is assumed to have both fading and AWGN characteristics. Finally, in Section 4.4, a unified framework is presented by proposing FE-resistant TAS/OSTBC (M-TAS/OSTBC) schemes and by investigating the effect of FEs on M-TAS/OSTBC schemes. Note that, although [78] has proposed three different TAS/Alamouti-OSTBC schemes with reduced feedback rate (which are used for comparison in Section 4.3), the analysis over Rayleigh fading channels has been limited to the asymptotic error performance under ideal feedback channels that lacks the exact performance analysis under practical feedback channels facing FEs.

4.2. Analysis of TAS/OSTBC with Receive Selection and Combining in the Presence of FEs

The related literature does not include an extensive and exact performance analysis of JTRAS/OSTBC scheme for several modulations, arbitrary system configurations and a generic fading model. JTRAS/OSTBC scheme, which consists of combined TAS/OSTBC at transmit end and single RAS at receive end, has been considered in [13], [31]-[33], [79]. In [13], JTRAS with space-time coding has been analyzed asymptotically for Rayleigh fading channels by using upper bounds and PEP-based approach. The error performance of a special case, that is based on performing the OSTBC transmission by selecting two transmit antennas (i.e., Alamouti Code [7]), has been examined in [31] for only BPSK modulation and Rayleigh fading channels. Although authors of [32] have considered the same problem with arbitrary number of selected antennas, they have given results for QPSK modulations and Rayleigh fading channels by using numerical integrations and not provided any closed-form expressions for the performance metrics. The OP performance of JTRAS/OSTBC scheme has been examined for perfect feedback conditions in Nakagami- m fading channels in [33]. The performance of JTRAS/OSTBC scheme

in the presence of CEEs has been analyzed in [79] for Rayleigh fading channels and some specific transmit and receive antenna configurations using approximations and PEP-based analysis method.

The literature does not include the exact error performance analysis of JTRAS/OSTBC scheme in i.i.d. Nakagami- m fading channels for binary/ M -ary signals and even for ideal feedback cases. Besides, no exact and closed-form expressions are provided for the feedback channel impairments of TAS/OSTBC scheme (with MRC-like receiver) in Rayleigh fading channels. This part of the thesis, focusing on the unified analysis of hybrid JTRAS/OSTBC scheme and TAS/OSTBC scheme in the presence of FEs in i.i.d. and slow and frequency-flat Nakagami- m fading channels, derives useful statistics of the output SNR such as the PDF and the CDF. It examines the system capacity from the OP perspective and the MGF of the output SNR and exact error probability expressions of different modulations such as BPSK, QPSK, DBPSK, CBFSK, NCBFSK, M -PAM, M -QAM signals and the approximate error probability expressions of M -PSK signals (for $M \geq 8$) are derived for the erroneous feedback case. In order to focus on the asymptotic performance of the investigated HDSs, the ADO analysis is used. Also, Monte Carlo simulations are performed in order to validate the analytical performance results in the presence of the FEs.

The detailed analysis of the erroneous feedback cases for JTRAS/OSTBC scheme is provided for integer values of the fading parameters (i.e., $m \in \mathbb{Z}^+$). Whereas, the analysis for TAS/OSTBC scheme is provided for arbitrary values of $m \geq 1/2$ that satisfy the product of the fading parameter m and the number of receive antennas (n_R) to be integer (i.e., $mn_R \in \mathbb{Z}^+$).

The rest of the section is organized as follows. The detailed system configurations of the investigated HDSs are presented in Section 4.2.1 by focusing on the statistics of the output SNR such as the PDF, CDF and OP expressions. Section 4.2.2 provides the derivation of the unified performance expression covering the MGF of the output SNR and exact (and approximate for M -PSK, $M \geq 8$) BER/SER performance expressions. The asymptotic performance analyses of both schemes are provided in Section 4.2.3. Section 4.2.4 presents the numerical results consisting of the theoretical ones and Monte Carlo simulations for OP and average error rates of M -ary signals. Finally, Section 4.2.5 draws conclusions about the theoretical and simulation results.

4.2.1. System Model and SNR Statistics for JTRAS/OSTBC and TAS/OSTBC Schemes

This section focuses on two different HDSs with identical transmitter structure equipped with n_T antennas but different receiver structures which employ n_R antennas as depicted in Figure 4.1. The transmitter side employs orthogonal OSTBC after determining the best available transmit antenna subset with n_S ones out of $K = \binom{n_T}{n_S}$ combinations. At every OSTBC codeword transmission, the transmit antenna subset that maximizes the post-processing SNR at the receiver is activated while the other antennas are kept silent which results in reduced number of RF chains (i.e., reduced number of power amplifiers and complexity) at the transmitter. Whereas, the receiver side employs single RAS and MRC-like structure for JTRAS/OSTBC and TAS/OSTBC cases, respectively. The envelopes of the channel gains $h_{i,j}$, $i =$

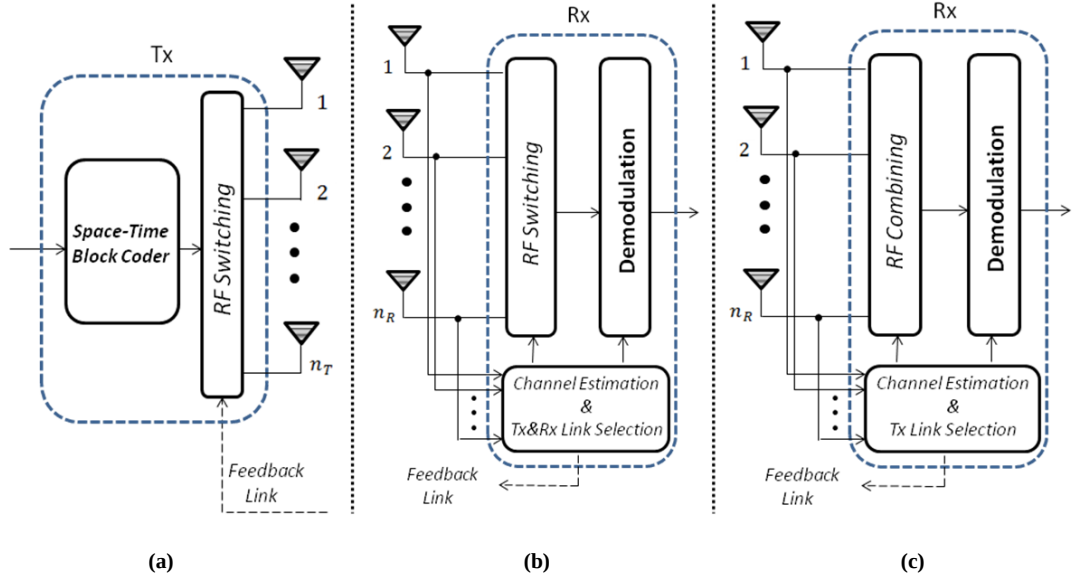


Figure 4.1: Block diagram of both schemes: (a) Transmitter of both schemes, (b) Receiver of JTRAS/OSTBC scheme, (c) Receiver of TAS/OSTBC scheme.

$1, 2, \dots, n_R$, $j = 1, 2, \dots, n_T$ defined between the transmit and receive ends, are modeled as i.i.d. Nakagami random variables with fading parameter m and squared mean $\Omega = E[|h_{i,j}|^2]$.

Before signal transmission, channel estimation and the selection of the transmit antenna subset for both schemes (and also the receive antenna for JTRAS/OSTBC case) are performed by using pilot symbols. For both schemes, by monitoring the

received signals, the optimal transmit antenna subset maximizing the post-processing SNR is determined. Besides, for JTRAS/OSTBC scheme, the receive antenna through which the overall received SNR would be the largest is activated. Whereas TAS/OSTBC scheme employs MRC-like combining through all receive branches after the optimal transmit antenna subset selection. Under the assumptions that the channel estimation and antenna selection processes are perfectly done, the antenna indices of the transmit and receive ends are determined and the index of the transmit antenna subset combination (TASC) that consists of the best transmit antennas (unlike [45] in which the optimal transmit antenna subset is not guaranteed) is sent back to the transmitter. Here, the feedback link is a low-rate link which can be modeled as BSC as considered in [45]. The low-rate structure of the feedback link can be maintained by sending only the index of the TASC to the transmitter as the feedback information. This can be provided by defining a look-up table-like codebook (at both transmit and receive ends) consisting of the indices of all TASCs. The deteriorating structure of the BSC causes some errors (i.e., FEs) at the delivery of the TASC index information (i.e., partial CSI) that results in degradation on the system performance [45]. In order to examine the effect of FEs on the performances of JTRAS/OSTBC and TAS/OSTBC schemes, the analysis is carried out based on the BSC assumptions and the total error probability expression given as

$$P_s(p_e) = p_{CF}P_{s,CF} + p_{EF}P_{s,EF}(p_e) \quad (4.1)$$

In (4.1), p_e denotes the average BER defined for each bit transmission over BSC. The feedback message should deliver $\eta = \lceil \log_2(K) \rceil$ bits in order to represent K different TASCs where the operator $\lceil \cdot \rceil$ denotes the smallest integer that is greater than or equal to the argument. But, since η bits can represent $L = 2^\eta \geq K$ different states, FE may cause the proper feedback messages denoted by $x_1, x_2, \dots, x_K$ that correspond to the proper TASC combinations to be de-mapped to one of the improper feedback messages $x_{K+1}, x_{K+2}, \dots, x_L$ (which is called as feedback failure (FF) in this section). Although many re-transmission or error control coding protocols (such as automatic repeat request (ARQ)) can be employed to overcome the FF, it is assumed that the transmitter arbitrarily selects and activates one of the proper TASCs in order not to reduce the

transmission rate that would be brought along with the usage of a re-signaling structure. Thus, assuming that all proper TASCs are selected equally-likely in the presence of FF as in [49] and [76], the a priori probabilities for the correct feedback (CF) and erroneous feedback (EF) cases are defined as

$$p_{CF} = (1 - p_e)^\eta + \frac{1}{K^2} \sum_{i=1}^K \sum_{j=0}^{L-K-1} p_e^{d_H(i,j)} (1 - p_e)^{\eta - d_H(i,j)} \quad (4.2)$$

and $p_{EF} = 1 - p_{CF}$, respectively concerning also the case of $L > K$ unlike the references [45] and [77] which consider $K = 2^b$, $b \in \mathbb{Z}^+$. In (4.2), $d_H(i, j) = d(x_i, x_{L-j})$ denotes the Hamming distance between the bit representations of the feedback messages x_i and x_{L-j} . In (4.1), $P_{s,CF}$ is the average SER expression related to the only correct TASC and $P_{s,EF}(p_e) = \sum_{\mathbf{c}_k, k=2, \dots, K} p_{\mathbf{c}_k} P_{s,EF}^{(\mathbf{c}_k)}$ is the average SER expression related to the EF case that is defined as the weighted sum of SERs $\left(P_{s,EF}^{(\mathbf{c}_k)}\right)$ related to $K - 1$ wrong TASCs (denoted as $\mathbf{c}_k$, $k = 2, \dots, K$) with equal probability of occurrence $p_{\mathbf{c}_k} = \frac{1}{K-1} + \frac{L-K}{LK}$. Here, the factor $\frac{L-K}{LK}$ appearing in $p_{\mathbf{c}_k}$, that is added to the equally-likely probability of occurrence $\frac{1}{K-1}$ for each wrong TASC combination, is the contribution of the equally-likely random selection probability of each proper transmit antenna subset in the presence of FF. Here, $\mathbf{c}_k = (n_1, n_2, \dots, n_{n_S})$, $n_{j+1} > n_j$, $j = 1, 2, \dots, n_S - 1$, denotes the vector of the selected transmit antenna indices and each vector is assigned to a feedback symbol x_k such that the single CF case is denoted by the first TASC ($\mathbf{c}_1 = (1, 2, \dots, n_S)$).

After the channel estimation, antenna selection and feedback processes, the OSTBC transmission results in an instantaneous SNR $\gamma_{i,j} = \frac{E_s}{n_S N_0 R_s} |h_{i,j}|^2$, $i = 1, 2, \dots, n_R$, $j = 1, 2, \dots, n_T$ and an average SNR $\bar{\gamma} = \frac{E_s \Omega}{n_S N_0 R_s}$ for each diversity branch. The SNR statistics related to the instantaneous SNR of each branch for JTRAS/OSTBC scheme have been obtained in [33] (also in [21], [29] and [70]) as $f(x) = \left(\frac{m}{\bar{\gamma}}\right)^m \frac{x^{m-1} e^{-xm/\bar{\gamma}}}{\Gamma(m)}$, $x \geq 0$, and $F(x) = \psi(m, xm/\bar{\gamma})/\Gamma(m)$, $x \geq 0$, respectively. For TAS/OSTBC scheme, the PDF and CDF expressions of the transmit branch (through all receive antennas) have been given similarly in [33] as $f(x) = \left(\frac{m}{\bar{\gamma}}\right)^{mn_R} \frac{x^{mn_R-1} e^{-xm/\bar{\gamma}}}{\Gamma(mn_R)}$, $x \geq 0$, and $F(x) = \psi(mn_R, xm/\bar{\gamma})/\Gamma(mn_R)$, $x \geq 0$, respectively. The SNR statistics given above for both HDSs can be unified by the following expressions:

$$f_U(x) = \left(\frac{m}{\bar{\gamma}}\right)^{mg} \frac{x^{mg-1} e^{-xm/\bar{\gamma}}}{\Gamma(mg)}, x \geq 0 \quad (4.3)$$

and

$$F_U(x) = \frac{\psi(mg, xm/\bar{\gamma})}{\Gamma(mg)}, x \geq 0 \quad (4.4)$$

which can also be written as

$$F_U(x) = 1 - e^{-\frac{xm}{\bar{\gamma}}} \sum_{k=0}^{mg-1} \frac{(xm/\bar{\gamma})^k}{k!}, x \geq 0 \quad (4.5)$$

while the product mg is integer [33, (1)] and [29, (1)]. Here the parameter is $g = 1$ for JTRAS/OSTBC and $g = n_R$ for TAS/OSTBC. The instantaneous output SNR for JTRAS/OSTBC scheme will be the maximum of the receive branch SNRs: $\gamma_{JTRAS/OSTBC} = \max_{1 \leq i \leq n_R} \{\gamma_{JTRAS/OSTBC}^{(i)}\}$. Thus, the CDF expression of the output SNR will be $F_{\gamma_{JTRAS/OSTBC}}(x) = \prod_{i=1}^{n_R} F_{\gamma_{JTRAS/OSTBC}^{(i)}}(x)$. Here, $F_{\gamma_{JTRAS/OSTBC}^{(i)}}(x)$ is the CDF expression related to the instantaneous branch SNR $\gamma_{JTRAS/OSTBC}^{(i)} = \sum_{k=1}^{n_S} Z_{n_k}^{(i)}$ where $Z_j^{(i)}$, $j = 1, 2, \dots, n_T$, denotes the sorted version of the branch SNRs such that the inequality $Z_1^{(i)} \geq Z_2^{(i)} \geq \dots \geq Z_{n_T}^{(i)}$ holds true. Also, for TAS/OSTBC scheme, the output SNR will be $\gamma_{TAS/OSTBC} = \sum_{k=1}^{n_S} Y_{n_k}$ where Y_j , $j = 1, 2, \dots, n_T$, denotes the sorted version of the branch SNRs (after the MRC-like combining structure of OSTBC) that satisfy the inequality $Y_1 \geq Y_2 \geq \dots \geq Y_{n_T}$. Thus, the output SNR of both JTRAS/OSTBC and TAS/OSTBC schemes can be unified by the SNR expression given as

$$\gamma_U = \max_{1 \leq v \leq N} \{\gamma^{(v)}\} \quad (4.6)$$

where the number of branches related to the receive selection is $N = n_R$ for JTRAS/OSTBC and $N = 1$ (no selection) for TAS/OSTBC.

The PDF of the branch SNR $\gamma^{(v)}$ given in (4.6) can be obtained by using the joint PDF given in [65, (2.2.3)] as

$$f_{(n_1), (n_2), \dots, (n_{n_S})}^{(v)}(x_1, x_2, \dots, x_{n_S}) = n_T! \left\{ \prod_{k=1}^{n_S} f_U(x_k) \right\} \times \prod_{k=1}^{n_S+1} \left\{ \frac{\{F_U(x_{k-1}) - F_U(x_k)\}^{n_k - n_{k-1} - 1}}{(n_k - n_{k-1} - 1)!} \right\} \quad (4.7)$$

where $x_k \geq x_{k+1}$, $x_0 = +\infty$, $x_{n_S+1} = 0$, $n_0 = 0$ and $n_{n_S+1} = n_T + 1$. By substituting binomial expansion of the difference of CDF expressions given in terms of $F_U(x)$ as

$$\{F_U(x_{k-1}) - F_U(x_k)\}^{n_k - n_{k-1} - 1} = \sum_{p_k^{(v)}=0}^{n_k - n_{k-1} - 1} c_{p,k}^{(v)} \times (F_U(x_{k-1}))^{p_k^{(v)}} (F_U(x_k))^{n_k - n_{k-1} - 1 - p_k^{(v)}} \quad (4.8)$$

where $c_{p,k}^{(v)} = (-1)^{n_k - n_{k-1} - 1 - p_k^{(v)}} \binom{n_k - n_{k-1} - 1}{p_k^{(v)}}$ and also by substituting (4.3) and the binomial expanded expression of $[F_U(x_k)]^{z_k^{(v)}}$

$$[F_U(x_k)]^{z_k^{(v)}} = \left(1 - e^{-mx_k/\bar{\gamma}} \sum_{r=0}^{mg-1} \frac{(mx_k/\bar{\gamma})^r}{r!} \right)^{z_k^{(v)}} = \sum_{t_k^{(v)}=0}^{z_k^{(v)}} \sum_{r_k^{(v)}=0}^{t_k^{(v)}(mg-1)} c_{tr,k}^{(v)} x_k^{r_k^{(v)}} e^{-t_k^{(v)} x_k m / \bar{\gamma}} \quad (4.9)$$

into (4.8), the joint PDF in (4.7) can be expressed as

$$f_{(n_1), (n_2), \dots, (n_{n_S})}^{(v)}(x_1, x_2, \dots, x_{n_S}) = \frac{c_0 (m/\bar{\gamma})^{mg n_S}}{\Gamma(mg)^{n_S}} \sum_{P^{(v)}} \sum_{T^{(v)}} \sum_{R^{(v)}} \times \prod_{k=1}^{n_S} \left\{ c_{p,k}^{(v)} c_{tr,k}^{(v)} x_k^{mg+r_k^{(v)}-1} e^{-x_k m (1+t_k^{(v)})/\bar{\gamma}} \right\} \quad (4.10)$$

In (4.9) and (4.10), the parameters are defined as $c_0 = n_T! / \prod_{k=1}^{n_S+1} \{(n_k - n_{k-1} - 1)!\}$, $d_k^{(v)} = n_k - n_{k-1} - 1 - p_k^{(v)} + p_{k+1}^{(v)}$ (note that $p_{n_S+1}^{(v)} = 0$) and $c_{tr,k}^{(v)} = \binom{d_k^{(v)}}{t_k^{(v)}} (-1)^{t_k^{(v)}} \beta_{r_k^{(v)}}(t_k^{(v)}, mg)$ where $\beta_{r_k^{(v)}}(t_k^{(v)}, mg)$ denotes the multinomial coefficients [62, (0.314)], [29]. The summations defined with the indices $P^{(v)}$, $T^{(v)}$ and $R^{(v)}$ in (4.10) all denote n_S -fold summations with the indices $p_k^{(v)}$, $t_k^{(v)}$ and $r_k^{(v)}$, respectively. Since the multivariate structure of the PDF expression

in (4.10) is analytically challenging, the Laplace transform and the inverse Laplace transform pair is applied in order to provide a more simplified representation. The Laplace transform of (4.10) can be obtained by substituting it into the multivariate Laplace integral as

$$H^{(v)}(s) = E[e^{-s(x_1+x_2+\dots+x_{n_S})}] = \frac{c_0(m/\bar{\gamma})^{mgn_S}}{\Gamma(mg)^{n_S}} \sum_{P^{(v)}} \sum_{T^{(v)}} \sum_{R^{(v)}} \prod_{k=1}^{n_S} \{c_{p,k}^{(v)} c_{tr,k}^{(v)}\} \quad (4.11)$$

$$\times \int_{x_{n_S}=0}^{\infty} x_{n_S}^{\mu_{n_S}^{(v)}-1} e^{-x_{n_S} b_{n_S}^{(v)}} \dots \int_{x_1=x_2}^{\infty} x_1^{\mu_1^{(v)}-1} e^{x_1 b_1^{(v)}} dx_1 dx_2 \dots dx_{n_S}$$

where $\mu_k^{(v)} = mg + r_k^{(v)}$ and $b_k^{(v)} = \left(s + \frac{m(1+t_k^{(v)})}{\bar{\gamma}}\right)$. It is possible to derive the Laplacian expression by manipulating [80, (13)] as

$$H^{(v)}(s) = \frac{c_0(m/\bar{\gamma})^{mgn_S}}{\Gamma(mg)^{n_S}} \sum_{P^{(v)}} \sum_{T^{(v)}} \sum_{R^{(v)}} \prod_{k=1}^{n_S} \{c_{p,k}^{(v)} c_{tr,k}^{(v)}\} \sum_{l_1^{(v)}=0}^{\mu_1^{(v)}-1} \frac{\Gamma(\mu_1^{(v)}) n_S^{u_1^{(v)}}}{l_1^{(v)}! (s + a_1^{(v)})^{u_1^{(v)}}}$$

$$\sum_{l_2^{(v)}=0}^{l_1^{(v)}+\mu_2^{(v)}-1} \frac{\Gamma(l_1^{(v)} + \mu_2^{(v)}) \left(\frac{n_S}{2}\right)^{u_2^{(v)}}}{l_2^{(v)}! (s + a_2^{(v)})^{u_2^{(v)}}} \dots \quad (4.12)$$

$$\dots \sum_{l_{n_S-1}^{(v)}=0}^{l_{n_S-2}^{(v)}+\mu_{n_S-1}^{(v)}-1} \frac{\Gamma(l_{n_S-2}^{(v)} + \mu_{n_S-1}^{(v)}) \left(\frac{n_S}{n_S-1}\right)^{u_{n_S-1}^{(v)}}}{l_{n_S-1}^{(v)}! (s + a_{n_S-1}^{(v)})^{u_{n_S-1}^{(v)}}}$$

$$\times \Gamma(l_{n_S-1}^{(v)} + \mu_{n_S}^{(v)}) (s + a_{n_S}^{(v)})^{-u_{n_S}^{(v)}}$$

which can be re-written in a more compact form as [84], [85]

$$H^{(v)}(s) = \frac{c_0(m/\bar{\gamma})^{mgn_S}}{\Gamma(mg)^{n_S}} \sum_{P^{(v)}} \sum_{T^{(v)}} \sum_{R^{(v)}} \sum_{L^{(v)}} \prod_{k=1}^{n_S} \{c_{p,k}^{(v)} c_{tr,k}^{(v)} c_{l,k}^{(v)}\} \quad (4.13)$$

$$\times \prod_{k=1}^{n_S} \left\{ \left(s + a_k^{(v)}\right)^{-u_k^{(v)}} \right\}$$

where $a_k^{(v)} = \frac{mn_S}{k\bar{\gamma}} \sum_{j=1}^k \left(1 + t_j^{(v)}\right)$ and the parameters $c_{l,k}^{(v)}$ and $u_k^{(v)}$ are defined as:

$$c_{l,k}^{(v)} = \begin{cases} \frac{\Gamma(\mu_1^{(v)})}{l_1^{(v)}!} n_S^{u_1^{(v)}}, & k = 1 \\ \frac{\Gamma(l_{k-1}^{(v)} + \mu_k^{(v)})}{l_k^{(v)}!} n_S^{u_k^{(v)}}, & k = 2, 3, \dots, n_S - 1 \\ \Gamma(l_{n_S-1}^{(v)} + \mu_{n_S}^{(v)}), & k = n_S \end{cases}$$

$$u_k^{(v)} = \begin{cases} \mu_1^{(v)} - l_1^{(v)}, & k = 1 \\ \mu_k^{(v)} - l_k^{(v)} + l_{k-1}^{(v)}, & k = 2, 3, \dots, n_S - 1 \\ \mu_{n_S}^{(v)} - l_{n_S-1}^{(v)}, & k = n_S \end{cases}$$

In (4.13), the summation defined with the index $L^{(v)}$ denotes $(n_S - 1)$ -fold summations with the indices $l_k^{(v)}$. Afterwards, performing the inverse Laplace transform over (4.13) will result in the univariate representation of the joint PDF in (4.10) as

$$f^{(v)}(x) = \mathcal{K}_0^{(v)} \sum_{P^{(v)}} \sum_{T^{(v)}} \sum_{R^{(v)}} \sum_{L^{(v)}} \mathcal{K}_1^{(v)} \mathcal{L}^{-1} \left\{ \prod_{k=1}^{n_S} \left(s + a_k^{(v)} \right)^{-u_k^{(v)}} \right\} \quad (4.14)$$

where $\mathcal{K}_0^{(v)} = \frac{c_0(m/\bar{\gamma})^{mg n_S}}{\Gamma(mg)^{n_S}}$, $\mathcal{K}_1^{(v)} = \prod_{k=1}^{n_S} \{c_{p,k}^{(v)} c_{tr,k}^{(v)} c_{l,k}^{(v)}\}$. The inverse Laplace transform can be obtained by using [72, (2.1.4-8)] that is valid for the condition $a_i^{(v)} \neq a_j^{(v)}, \forall i \neq j$. In order to utilize this identity and obtain the PDF expression, the coefficients $a_k^{(v)}, k = 1, 2, \dots, n_S$, are renamed as $\tilde{a}_d^{(v)}, d = 1, 2, \dots, D^{(v)} \leq n_S$, such that all the coefficients $(\tilde{a}_d^{(v)})$ are different from each other. Hence, the coefficients with equal values will result in the coefficient with the new power $(\tilde{u}_d^{(v)})$ that is the sum of the powers of the equal-valued coefficients (i.e., $\tilde{u}_1^{(v)} = u_1^{(v)} + u_2^{(v)}$ if $a_1^{(v)} = a_2^{(v)}$). Thus, by considering this rearrangement, the PDF in (4.14) can be obtained as

$$f^{(v)}(x) = \mathcal{K}_0^{(v)} \sum_{P^{(v)}} \sum_{T^{(v)}} \sum_{R^{(v)}} \sum_{L^{(v)}} \sum_{k^{(v)}=1}^{D^{(v)}} \sum_{q^{(v)}=1}^{\tilde{u}_k^{(v)}} \frac{\mathcal{K}_1^{(v)} A_{kq} \left(-\tilde{a}_k^{(v)} \right)}{\left(\tilde{u}_k^{(v)} - q^{(v)} \right)!} \frac{x^{\tilde{u}_k^{(v)} - q^{(v)}} e^{-x \tilde{a}_k^{(v)}}}{(q^{(v)} - 1)!} \quad (4.15)$$

$A_{kq}(\cdot)$, the residue coefficients in (4.15), have been defined in [72, (2.1.4-8)] in detail. Integrating the PDF in (4.15), the CDF of the v^{th} diversity branch can be derived as

$$F^{(v)}(x) = \mathcal{K}_0^{(v)} \sum_{P^{(v)}} \sum_{T^{(v)}} \sum_{R^{(v)}} \sum_{L^{(v)}} \mathcal{K}_1^{(v)} \sum_{k^{(v)}=1}^{D^{(v)}} \sum_{q^{(v)}=1}^{\tilde{u}_k^{(v)}} \mathcal{K}_2^{(v)} \psi \left(\tilde{u}_k^{(v)} - q^{(v)} + 1, x \tilde{a}_k^{(v)} \right) \quad (4.16)$$

where $\mathcal{K}_2^{(v)} = \frac{A_{kq}(-\tilde{a}_k^{(v)})}{(\tilde{u}_k^{(v)} - q^{(v)})!(q^{(v)} - 1)!} \left(\tilde{a}_k^{(v)}\right)^{q^{(v)} - \tilde{u}_k^{(v)} - 1}$. Also, by using the identity given in [62, (8.351.2)] as $\psi(c_1, c_2 t) = \frac{(c_2 t)^{c_1}}{c_1} e^{-c_2 t} {}_1F_1(1; 1 + c_1; c_2 t)$, the CDF in (4.16) can be re-organized by means of the confluent hypergeometric function ${}_1F_1$ [62, (210.1)] as

$$F^{(v)}(x) = \mathcal{K}_0^{(v)} \sum_{P^{(v)}} \sum_{T^{(v)}} \sum_{R^{(v)}} \sum_{L^{(v)}} \mathcal{K}_1^{(v)} \sum_{k^{(v)}=1}^{D^{(v)}} \sum_{q^{(v)}=1}^{\tilde{u}_k^{(v)}} \mathcal{K}_3^{(v)} e^{-x\tilde{a}_k^{(v)}} x^{-q^{(v)} + \tilde{u}_k^{(v)} + 1} \times {}_1F_1\left(1; \tilde{u}_k^{(v)} - q^{(v)} + 2; x\tilde{a}_k^{(v)}\right) \quad (4.17)$$

where $\mathcal{K}_3^{(v)} = \frac{A_{kq}(-\tilde{a}_k^{(v)})}{(\tilde{u}_k^{(v)} - q^{(v)} + 1)!(q^{(v)} - 1)!}$.

After the derivation of the statistical properties of branch SNRs as shown above, the distribution of the output SNR can be easily derived. By using (4.6) and the highest order statistics [64], the CDF of the output SNR would be the product of the identical marginal CDF expressions $F^{(v)}(x)$ given in (4.16):

$$F_{\gamma_U}(x) = \mathcal{K}_0 \sum_P \sum_T \sum_R \sum_L \mathcal{K}_1 \sum_K \sum_Q \mathcal{K}_2 \prod_{v=1}^N \psi\left(\tilde{u}_k^{(v)} - q^{(v)} + 1, x\tilde{a}_k^{(v)}\right) \quad (4.18)$$

where $\mathcal{K}_0 = \prod_{v=1}^N \mathcal{K}_0^{(v)}$, $\mathcal{K}_1 = \prod_{v=1}^N \mathcal{K}_1^{(v)}$ and $\mathcal{K}_2 = \prod_{v=1}^N \mathcal{K}_2^{(v)}$. In (4.18), the summations defined with the indices by P , T , R , L , K and Q all denote N -fold summations with the indices $P^{(v)}$, $T^{(v)}$, $R^{(v)}$, $L^{(v)}$, $k^{(v)}$ and $q^{(v)}$, respectively. Similarly, by using (4.17), the CDF of the output SNR can also be re-written by means of the confluent hypergeometric function as [84], [85];

$$F_{\gamma_U}(x) = \mathcal{K}_0 \sum_P \sum_T \sum_R \sum_L \mathcal{K}_1 \sum_K \sum_Q \mathcal{K}_3 \frac{x^{N + \sum_{v=1}^N \tilde{u}_k^{(v)} - q^{(v)}}}{e^{x \sum_{v=1}^N \tilde{a}_k^{(v)}}} \times \prod_{v=1}^N {}_1F_1\left(1; \tilde{u}_k^{(v)} - q^{(v)} + 2; x\tilde{a}_k^{(v)}\right) \quad (4.19)$$

where $\mathcal{K}_3 = \prod_{v=1}^N \mathcal{K}_3^{(v)}$. By simply differentiating $F_{\gamma_U}(x) = \prod_{v=1}^N F^{(v)}(x)$ with respect to x , the PDF of the output SNR can also be obtained as $f_{\gamma_U}(x) = N f^{(v)}(x) \prod_{v=1}^{N-1} F^{(v)}(x)$ in terms of the same coefficients, incomplete Gamma function

and the confluent hypergeometric function. $F_{\gamma_U}(\cdot)$ constitutes the main metric required for the derivation of other performance metrics (e.g., the MGF and exact average error probability). Derivation of $F_{\gamma_U}(\cdot)$ also provides obtaining the OP of the unified scheme by easily evaluating $P_{out} = F_{\gamma_U}(2^{\mathcal{R}} - 1)$ as shown in [20], [28], [29], [33], [42], [70] and [73]. Here, $\mathcal{R}$ denotes the desired spectral efficiency in bit/s/Hertz. Note that, although (4.1) is given for BER/SERs of the investigated schemes, replacing OP expressions instead of the BER/SER expressions will result in the average OP of the diversity scheme for any system configurations.

4.2.2. Exact Error Performance Analysis for JTRAS/OSTBC and TAS/OSTBC Schemes

This subsection presents the unified performance expression that covers the MGF and BER/SER expressions of the investigated HDSs by using the SNR statistics derived in the previous section. For this purpose, the analysis is focused on the unified integral representation given in [73, (26)] as

$$\mathcal{J}_{\gamma_U}(\theta, \epsilon, \Phi) = \theta \int_0^\infty x^\epsilon e^{-x\Phi} F_{\gamma_U}(x) dx \quad (4.20)$$

Here, setting different values in the parameters of this function $(\theta, \epsilon; \Phi)$ will result in the MGF and the SER of different modulation signals as defined in [73]. In addition to this unified function, another unified function that is defined in (3.48) is used in order to cover the SER expressions of some M -ary modulations as well:

$$\hat{\mathcal{J}}_{\gamma_U}(\theta, \Phi) = \theta \int_0^\infty e^{-x\Phi} F_{\gamma_U}(x) {}_1F_1\left(1; \frac{3}{2}; \frac{x\Phi}{2}\right) dx \quad (4.21)$$

where ${}_1F_1$ denotes the confluent hypergeometric function [62, (210.1)]. By considering both (4.20) and (4.21), the MGF of the output SNR can be evaluated with the help of the identity $M_{\gamma_U}(s) = \mathcal{J}(s, 0, s)$. Also, the average error performance expressions of different modulation types have been defined in terms of (4.20) and (4.21) and listed below in Table 4.1. Thus, deriving the exact expressions of (4.20) and (4.21) will lead us to obtain and the performance metrics of the investigated diversity schemes for several modulations. By substituting (4.19) into (4.20) and (4.21), and re-writing the

resulting integrals by means of the Lauricella function described in [65, (2.4.2)], the unified functions in (4.20) and (4.21) can be respectively expressed as [85]

Table 4.1: Average BER/SER Expressions of binary and M -ary modulations in terms of (4.20) and (4.21).

Modulation	BER/SER Expression
BPSK	$P_{b,U} = \mathcal{J}_{\gamma_U} \left(\frac{1}{2\sqrt{\pi}}, -\frac{1}{2}, 1 \right)$
CBFSK	$P_{b,U} = \mathcal{J}_{\gamma_U} \left(\frac{1}{8\sqrt{\pi}}, -\frac{1}{2}, \frac{1}{2} \right)$
NCBFSK	$P_{b,U} = \mathcal{J}_{\gamma_U} \left(\frac{1}{4}, 0, \frac{1}{2} \right)$
DBPSK	$P_{b,U} = \mathcal{J}_{\gamma_U} \left(\frac{1}{2}, 0, 1 \right)$
M -PSK (Approx.)	$P_{s,U} = \mathcal{J}_{\gamma_U} \left(\frac{\sin(\pi/M)}{\sqrt{\pi}}, -\frac{1}{2}, \frac{\sin(\pi)}{M} \right)$
M -PAM	$P_{s,U} = \mathcal{J}_{\gamma_U} \left(\sqrt{\frac{3(M-1)}{\pi M^2(M+1)}}, -\frac{1}{2}, \frac{3}{(M^2-1)} \right)$
QPSK	$P_{s,U} = \mathcal{J}_{\gamma_U} \left(\sqrt{\frac{1}{8\pi}}, -\frac{1}{2}, \frac{1}{2} \right) + \hat{\mathcal{J}}_{\gamma_U} \left(\frac{1}{2\pi}, 1 \right)$
M -QAM	$P_{s,U} = \mathcal{J}_{\gamma_U} \left(\sqrt{\frac{6(\sqrt{M}-1)^2}{\pi M^2(M-1)}}, -\frac{1}{2}, \frac{3}{2(M-1)} \right) + \hat{\mathcal{J}}_{\gamma_U} \left(\frac{6(\sqrt{M}-1)}{\pi(\sqrt{M}+1)}, \frac{3}{M-1} \right)$

$$\begin{aligned} \mathcal{J}_{\gamma_U}(\theta, \epsilon, \Phi) &= \theta \mathcal{K}_0 \sum_P \sum_T \sum_R \sum_L \mathcal{K}_1 \sum_K \sum_Q \mathcal{K}_3 \frac{\Gamma(\mathcal{B}(\epsilon))}{\mathcal{C}_k^{\mathcal{B}(\epsilon)}} \\ &\times F_A^{(N)} \left(\mathcal{B}(\epsilon); 1, \dots, 1; b_1, \dots, b_N; \frac{\tilde{a}_k^{(1)}}{\mathcal{C}_k}, \dots, \frac{\tilde{a}_k^{(N)}}{\mathcal{C}_k} \right) \end{aligned} \quad (4.22)$$

and

$$\begin{aligned} \hat{\mathcal{J}}_{\gamma_U}(\theta, \Phi) &= \theta \mathcal{K}_0 \sum_P \sum_T \sum_R \sum_L \mathcal{K}_1 \sum_K \sum_Q \mathcal{K}_3 \frac{\Gamma(\mathcal{B}(0))}{\mathcal{C}_k^{\mathcal{B}(0)}} \\ &\times F_A^{(N+1)} \left(\mathcal{B}(0); 1, \dots, 1; \frac{3}{2}, b_1, \dots, b_N; \frac{\Phi}{2\mathcal{C}_k}, \frac{\tilde{a}_k^{(1)}}{\mathcal{C}_k}, \dots, \frac{\tilde{a}_k^{(N)}}{\mathcal{C}_k} \right) \end{aligned} \quad (4.23)$$

where $\mathcal{B}(\epsilon) = 1 + \epsilon + N + \sum_{v=1}^N \tilde{u}_k^{(v)} - q^{(v)}$, $\mathcal{C}_k = \Phi + \sum_{v=1}^N \tilde{a}_k^{(v)}$, $b_v = \tilde{u}_k^{(v)} - q^{(v)} + 2$. The summation indices given in (4.22) and (4.23) decomposes into multi-fold summation indices that are denoted as $\sum_S = \sum_{S^{(1)}} \sum_{S^{(2)}} \dots \sum_{S^{(N)}}$ and $\sum_{S^{(v)}} = \sum_{s_1^{(v)}} \sum_{s_2^{(v)}} \dots \sum_{s_N^{(v)}}$ for all values of $v = 1, 2, \dots, N$. Since the convergence criterion $\sum_{i=1}^n |x_i| < 1$ for the Lauricella function $F_A^{(n)}(a; b_1, \dots, b_n; c_1, \dots, c_n; x_1, \dots, x_n)$

used in (4.22) and (4.23) is satisfied for possible values of the function arguments, $F_A^{(n)}$ is computed via the simple integral definition in [65, (2.3.3)]. Also, [21] and [81] provide efficient methods for the numerical evaluation of this function by using the Gauss-Laguerre integration and semi-infinite Gauss-Hermite quadrature method, respectively. Note that setting the parameters given in (4.4)-(4.6) as $(g, N) = (1, n_R)$ will result in the exact expressions related to JTRAS/OSTBC scheme whereas setting $(g, N) = (n_R, 1)$ will result in those of TAS/OSTBC scheme. For instance, the exact expressions for the TAS/OSTBC scheme would be obtained as

$$\begin{aligned} \mathcal{J}_{\gamma_{TAS/OSTBC}}(\theta, \epsilon, \Phi) &= \theta \mathcal{K}_0 \sum_P \sum_T \sum_R \sum_L \mathcal{K}_1 \sum_k \sum_q \mathcal{K}_3 \\ &\times \frac{\Gamma(\mathcal{D}(\epsilon))}{\mathcal{E}_k^{\mathcal{D}(\epsilon)}} {}_2F_1\left(1; \mathcal{D}(\epsilon); \mathcal{D}(0); \frac{\tilde{a}_k}{\mathcal{E}_k}\right) \end{aligned} \quad (4.24)$$

and

$$\begin{aligned} \hat{\mathcal{J}}_{\gamma_{TAS/OSTBC}}(\theta, \Phi) &= \theta \mathcal{K}_0 \sum_P \sum_T \sum_R \sum_L \mathcal{K}_1 \sum_k \sum_q \mathcal{K}_3 \\ &\times \frac{\Gamma(\mathcal{D}(0))}{\mathcal{E}_k^{\mathcal{D}(0)}} F_2\left(\mathcal{D}(\epsilon); 1, 1, \frac{3}{2}, \mathcal{D}(0); \frac{\Phi}{2\mathcal{E}_k}, \frac{\tilde{a}_k}{\mathcal{E}_k}\right) \end{aligned} \quad (4.25)$$

where $\mathcal{D}(\epsilon) = 2 + \epsilon + N + \tilde{u}_k - q$, $\mathcal{E}_k = \Phi + \tilde{a}_k$, ${}_2F_1$ [62, (3.197-3)] and F_2 [65, (3.35.7-1)] denote the Gauss hypergeometric function and the Appell hypergeometric function (of the second kind), respectively. Both of these functions can be easily evaluated via many well known computation tools (e.g., MATHEMATICA and MAPLE).

4.2.3. Asymptotic Diversity Order Analysis of JTRAS/OSTBC and TAS/OSTBC Schemes under the Effect of FEs

Since exact theoretical performance expressions given in (4.22)-(4.25) do not provide any insights for interpreting the ADO, the asymptotic behavior of the unified diversity scheme is examined. Hence, by using the high-SNR approximation suggested in [69], the asymptotic analysis is carried out to derive approximate expressions for the SNR statistics that provide observing the ADO of the investigated schemes clearly. According to [69], expressing the PDF of the total channel power gain $H_U = \gamma_U/\bar{\gamma}$ as

$f_{H_U}(x) = ax^t + o(x^t)$, $a > 0$, paves the way to express the error probability expression approximately as

$$P_{e,U} = \frac{2^t a \Gamma(t + 1.5)}{\sqrt{\pi}(t + 1)} (k\bar{\gamma})^{-t-1} + o(\bar{\gamma}^{-t-1}) \quad (4.26)$$

for any modulation signals with a CEP $P(e|h_U)$ that is defined by means of $Q(\sqrt{h_U k \bar{\gamma}})$ where k is the modulation-dependent parameter. As explored from (4.26), the term $t + 1$ denotes the ADO of the overall system. Thus, deriving the value of t will be enough to examine the ADO. In order to obtain the asymptotic representation of the PDF/CDF of the total channel power gain (H_U), the PDF/CDF of each diversity branch can be derived asymptotically. Denoting the sorted versions of the channel power gains of the v^{th} branch as $W_1^{(v)} \geq W_2^{(v)} \geq \dots \geq W_{n_T}^{(v)}$ and the total channel power gain of this branch as $H_U^{(v)} = \sum_{k=1}^{n_S} w_{n_k}^{(v)}$, and using the union bounding technique, the CDF of $H_U^{(v)}$ can be upper bounded as given below:

$$\begin{aligned} F_{H_U^{(v)}}(x) &= Pr\left(H_U^{(v)} \leq x\right) \\ &= Pr\left(\sum_{k=1}^{n_S} w_{n_k}^{(v)} \leq x\right) \leq \sum_{k=1}^{n_S} Pr(w_{n_k}^{(v)} \leq x) = \sum_{k=1}^{n_S} F_{n_k}(x) \end{aligned} \quad [-12pt] \quad (4.27)$$

Here $F_{n_k}(x)$ denotes the CDF of the n_k^{th} highest order statistics. Since the upper bounded expression of the PDF of the n_k^{th} highest order statistics has been given in [82, (13)] as $f_{n_k}(x) \leq A_k x^{mg(n_T - n_k + 1) - 1} + o(x^{mg(n_T - n_k + 1) - 1})$ where $A_k = \binom{n_T}{n_T - n_k + 1} \frac{n_T - n_k + 1}{\Gamma(mg)\Gamma(mg + 1)^{n_T - n_k}}$, it is possible to derive the upper bounded representation of $F_{n_k}(x)$ by simply integrating this PDF expression

$$F_{n_k}(x) \leq B_k x^{mg(n_T - n_k + 1)} + o(x^{mg(n_T - n_k + 1)}) \quad (4.28)$$

where $B_k = \frac{A_k}{mg(n_T - n_k + 1)}$. Thus, $F_{H_U^{(v)}}(x)$ can be written as

$$F_{H_U^{(v)}}(x) \leq B_{n_{min}} x^{mg(n_T - n_{min} + 1)} + o(x^{mg(n_T - n_{min} + 1)}) \quad (4.29)$$

where $n_{min} = \min_{1 \leq k \leq n_S} n_k$. The right hand side of (4.29) has been written by considering the fact that the whole expression is almost dominated by the polynomial component with the greatest exponent. Using the approximate CDF expression obtained in (4.29), the CDF of the overall channel power gain can be derived by using the product definition $F_{H_U}(x) = \left(F_{H_U^{(v)}}(x)\right)^N \leq B_{n_{min}}^N x^{mgN(n_T - n_{min} + 1)} + o\left(x^{mgN(n_T - n_{min} + 1)}\right)$. Finally, derivative of $F_{H_U}(x)$ will result in the approximate PDF expression of the overall channel power gain:

$$f_{H_U}(x) \leq B_{n_{min}}^N mgN(n_T - n_{min} + 1)x^{mgN(n_T - n_{min} + 1) - 1} + o\left(x^{mgN(n_T - n_{min} + 1) - 1}\right) \quad (4.30)$$

Thus, the parameters are obtained as $a = B_{n_{min}}^N mgN(n_T - n_{min} + 1)$ and $t = mgN(n_T - n_{min} + 1) - 1$. By substituting these parameters into (4.26), the asymptotic BER/SER of both schemes can be easily obtained. Since, the product $gN = n_R$ for both JTRAS/OSTBC and TAS/OSTBC schemes, ADO of the unified scheme is derived as $mn_R(n_T - n_{min} + 1)$. Considering that n_{min} has integer values satisfying $1 \leq n_{min} \leq n_T - n_S + 1$, the effective diversity order (EDO) of the unified scheme which is defined in [83] as simply the ratio $-\log(P_s)/\log(E_s/N_0)$, is obtained to vary in the interval $[mn_R n_S, mn_R n_T]$, where $mn_R n_T$ is the maximum achievable ADO value of any diversity system equipped with n_T transmitter and n_R receiver antennas and operating in a Nakagami- m fading environment. The value of the EDO has been lower bounded by the product of the fading parameter m , the number of receive antennas n_R and the number of selected antennas (n_S) that are used for OSTBC transmission.

4.2.4. Numerical Results for JTRAS/OSTBC and TAS/OSTBC Schemes in the Presence of FEs

The effects of erroneous feedback on the performance of both JTRAS/OSTBC and TAS/OSTBC schemes in Nakagami- m fading channels have been presented via theoretical and simulation results of various system and channel conditions. All the numerical evaluations related to the performance metrics exposed in Figures 4.2-7 are provided for $\Omega = 1$, $E_s = 1$ and the average BERs specified for the feedback

link: $p_e \in 0.0001, 0.01, 0.05, 0.1, 0.2, 0.5, 0.7, 0.9$.

Since OP provides a

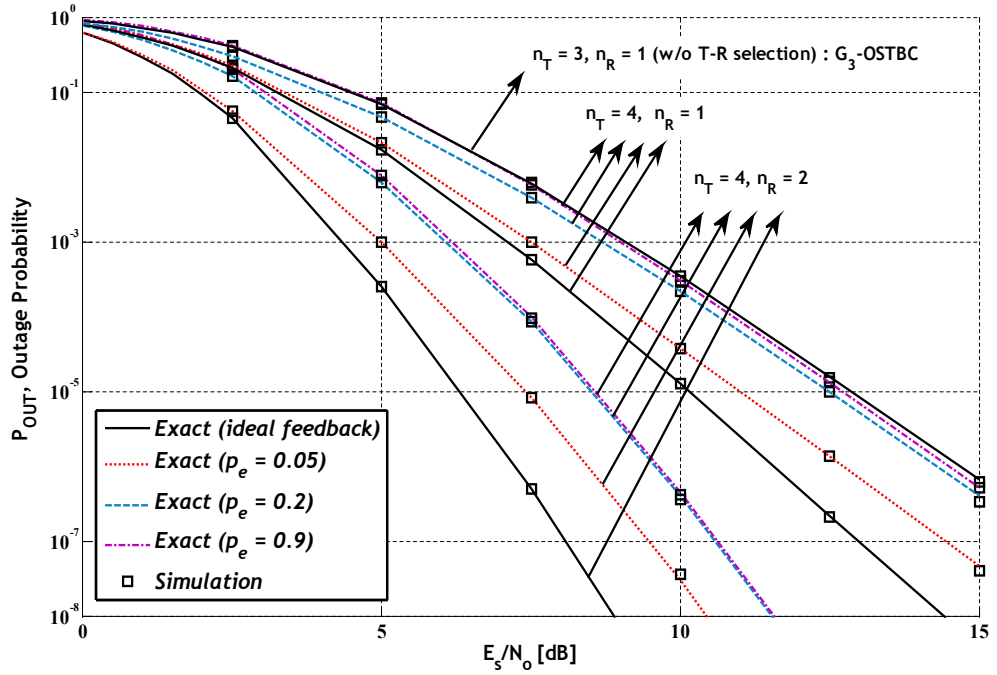


Figure 4.2: Outage probability vs. average SNR per symbol for JTRAS/ G_3 -OSTBC scheme for several n_T, n_R ($\mathcal{R} = 2$ bit/s/Hertz, $m = 2$, $p_e = 0.05$, $p_e = 0.2$ and $p_e = 0.9$).

useful and realistic insight for interpreting the system capacity, theoretical results are presented using (4.18) and simulation results of the OP of three-branch ($n_S = 3$) JTRAS/ G_3 -OSTBC scheme in Figure 4.2 for $n_R = 1$ and $n_R = 2$ in Nakagami- m fading environment ($m = 2$). For the average BER values of $p_e = 0.05$, $p_e = 0.20$ and $p_e = 0.90$ and the desired bandwidth efficiency of $R = 2$ bit/s/Hertz, this figure depicts the effect of the variation of p_e on the OP performance. The OP curves clearly show that the OP performance and the decreasing ADO due to increasing p_e is limited by that of pure OSTBC (i.e., G_3 -OSTBC with no antenna selection) case. Figure 4.2 also shows the perfect agreement between theoretical and simulation results.

In Figure 4.3, the SER performances of TAS/ G_2 -OSTBC scheme using QPSK signals are given for $n_R = 3$ and several n_T , in Rayleigh fading channels ($m = 1$) in the presence of FEs. For the average BER values of $p_e = 0.01$, $p_e = 0.2$, $p_e = 0.5$ and $p_e = 0.7$, the SER curves of TAS/ G_2 -OSTBC schemes are depicted. Besides, for the average BER values of $p_e = 0.0001$, $p_e = 0.05$, the SER curves of TAS/ G_2 -OSTBC $\{n_T = 4, n_R = 3\}$ scheme are included in order to point out that the SER performances face low SNR degradations for slight FE cases.

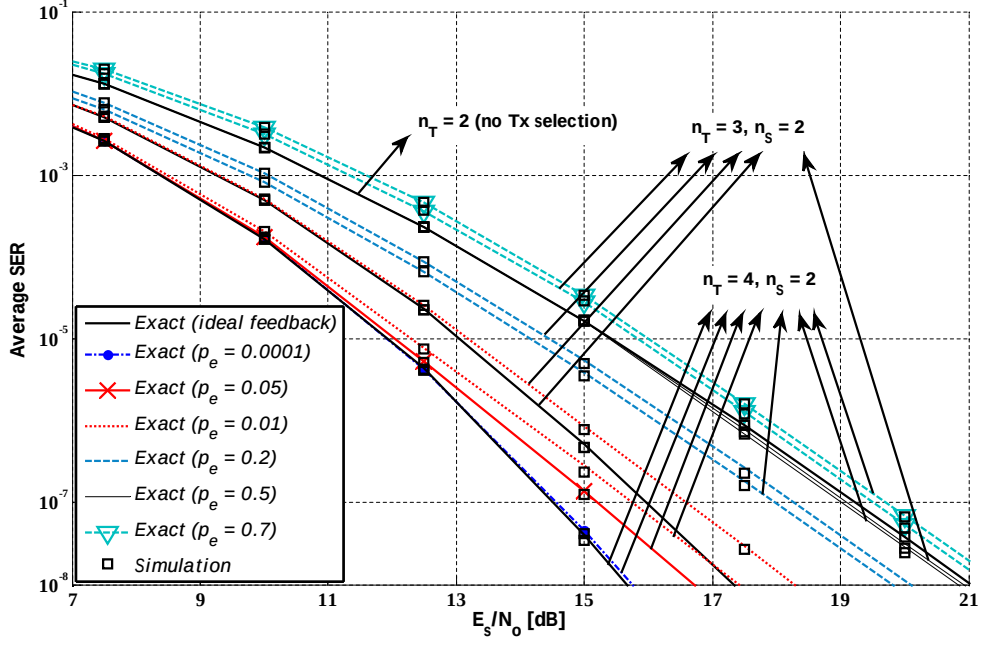


Figure 4.3: Average SER vs. average SNR per symbol for TAS/G₂-OSTBC scheme using QPSK signals for several n_T and p_e ($n_R = 3$, $m = 1$ (Rayleigh fading channels)).

The SER performances of TAS/G₃-OSTBC scheme using 16-QAM signals in the presence of FEs are given in Figure 4.4 for $n_R = 2$ and several n_T , in One-sided Gaussian fading channels ($m = 0.5$). By examining the SNR difference of the ideal performance curves and the erroneous ones as seen in Table 4.2, the effect of FEs on the SER performance of TAS/OSTBC scheme can be observed.

Table 4.2: SNR degradation of TAS/OSTBC&MRC at a SER of 10^{-5} .

Scheme	TAS/G ₂ -OSTBC		TAS/G ₃ -OSTBC	
n_R	3		2	
n_T	3	4	4	5
Degradation for $p_e = 0.01$	0.14 dB	0.36 dB	0.45 dB	0.80 dB
Degradation for $p_e = 0.20$	1.40 dB	2.20 dB	2.00 dB	5.50 dB
Degradation for $p_e = 0.50$	2.40 dB	3.50 dB	3.90 dB	6.10 dB
Degradation for $p_e = 0.70$	2.85 dB	4.10 dB	4.00 dB	6.90 dB

As seen from Figures 4.3 and 4.4, the SER performances related to the erroneous feedback cases are upper bounded by the pure OSTBC (i.e., OSTBC with no transmit antenna selection) case for the BSC BER values satisfying $p_e \leq 0.5$ whereas they

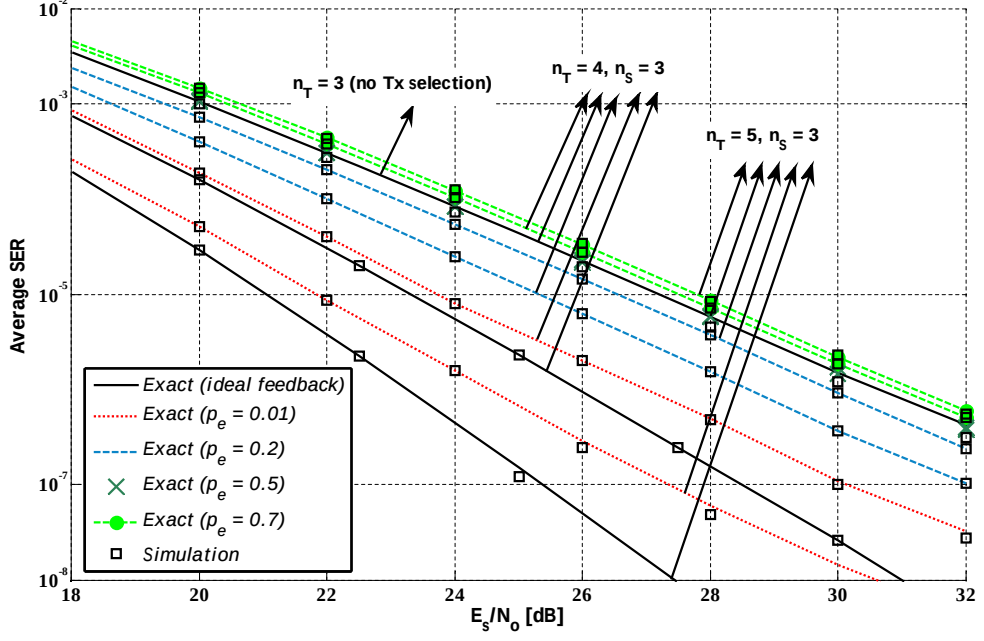


Figure 4.4: Average SER vs. average SNR per symbol for TAS/G₃-OSTBC scheme using 16-QAM signals for several n_T and p_e ($n_R = 2$, $m = 0.5$ (One-sided Gaussian fading channels)).

get worse for the higher BSC BER values such as $p_e = 0.7$. Besides, the EDO of TAS/OSTBC schemes in the presence of FEs would be lower bounded by that of the pure OSTBC case (which is mn_Rn_S). The SER curves related to JTRAS/G₂-OSTBC scheme using QPSK signals in the presence of FEs are given in Figure 4.5 for $n_R = 3$ and several n_T , in Rayleigh fading channels ($m = 1$).

The BER performances of JTRAS/G₃-OSTBC scheme using CBFSK signals in the presence of FEs are given in Figure 4.6 for $n_R = 2$ and several n_T , in Nakagami- m fading channels ($m = 2$). Similar to Figures 4.3 and 4.4, the effect of FEs on the SER performance of JTRAS/OSTBC schemes can be observed by interpreting the SNR difference of the ideal performance curves and the erroneous ones given in Figures 4.5 and 4.6. The SNR degradation values corresponding to JTRAS/OSTBC schemes are listed in Table 4.3. The variation of the error performance of JTRAS/G₂-OSTBC

$\{n_T = 3\}$ scheme due to the variation of the number of receive antennas ($n_R \in \{1, 2, 3\}$) and BSC BER values ($p_e \in \{0.01, 0.1\}$) has been depicted in Figure 4.7 for BPSK modulation signals and Rayleigh fading case ($m = 1$). For a BER of 10^{-5} and BSC BER of $p_e = 0.01$, the SNR degradation values for $n_R = 1, 2$ and 3 are seen to be

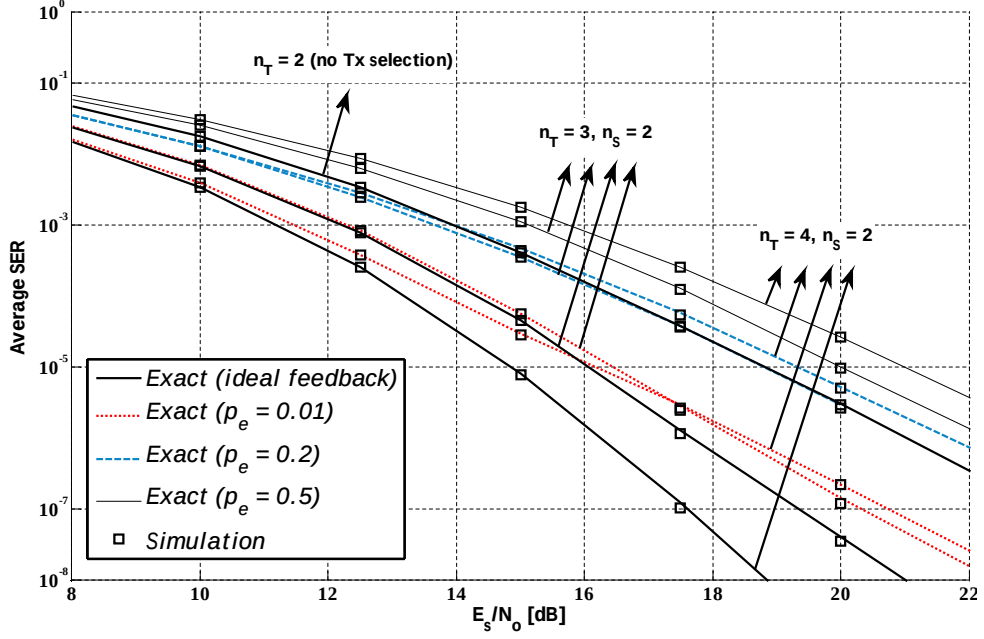


Figure 4.5: Average SER vs. average SNR per symbol for JTRAS/G₂-OSTBC scheme using QPSK signals for several n_T and p_e ($n_R = 3$, $m = 1$ (Rayleigh fading channels)).

Table 4.3: SNR degradation of JTRAS/OSTBC at a SER of 10^{-5} .

Scheme	JTRAS/G ₂ -OSTBC		JTRAS/G ₃ -OSTBC	
n_R	3		2	
n_T	3	4	4	5
Degradation for $p_e = 0.01$	0.40 dB	1.30 dB	0.30 dB	0.25 dB
Degradation for $p_e = 0.20$	2.75 dB	4.50 dB	1.10 dB	1.75 dB
Degradation for $p_e = 0.50$	4.00 dB	6.00 dB	1.65 dB	2.60 dB

1.6 dB, 1.2 dB and 1.1 dB, whereas for a BSC BER of $p_e = 0.1$, these values are 5.7 dB, 3.7 dB and 3.1 dB. Figure 4.7 shows that the SNR degradation caused by two different FE cases ($p_e = 0.01$ and $p_e = 0.1$) are decreasing for the increasing values of n_R . The EDO of JTRAS/OSTBC scheme in the presence of FEs would be lower bounded by the ADO provided by the OSTBC&RAS scheme (i.e., JTRAS/OSTBC with no TAS) similar to the TAS/OSTBC case. However, for the BSC BER values for which the erroneous feedback cases of TAS/OSTBC schemes are still superior or comparable to pure OSTBC scheme (e.g., $p_e = 0.2$ and $p_e = 0.5$), the error performance curves related to JTRAS/OSTBC scheme are not upper bounded by the OSTBC&RAS case (that can easily be seen from the scenarios in Figures 4.5 and 4.6). The robustness

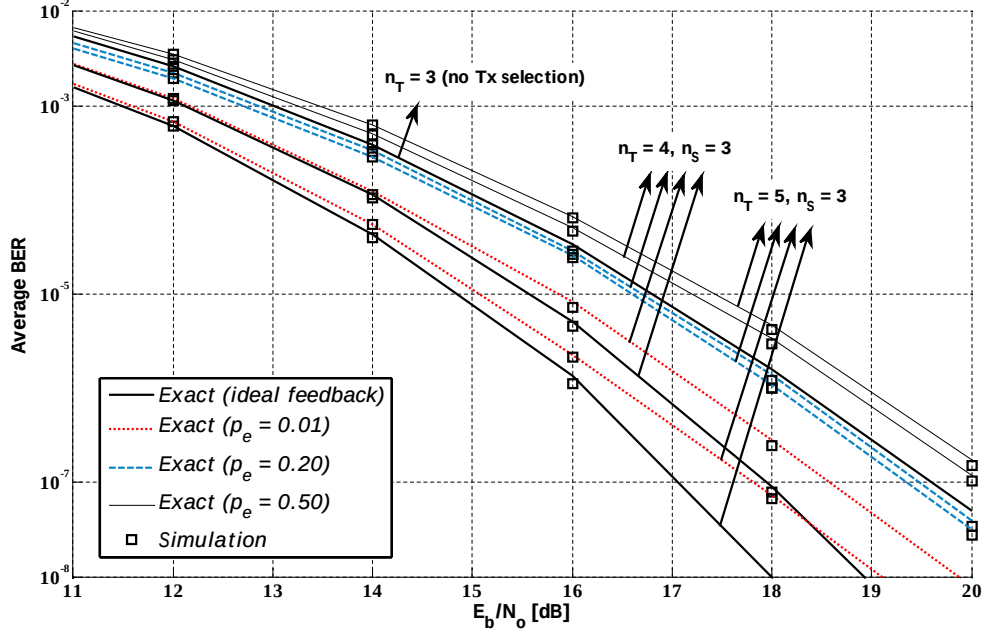


Figure 4.6: Average BER vs. average SNR per bit for JTRAS/G₃-OSTBC scheme using CBFSK signals for several n_T and p_e ($n_R = 2$, $m = 2$).

of TAS/OSTBC scheme when compared to JTRAS/OSTBC scheme against FEs can also be observed by comparing the SNR degradation values of both schemes for the same BSC BER values given in Figures 4.3 and 4.5 (also the values given in Table 4.2 and 4.3 for G₂-OSTBC). The SER curves in Figures 4.3-6 clearly show that increasing the number of total transmit antennas results in performance improvement of both schemes for some p_e values smaller than a specific threshold. However, they point out the fact that increase in the total number of transmitter antennas of both schemes deteriorates the error performance for p_e values that are greater than a specific threshold, it would be important to note that this specific threshold value of p_e is observed to be greater for TAS/OSTBC than that of JTRAS/OSTBC. This result reveals the sensitive selection structure that JTRAS scheme has against the MRC-like combining structure of TAS/OSTBC scheme which can also be interpreted as the superiority of conventional TAS/OSTBC scheme in the presence of FEs when compared to the JTRAS/OSTBC scheme. The differing behaviors of these schemes in the presence of FEs are heavily related to the selection criteria employed. The selection structure employed in TAS/OSTBC scheme causes impairments in transmit diversity in the presence of FEs while perfectly maintaining the receiver diversity (provided by MRC-like structure). Whereas, in the presence of FEs, the joint selection structure

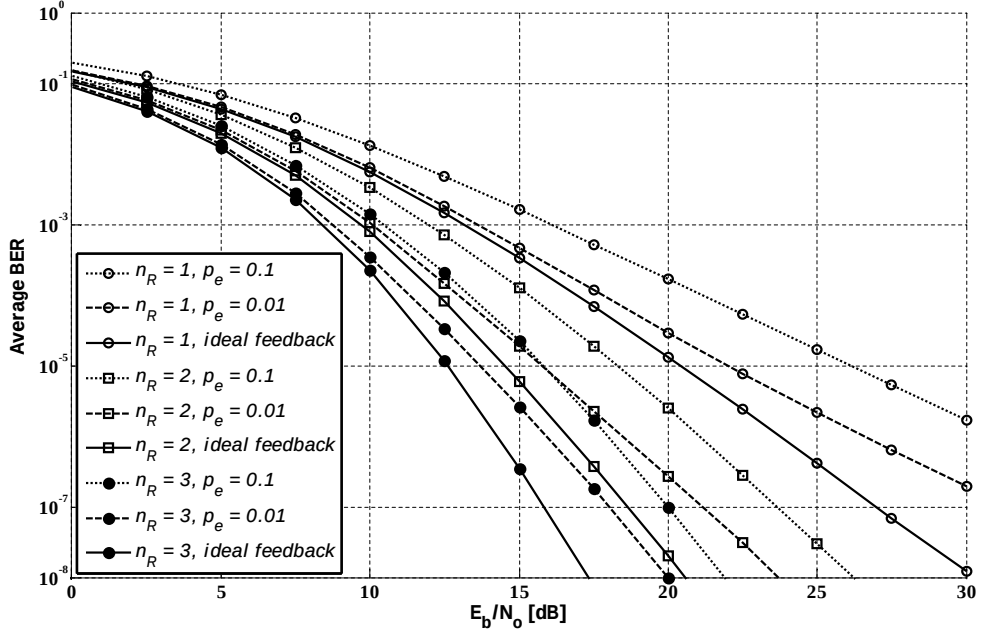


Figure 4.7: Average BER vs. average SNR per bit for JTRAS/G₂-OSTBC ($n_T = 3$, $n_S = 2$) scheme using BPSK signals and Rayleigh fading channels ($m = 1$) ($p_e \in \{0.01, 0.1\}$, $n_R \in \{1, 2, 3\}$).

employed in JTRAS/OSTBC scheme would face impairments both at transmit and receive antenna selection that may result in the exclusion of the best transmit-receive links. Besides, Monte Carlo simulation results have validated the theoretical results in all performance figures.

4.2.5. Conclusions for the Effects of FE on the Performances of JTRAS/OSTBC and TAS/OSTBC Schemes

This section has presented the extensive performance examination of the unified diversity scheme covering both JTRAS/OSTBC and TAS/OSTBC schemes. Focusing on the practical feedback imperfections (i.e., FEs) in Nakagami- m fading channels, detailed analyses have been provided for the exact expressions related to the statistics of the output SNR (PDF, CDF and MGF) and the capacity and error performances (OP and BER/SER) for both schemes. Also, by examining the ADO analysis of the unified scheme, it is shown that both schemes achieve full diversity order (i.e., mn_Rn_T) for ideal conditions, while in the presence of FEs they keep maintaining the ADO provided by the OSTBC (with no transmit antenna selection) scheme itself and the receive diversity order provided by RAS or MRC-like structure (i.e., mn_Rn_S). Hence,

the robustness that is gathered on the ADO by the combined TAS&OSTBC structure employed at the transmitters of both TAS/OSTBC and JTRAS/OSTBC schemes exposes the importance of these schemes in practical system scenarios when compared to the single TAS scheme that is shown in [19] to lose the transmitter diversity and to face severe deteriorations in the presence of FEs. As indicated in Figures 4.2-7, for the ideal and erroneous conditions of both schemes, the OP and BER/SER performances can easily achieve considerable enhancements brought along with higher diversity orders provided by the increment in the number of transmit and receive antennas for BSC BER values that are smaller than a specific threshold. Also, this threshold is seen to be greater for TAS/OSTBC scheme than that of JTRAS/OSTBC scheme which clearly point out the superiority of TAS/OSTBC scheme to JTRAS/OSTBC scheme especially in the presence of FEs. This superiority is provided by the virtue of multiple RF chains employed in the receiver side of TAS/OSTBC. However the simple receiver structure of JTRAS/OSTBC (employing only a single RF chain) tends this scheme to be a more feasible technique for multi-antenna communications systems rather than TAS/OSTBC scheme. Also, for reasonable average BER values (such as $p_e = 0.01$) specified for the feedback link, both schemes still achieve considerable performances (with tolerable SNR degradation) that provide both schemes to maintain significant importance in real-world wireless communications system design.

4.3. Feedback-Rate Efficient TAS/Alamouti-OSTBC Scheme with Robust Error Performance in the Presence of FEs¹

This section, proposing a new modified TAS/Alamouti-OSTBC scheme (with MRC-like receiver) with reduced feedback-rate (and full-diversity for ideal feedback cases), analyzes the error performance over flat Nakagami- m fading channels in the presence of FEs. In contrast to many of the research papers in the literature and to the analysis provided in Section 4.2, the performance analyses throughout the following two sections consider the feedback link to have identical fading characteristics as the main transmission link. The error performance results highlight

¹This work is supported by the Scientific and Technological Research Council of Turkey (TÜBİTAK) under grant EEEAG-112E010

the robustness of the proposed scheme under imperfect feedback conditions when compared to the conventional TAS/Alamouti-OSTBC scheme and other modified TAS/Alamouti-OSTBC schemes proposed in [78].

Under ideal channel estimation and delay-free feedback assumptions, statistical expressions and performance metrics related to the post-processing SNR of the proposed scheme are derived for Nakagami- m fading channels. Exact expressions for OP and BERs/SERs of M -ary signals are presented in order to provide an extensive examination on the system performance in the presence of FEs. The proposed scheme is shown to provide not only full-diversity order as the conventional and other reduced feedback-rate schemes for perfect feedback case but also superior error performance in the presence of FEs. The detailed analysis in this section makes the following specific contributions:

- The PDF and CDF of the output SNR,
- The system capacity examination from the OP perspective.
- The MGF of the output SNR and exact error probability expressions of M -ary signals for erroneous cases by considering a more realistic feedback link model (i.e., assuming that the feedback link is reciprocal to the main transmission link).

The remainder of this section is organized as follows. Section 4.3.1 presents the proposed scheme and the channel model, briefly defines the statistics of the output SNR by deriving the PDF, CDF and OP expressions. Section 4.3.2 presents the derivation of the unified expression covering the MGF of the output SNR and exact (and approximate for M -PSK, $M \geq 8$) BER/SER performance expressions. In Section 4.3.3, some numerical results are presented that are validated by simulations for average error rates of M -ary signals. Finally, Section 4.3.4 draws conclusions about the analytical and numerical results.

4.3.1. System Model and SNR Statistics for the Modified TAS/Alamouti-OSTBC Scheme

The proposed scheme is constructed as a closed-loop multi-antenna system with n_T and n_R antennas at transmit and receive ends, respectively. The transmitter side employs hybrid TAS/Alamouti-OSTBC scheme with some modifications on

the antenna selection/switching structure of the conventional TAS/Alamouti-OSTBC scheme [45], [78]. By setting one of the transmit antennas fixed for all transmission intervals and performing TAS among $n_T - 1$ ones, proposed scheme reduces the feedback rate while providing full-diversity order transmission in ideal feedback cases and a robust error performance in the presence of FEs. The number of bits required for the feedback data would be reduced to $k_{ps} = \lceil \log_2(n_T - 1) \rceil$ whereas this value was $k_{cs} = \lceil \log_2\binom{n_T}{2} \rceil$ for the conventional TAS/Alamouti-OSTBC with antenna subset selection. As clearly seen from Table 4.4, for a given n_T , number of feedback bits needed by the modified TAS/Alamouti-OSTBC is lower than or comparable to those needed for the modified TAS/Alamouti-OSTBCs (denoted as Scheme-1 (S-1), Scheme-2 (S-2) and Scheme-3 (S-3)) given in [78]. Considering

Table 4.4: Feedback rate variation of the proposed scheme compared to conventional scheme and other reduced-rate schemes.

	$n_T = 3$	$n_T = 4$	$n_T = 5$	$n_T = 6$	$n_T = 7$
Scheme-1	1	2	3	4	4
Scheme-2	2	2	3	3	3
Scheme-3	1	1	2	2	2
Proposed Sch.	1	2	2	3	3
Conv. Scheme	2	3	4	4	5

that the channel coefficients $h_{i,j}, i = 1, 2, \dots, n_R, j = 1, 2, \dots, n_T$ are estimated perfectly, the receiver combines the independent signal replicas gathered from the j^{th} transmit antenna through n_R receive antennas during the Alamouti codeword interval by using the optimal combiner weights (i.e., MRC) and puts out the combined SNR: $\gamma_j = \sum_{i=1}^{n_R} |h_{i,j}|^2$. While the envelopes $|h_{i,j}|$ are assumed to be i.i.d. Nakagami- m random variates with fading parameter m and squared mean $\Omega = E[|h_{i,j}|^2]$, γ_j can be easily shown to be a Gamma random variate with PDF and CDF given as $f(x) = (m/\bar{\gamma})^{mn_R} x^{mn_R-1} e^{-xm\bar{\gamma}} / \Gamma(mn_R), x \geq 0$, and $F(x) = \psi(mn_R, xm\bar{\gamma}) / \Gamma(mn_R), x \geq 0$, respectively. By comparing the instantaneous SNRs $\gamma_j, j = 2, 3, \dots, n_T$, the receiver determines the transmit antenna that will be used for transmission: $\gamma_{max} = \max_{2 \leq j \leq n_T} \{\gamma_j\}$. The index of the selected transmit antenna is fed to the transmitter back over a feedback link that is considered to have fading characteristics from a

more realistic viewpoint [19], [50]. Since the feedback channel is assumed to be reciprocal to the main communications channel, the fading parameter m is identical for both signaling directions. The k_{ps} -bit feedback information is sent to the transmitter followed by an L_{ps} -PSK modulation where $L_{ps} = 2^{k_{ps}}$. Due to degrading effects of the feedback channel with fading characteristics, the index information would be decoded erroneously. With the probability p_k , $k = 1, 2, \dots, n_T - 1$, the k^{th} best transmit antenna will be activated together with the fixed transmit antenna while the others are kept silent. The a priori probabilities p_k can be easily defined as $p_k = \sum_{m=1}^{K_{ps}} p(s_m)P(s_m \rightarrow s_k)$, where $K_{ps} = n_T - 1$ and s_m denotes the baseband symbol belonging to L_{ps} -PSK modulation, $p(s_m)$ represents the occurrence probability of the symbol s_m and $P(s_m \rightarrow s_k)$ is the PEP related to the k^{th} and m^{th} symbols corresponding to the k^{th} and m^{th} best antennas. The PEP values have been evaluated for different modulation levels (L_{ps}) in the literature [19]. Based on the assumption of feedback channel with fading characteristics, the total error probability expression can be expressed as $P_s(\bar{\gamma}_{FC}) = \sum_{k=1}^{K_{ps}} p_k P_{s,k}(\bar{\gamma}_{FC})$. Here, $P_{s,k}(\bar{\gamma}_{FC})$ is the average SER related to the k^{th} best transmit antenna together with the fixed one and $\bar{\gamma}_{FC}$ is the average SNR of the feedback channel. The feedback message, delivering k_{ps} bits in order to represent $K_{ps} = n_T - 1$ transmit antennas, can be decoded to $L_{ps} = 2^{k_{ps}} \geq K_{ps}$ index values in the presence of FEs. This may cause the proper transmit antenna indices $(1, 2, \dots, K_{ps})$ to be de-mapped to one of the improper transmit antenna indices $(K_{ps}+1, K_{ps}+2, \dots, L_{ps})$ (which can be called FF). Although many re-transmission or error control coding protocols such as ARQ can be employed to overcome the FF, it is assumed that the transmitter arbitrarily selects and activates one of the transmit antennas in order not to reduce the transmission rate that would be brought along with the usage of a re-signaling structure. Thus, assuming that each transmit antenna is selected equally-likely in the presence of FF as in [84] the a priori probabilities would be increased by the factor $(1/K_{ps}) \sum_{j=K_{ps}+1}^{L_{ps}} p_j$ yielding $p'_k = p_k + \frac{1}{K_{ps}} \sum_{j=K_{ps}+1}^{L_{ps}} p_j$, $k = 1, 2, \dots, K_{ps}$. Hence, it would be possible to concern also the case of $L_{ps} > K_{ps}$ unlike [45] which considers $K_{ps} = 2^b$, $b \in \mathbb{Z}^+$.

After switching the k^{th} best transmit antenna as the result of erroneous decoding, the resulted instantaneous post-processing SNR at the receive-combiner output will be $\gamma_{o,k} = \frac{1}{2}(\gamma_{max,k} + \gamma_1)$ where $\gamma_{max,k}$ denotes the k^{th} maximum instantaneous SNR among $n_T - 1$ transmit branch SNRs and γ_1 denotes the instantaneous SNR of the

fixed transmit branch. The total transmit power is equally allocated to both transmit antennas that yields the factor $\frac{1}{2}$. By simple modifications on [45, (6)], the PDF of $\gamma_{max,k}$ can be obtained as $f_{(k)}(x) = (n_T - 1) \binom{n_T-2}{k-1} f(x) [1 - F(x)]^{k-1} F(x)^{n_T-k-1}$. Binomial expansion over the series definition of $F(x)$ given in [70, (10)] yields

$$f_{(k)}(x) = a_0 \sum_{p=0}^{k-1} a_1 \sum_{q=0}^{n_T-k+p-1} a_2 \sum_{r=0}^{q(mn_R-1)} \frac{a_3 x^{r+mn_R-1}}{e^{xm(1+q)/\bar{\gamma}}} \quad (4.31)$$

Note that, (4.31) is obtained for arbitrary values of $m \geq 1/2$ that satisfy the product mn_R to be integer (i.e., $mn_R \in \mathbb{Z}^+$). In (4.31), the parameters are defined as $a_0 = (n_T - 1) \binom{n_T-2}{k-1} / \Gamma(mn_R) (m/\bar{\gamma})^{-mn_R}$, $a_1 = \binom{k-1}{p} (-1)^p$, $a_2 = \binom{n_T-k+p-1}{q} (-1)^q$ and $a_3 = \mu_r(q, mn_R)$ are the multinomial coefficients [62, (0.314)]. The MGF of the instantaneous output SNR $\gamma_{o,k}$ can be easily obtained via the product of MGFs of the instantaneous SNRs $\gamma_{max,k}$ and γ_1 : $\mathcal{M}_{\gamma_{o,k}}(s) = \mathcal{M}_{\gamma_{max,k}}(\frac{s}{2}) \mathcal{M}_{\gamma_1}(\frac{s}{2})$. By performing the MGF integral [19, (7.44)] over (4.31), $\mathcal{M}_{\gamma_{max,k}}(\frac{s}{2})$ can be obtained as

$$\mathcal{M}_{\gamma_{max,k}}\left(\frac{s}{2}\right) = a_0 \sum_p a_1 \sum_q a_2 \sum_r \frac{a_3 \Gamma(r + mn_R)}{\left[\frac{s+s_q}{2}\right]^{r+mn_R}} \quad (4.32)$$

where $s_q = (2m(q+1))/\bar{\gamma}$. Also, the MGF of γ_1 has been derived in [19, (6.65)] as $\mathcal{M}_{\gamma_1}(\frac{s}{2}) = (2m/\bar{\gamma})^{mn_R} (s + \frac{2m}{\bar{\gamma}})^{-mn_R}$. Thus, the MGF expression of the output SNR can be written as [74]

$$\mathcal{M}_{\gamma_{o,k}}(s) = \bar{a}_0 \sum_p a_1 \sum_q a_2 \sum_r \frac{a_3 a_4 [s + s_0]^{mn_R}}{[s + s_q]^{r+mn_R}} \quad (4.33)$$

where $\bar{a}_0 = a_0 s_0^{mn_R}$, $a_4 = \Gamma(r + mn_R) / 2^{r+mn_R}$ and $s_0 = 2m/\bar{\gamma}$. In order to obtain a simplified and univariate expression of the CDF related to the output SNR, the inverse Laplace transform is applied over (4.33) as

$$F_{\gamma_{o,k}}(x) = \bar{a}_0 \sum_p a_1 \sum_q a_2 \sum_r a_3 a_4 \mathcal{L}^{-1} \left\{ \frac{[s + s_0]^{-c_0}}{s[s + s_q]^{c_r}} \right\} \quad (4.34)$$

where $c_r = r + mn_R$. Inverse Laplace transform can be obtained by using [72, (2.1.4-8)] that is valid for $s_i \neq s_j \neq 0, \forall i \neq j$. In order to utilize this identity, the coefficients $s_k, k = 1, 2, 3, (s_1 = 0, s_2 = s_0, s_3 = s_q)$ are renamed as $\hat{s}_d, d = 1, \dots, N \leq 3$, such that all the coefficients ($\hat{s}_d$) are different from each other. The powers of equal coefficients will sum up and result in the new powers ($\hat{c}_d$) (i.e., $\hat{c}_1 = c_1 + c_2$ if $s_1 = s_2$). Thus, the CDF in (4.34) can be expressed as [74]

$$F_{\gamma_{o,k}}(x) = \bar{a}_0 \sum_p a_1 \sum_q a_2 \sum_r a_3 a_4 \sum_{n=1}^N \sum_{l=1}^{\hat{c}_n} \frac{a_5 x^{\hat{c}_n - l}}{e^{x \hat{s}_n}} \quad (4.35)$$

where $a_5 = \frac{A_{nl}(-\hat{s}_n)}{(\hat{c}_n - l)!(l-1)!}$, $A_{nl}(\cdot)$ is defined in [72, (2.1.4-8)]. The derivation of (4.35) helps obtaining not only the exact error probability expressions that would be handled in the following section but also the OP which can be expressed as $P_{out} = F_{\gamma_o}(2^{\mathcal{R}} - 1)$ [70], [84] where $\mathcal{R}$ denotes the desired value of spectral efficiency. By the weighted sum of OPs related to each selection case (i.e., $P_{out,k}$) and using the a priori probabilities p'_k , the average OP can easily be evaluated: $P_{out} = \sum_{k=1}^{n_T-1} p'_k P_{out,k}$ where $P_{out,k}$ can be easily obtained by substituting $2^{\mathcal{R}} - 1$ into (4.35).

4.3.2. Exact Error Performance Analysis for the Modified TAS/Alamouti-OSTBC Scheme

Based on the output SNR statistics, this section derives the MGF and BER/SER expressions of the proposed scheme. Hence, the analysis is focused on the unified integral definition given in [73] as

$$\mathcal{J}_{\gamma_{o,k}}(\theta, \epsilon, \Phi) = \theta \int_0^\infty x^\epsilon e^{-x\Phi} F_{\gamma_{o,k}}(x) dx \quad (4.36)$$

Here, setting different values in the parameters of this function ($\theta, \epsilon; \Phi$) will result in the MGF and the BER/SER of some binary/ M -ary modulations. In addition to this unified function, the unified function given in (3.48) is used in order to cover the SER expressions of some M -ary modulations as well:

$$\hat{\mathcal{J}}_{\gamma_{o,k}}(\theta, \Phi) = \theta \int_0^\infty e^{-x\Phi} F_{\gamma_{o,k}}(x) {}_1F_1\left(1; \frac{3}{2}; \frac{x\Phi}{2}\right) dx \quad (4.37)$$

In (4.37), ${}_1F_1(\cdot, \cdot, \cdot)$ denotes the confluent hypergeometric function [62, (210.1)]. By considering both (4.36) and (4.37), it is possible to evaluate the MGF of $\gamma_{o,k}$ by using the identity $\mathcal{M}_{\gamma_{o,k}}(s) = \mathcal{J}_{\gamma_{o,k}}(s, 0, s)$, BER of BPSK, CBFSK, NCBFSK and DBPSK signals and the approximate SER of M -PSK signals ($M \geq 8$) by using the identity $P_{s,k} = \mathcal{J}_{\gamma_{o,k}}\left(\frac{\lambda_3(\lambda_2^{\lambda_1})}{2\Gamma(\lambda_1)}, \lambda_1 - 1, \lambda_2\right)$, SER of M -PAM signals by using the identity $P_{s,k} = \mathcal{J}_{\gamma_{o,k}}\left(\sqrt{\frac{3(M-1)}{\pi M^2(M+1)}}, \frac{-1}{2}, \frac{3}{(M^2-1)}\right)$ and SER of QPSK and M -QAM signals by using the identity $P_{s,k} = \mathcal{J}_{\gamma_{o,k}}\left(\sqrt{\frac{\lambda_4}{8\pi}}(\lambda_5 - \lambda_6), \frac{-1}{2}, \frac{\lambda_4}{2}\right) + \hat{\mathcal{J}}_{\gamma_{o,k}}\left(\frac{\lambda_4\lambda_6}{2\pi}, \lambda_4\right)$. The parameters for binary modulations and M -PSK signals are defined for BPSK, CBFSK, NCBFSK, DBPSK, M -PSK (for $M \geq 8$), QPSK and M -QAM signals are defined in Table 3.1. Thus, deriving the exact expressions of (4.36) and (4.37) will lead us to obtain and examine the performance metrics of the proposed scheme for several modulations. By substituting (4.35) into the unified integrals in (4.36) and (4.37), and rearranging the resulting integrals in terms of the Gamma function and the Gauss hypergeometric function ${}_2F_1$ [62, (3.197-3)], respectively, the unified performance expressions can be obtained as [74]

$$\mathcal{J}_{\gamma_{o,k}}(\theta, \epsilon, \Phi) = \theta \bar{a}_0 \sum_p a_1 \sum_q a_2 \sum_r a_3 a_4 \sum_n \sum_l a_5 \frac{\Gamma(\epsilon + 1 + \hat{c}_n - l)}{(\Phi + \hat{s}_n)^{\epsilon+1+\hat{c}_n-l}} \quad (4.38)$$

and

$$\begin{aligned} \hat{\mathcal{J}}_{\gamma_{o,k}}(\theta, \epsilon, \Phi) &= \theta \bar{a}_0 \sum_p a_1 \sum_q a_2 \sum_r a_3 a_4 \sum_n \sum_l a_5 \\ &\times \frac{{}_2F_1\left(1; 2 + \hat{c}_n - l; \frac{3}{2}; \frac{\hat{s}_n}{\Phi + \hat{s}_n}\right) \Gamma(1 + \hat{c}_n - l)}{(\Phi + \hat{s}_n)^{1+\hat{c}_n-l}} \end{aligned} \quad (4.39)$$

As in the OP analysis, the average error probability performance of the proposed scheme can be obtained by simply averaging over all possible transmit antenna combinations.

4.3.3. Numerical Results for the Modified TAS/Alamouti-OSTBC Scheme in the Presence of FEs

In order to examine the performance of the proposed scheme for erroneous feedback cases in Nakagami- m fading channels, and to compare it to other schemes,

this subsection presents numerical results for BPSK and QPSK signals, unit channel power ($\Omega = 1$) and total transmit energy of $E_s = 1$ and several average SNRs specified for the feedback link. Both figures include the theoretical and simulation results related to the proposed scheme, whereas the numerical results related to the conventional scheme and other schemes have been obtained via Monte Carlo simulations as in [45]. In Figure 4.8, the SER performances of all schemes using QPSK signals under ideal feedback assumption are given for $n_T = 4$, $n_R = 2$ and $m = 0.5$. In Figure 4.8, in the absence of FEs, conventional scheme has the best SER performance as also stated in [45]. The SNR degradation caused by the sub-optimal antenna selection strategies of the proposed scheme, S-1, S-2 and S-3 are 0.83 dB, 0.37 dB, 1 dB, 0.83 dB, respectively at a SER of 10^{-5} . Under error-free feedback conditions, proposed scheme, S-1, S-2 and S-3 all provide comparable error performance with full-diversity order (i.e., mn_Rn_T) while also reducing the feedback bit-rate as shown in Table 4.4. Figure 4.8 also presents the error performance variation of the proposed scheme due to the variation of average SNR of the feedback link ($\bar{\gamma}_{FC,dB} \in -40, -10, 0, 10, 20, 30, 40$). The effect of $\bar{\gamma}_{FC}$ on the SER performance of the proposed scheme can be observed by interpreting the SNR difference between the ideal performance curve (the solid line given at the bottom) and the erroneous ones. For average SNRs of 20 dB, 30 dB and 40 dB, the SNR degradations are approximately 4.5 dB, 2.3 dB and 0.9 dB, respectively at a SER of 10^{-5} . The SER curves for all $\bar{\gamma}_{FC,dB}$ are upper bounded by that of the pure Alamouti (i.e., Alamouti coding w/o TAS) scheme. Besides, the ADO of the proposed TAS/Alamouti-OSTBC scheme at erroneous feedback cases is equal to the ADO provided by the pure Alamouti case (i.e., $2mn_R$).

SER curves of QPSK signals for ideal and erroneous cases are presented in Figure 4.9 for $n_T = 5$, $n_R = 1$ and $m = 1$. For $\bar{\gamma}_{FC,dB} = -10$ dB, the proposed scheme has provided the best performance with an approximate SNR advantage of 1 dB to the conventional scheme at a SER of 10^{-3} , while this value is 0.37 dB, 0.3 dB, -0.2 dB, respectively for S-2, S-3 and S-1 when compared to the conventional scheme. Also, for the SER curves related to a more reliable feedback link with an average SNR of $\bar{\gamma}_{FC,dB} = 5$ dB, M-TAS/Alamouti-OSTBC has provided the best SER performance with 2.8 dB SNR advantage to the conventional scheme at 10^{-3} . SNR advantages brought along with S-3, S-1 and S-2 when compared to the conventional scheme are 1.5 dB, 1 dB and -1.1 dB, respectively. For the ideal feedback case, the

conventional scheme is better than all sub-optimal schemes with SNR gains of 0.36 dB, 0.8 dB, 0.9 dB, 1.05 dB when compared to S-1, S-3, M-TAS/Alamouti-OSTBC and S-2, respectively at a SER of 10^{-3} . The scenario considered in Figure 4.9 has been examined again in Figure 4.10 for BPSK signals and the average feedback link SNR of $\bar{\gamma}_{FC,dB} = 10$ dB. Similar to the results of Figure 4.9, the BER curves depicted in Figure 4.10 show that the PS achieves an SNR gain of 2.3 dB, 2.9 dB, 4.9 dB and 5.1 dB when compared to S-3, S-2, CS and S-1 at a BER of 10^{-5} .

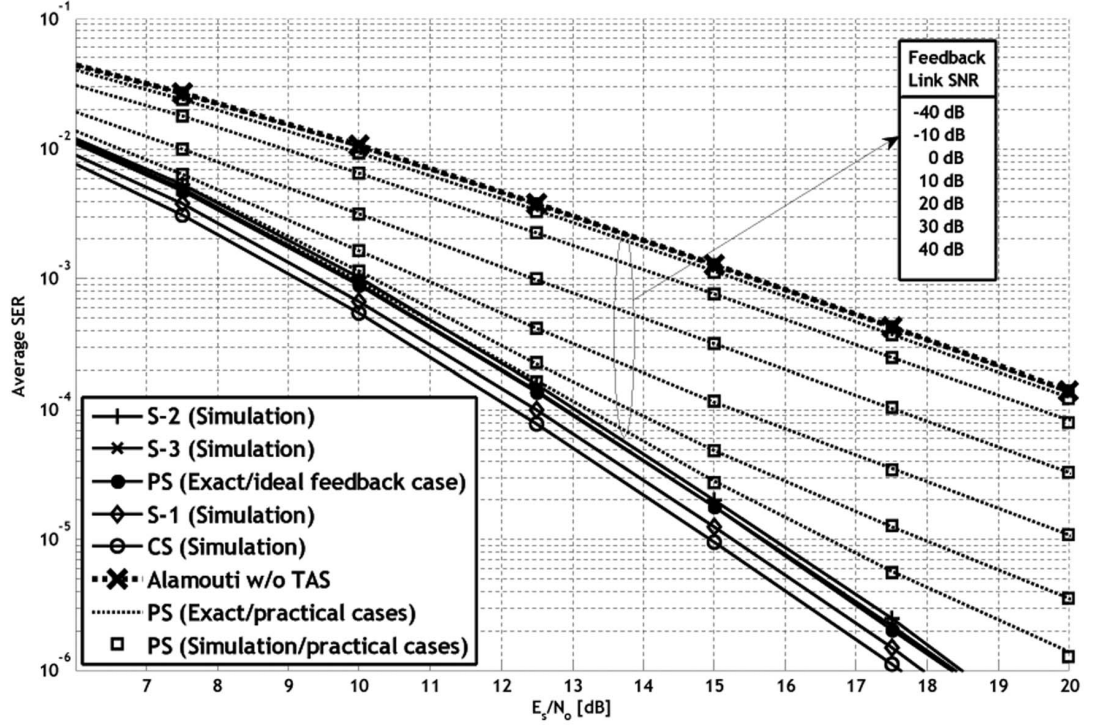


Figure 4.8: Average SER vs. $\bar{\gamma}$ for the proposed scheme compared to other schemes ($n_T = 4, n_R = 2, m = 0.5$) and the variation of SER due to $\bar{\gamma}_{FC}$.

4.3.4. Conclusions for the Effects of FEs on the Performance of the Modified TAS/Alamouti-OSTBC Scheme

We have proposed a reduced feedback-rate diversity scheme employing a modified TAS/Alamouti-OSTBC structure, and has exposed the robustness of this scheme by focusing on its performance in the presence of FEs. Considering a generic fading model like Nakagami- m , statistics related to the output SNR and the performance metrics have been presented for ideal feedback case and practical impairments. SER curves have shown that all schemes achieve full diversity order

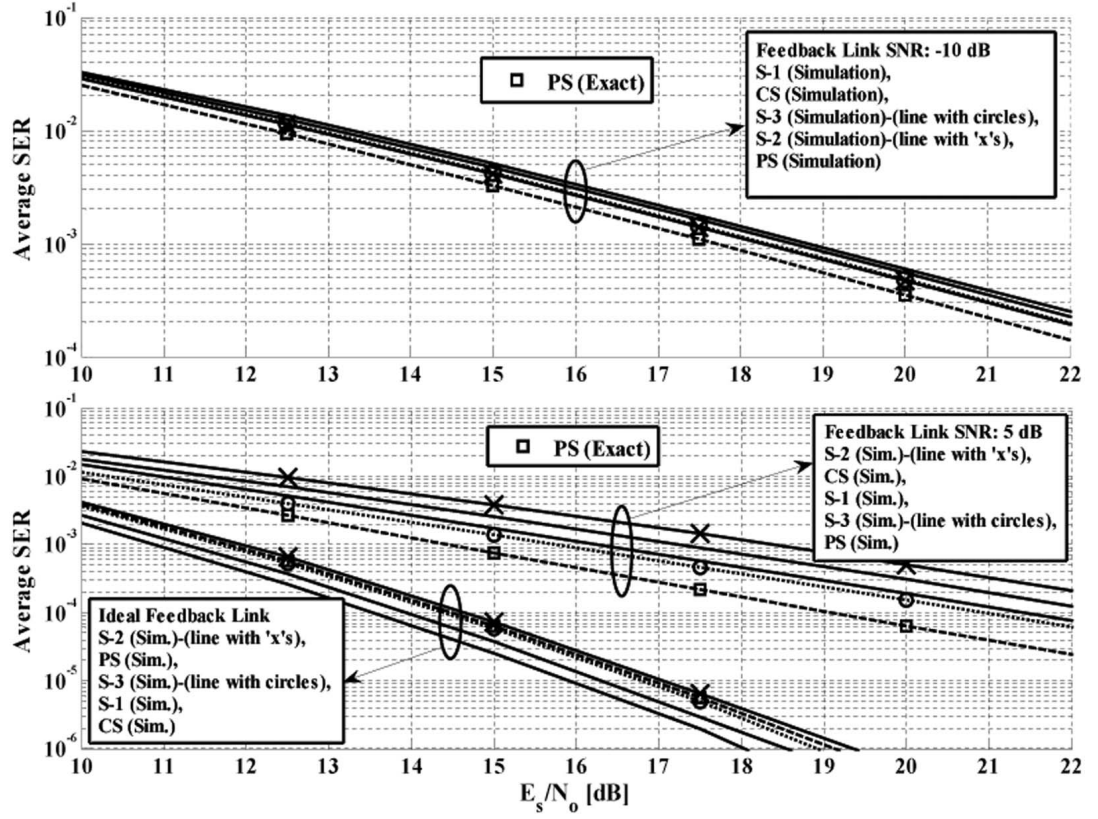


Figure 4.9: Average SER vs. $\bar{\gamma}$ for the proposed scheme compared to other schemes ($n_T = 5, n_R = 1, m = 1$, average SNRs -10 dB and 5 dB).

(mn_Rn_T) for ideal cases, while in the presence of FEs they maintain the ADO of $2mn_R$ that is provided by Alamouti scheme itself (w/o TAS) and the receive diversity provided by the MRC-like structure.

Performance results clearly point out the superiority of the proposed scheme to the conventional one, S-1, S-2 and S-3 in the presence of FEs. This superiority is provided by the compromise between the advantage of feedback rate reduction and the gain based on antenna subset selection strategy. The number of feedback bits per each index transmission limits the number of imperfect TAS cases under erroneous feedback conditions that directly degrades the error performance. Besides, the antenna selection strategy has a prepotent role on the system performance depending on the selection efficiency of most qualified transmit branches. Also, for higher values of $\bar{\gamma}_{FC, dB}$, M-TAS/Alamouti-OSTBC scheme still achieves considerable error performance that, together with the reduced feedback link traffic, provides this scheme to maintain significant importance in real-world wireless communications system design.

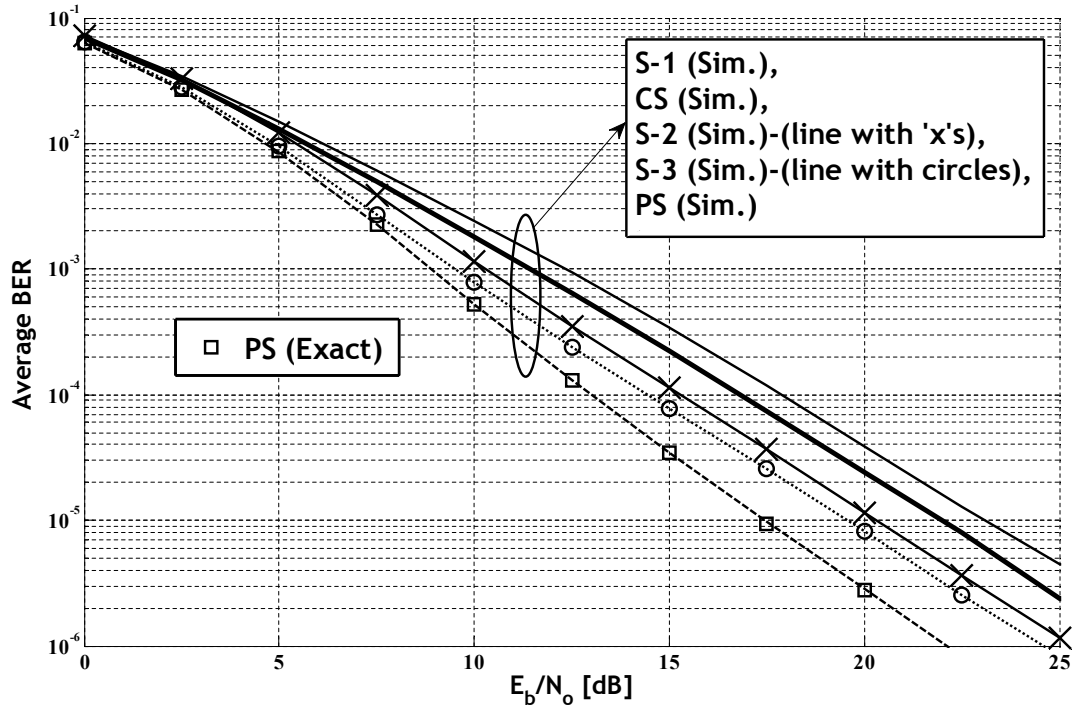


Figure 4.10: Average BER vs. $\bar{\gamma}$ for the proposed scheme compared to other schemes ($n_T = 5$, $n_R = 1$, $m = 1$, average feedback link SNR of 10 dB).

4.4. Feedback-Error-Resistant and Reduced Feedback-Rate Orthogonal Space-Time Block Coding Scheme Employing Single Transmit Antenna Selection²

A novel and generic transmit diversity scheme that employs combined single TAS/OSTBC scheme with reduced feedback-rate and robust error performance in the presence of FEs is introduced in this section. Statistical expressions and performance metrics related to the post-processing SNR of the proposed scheme are derived for Nakagami- m fading channels considering that the channel estimation process is error-free and the feedback information is delivered to the transmitter side with no delay. Exact expressions for OP and BERs/SERs of M -ary signals are presented in order to examine the system performance in the presence of FEs extensively. Also, the asymptotic analysis is provided to show the ADO and the asymptotic behavior of the proposed scheme. The proposed scheme is shown to provide not only full-diversity order as the conventional scheme for perfect feedback case but also superior error

²This work is supported by the Scientific and Technological Research Council of Turkey (TÜBİTAK) under grant EEEAG-112E010

performance in the presence of FEs. Besides, the asymptotic coding gain of the proposed scheme is investigated and compared to that of the conventional TAS/OSTBC scheme from different perspectives.

Two simplest members of the modified TAS/OSTBC (M-TAS/OSTBC) schemes are introduced in [74] (or Section 4.3) and [86], and the performance enhancements they provided are exposed by some limited perspectives. The modified scheme presented in [74] provides considerable reduction in feedback burden and robustness in the error performance in the presence of FEs by regardlessly using the first transmit antenna together with a selected one (among $n_T - 1$ ones) for G_2 -OSTBC codeword transmission. The exact SER performance of M-TAS/ G_2 -OSTBC scheme in the presence of FEs is examined in [74] whereas no insights for the OP and ADO of the diversity scheme. The other member of the M-TAS/OSTBC scheme, which has used the first two transmit antennas regardlessly together with a selected one (among $n_T - 2$ ones) for G_3 -OSTBC codeword transmission, has been introduced in [86]. The average SER performance of M-TAS/ G_3 -OSTBC scheme in Rayleigh fading channels is examined via Monte Carlo simulations and the ADO of the investigated scheme is observed by using a high SNR approximation. Whereas [86] has not provided the derivations of the exact performance metrics for a generic fading channel model like Nakagami- m .

By also including the exact derivations of and the important insights on the performance metrics that are missing in the consideration of [74] and [86], this section introduces and analyzes the feedback-error-resistant and reduced feedback-rate structure of the M-TAS/OSTBC schemes. Considering erroneous feedback conditions, the exact and asymptotic performance metrics are derived for flat Nakagami- m fading channels and M -ary modulations. In contrast to many of the research papers in the literature, the performance analysis throughout this section considers the feedback link to have identical fading characteristics as the main transmission link. The performance results highlight the robustness of the M-TAS/OSTBC schemes under imperfect feedback conditions when compared to the conventional TAS/OSTBC schemes.

The detailed analysis in this section makes the following specific contributions:

- The PDF and the CDF of the output SNR,
- the system capacity examination from the OP perspective,

- the MGF of the output SNR and exact error probability expressions of M -ary signals for erroneous cases by considering a more realistic feedback link model (i.e., assuming that the feedback link is reciprocal to the main transmission link),
- the asymptotic performance metrics such as the ADO and asymptotic coding gain (ACG) are derived for the modified scheme that yields to present an insightful comparison to the conventional scheme in the presence of feedback impairments.

The remainder of this section is organized as follows. Section 4.4.1 presents the investigated M-TAS/OSTBC schemes and the channel model, briefly define the statistics of the output SNR by deriving the PDF, CDF and MGF expressions. The OP analysis is provided in Section 4.4.2, while Section 4.4.3 presents the derivation of the unified expression covering the MGF of the output SNR and exact (and approximate for M -PSK, $M \geq 8$) BER/SER performance expressions. By using a high SNR approximation technique, the asymptotic performance metrics (i.e., ADO and ACG) are derived in Section 4.4.4. Section 4.4.5 presents some numerical results consisting of the average SNR gain variation and average BER/SER performances. The theoretical BER/SER results have been validated by Monte Carlo simulations for M -ary signals. Finally, Section 4.4.6 draws conclusions about the analytical and numerical results.

4.4.1. System Model and SNR Statistics for the Modified TAS/OSTBC Schemes

The diversity scheme proposed in this section introduces a generic baseline for modified TAS-aided OSTBC schemes. These modified schemes can be constructed by properly sharing the active radio-frequency chains that exist in the transmitter side such that some of them are switched on/activated at the end of an antenna selection process and the remaining are unquestioningly used as a part of the OSTBC codeword transmission.

Different sharing methods that are used for antenna selection/switching and OSTBC transmission sessions would bring along with differing advantages on feedback data-rate reduction and feedback-error resistance when compared to the conventional (optimal-antenna-subset-selecting) TAS/OSTBC scheme. For the efficiency it provides on feedback-rate reduction and its antenna selection/switching

simplicity, the derivations in this section have focused on the modified TAS/OSTBC scheme that employs single TAS among $n_T - n_S + 1$ transmit antennas where n_S is the number of activated transmit antennas. The OSTBC mapper inserts the message symbols into the first $n_S - 1$ rows of the codeword matrix $\mathcal{C}_{n_S \times N_{TS}}$ that correspond to the transmit antennas regardlessly allocated for signal transmission where N_{TS} denotes the number of OSTBC time slots. As the single transmit antenna that maximizes the output SNR is determined, the signal transmission of the last n_S^{th} row of the OSTBC codeword matrix is directed to the selected transmit antenna.

The proposed multi-antenna diversity scheme, that is constructed as a combination of the M-TAS/OSTBC transmitter and the receiver employing a MRC-like combining through n_R antennas as depicted in Figure 4.11 below, provides considerable feedback-rate reduction caused by the reduction in the number of possible antenna subsets considered in TAS process. In Figure 4.11, $w_i, i = 1, 2, \dots, n_R$

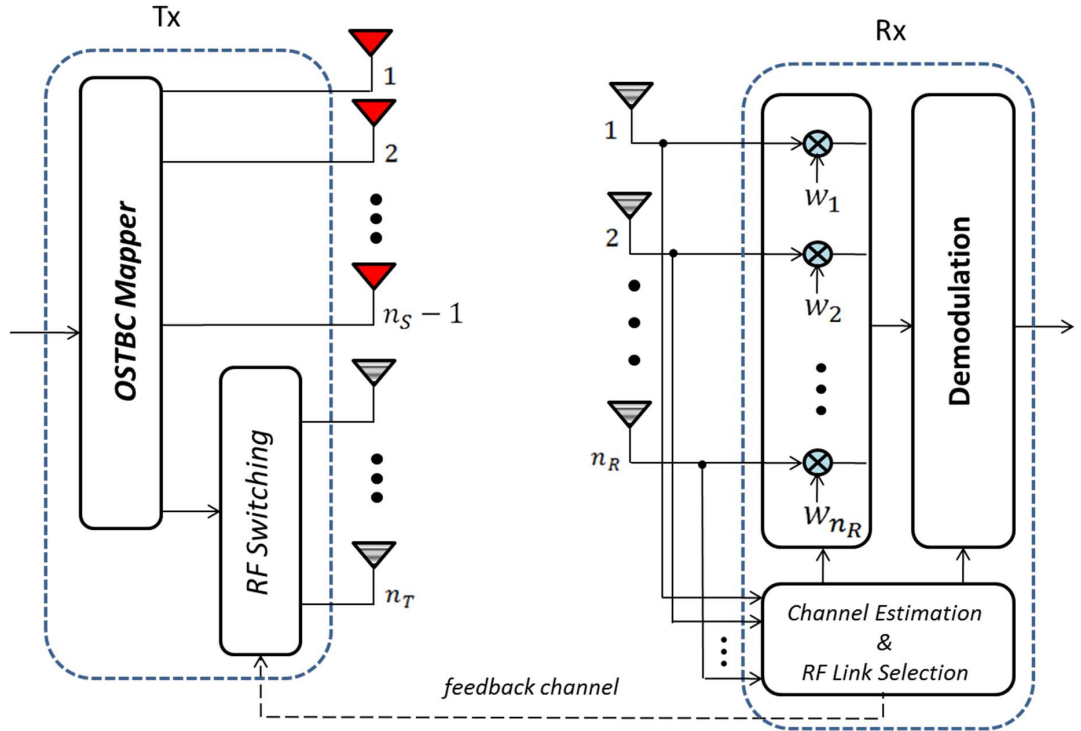


Figure 4.11: Block diagram of M-TAS/OSTBC scheme.

denote the optimal receive combining weights that are adaptively estimated in terms of the complex channel gains at the signal reception and processing period. For a multi-antenna setup of n_T transmit antennas that performs n_S -branch OSTBC transmission, the number of bits required for the feedback data would be reduced

to $k_{ms} = \lceil \log_2(n_T - n_S + 1) \rceil$ whereas this value was $k_{cs} = \lceil \log_2 \binom{n_T}{n_S} \rceil$ for the C-TAS/OSTBC scheme with optimal antenna subset selection. Here, $\lceil \cdot \rceil$ denotes the smallest integer that is greater than its argument. Figure 4.12 depicts the reduction in the number of feedback data bits, that is provided by employing M-TAS/OSTBC schemes instead of the conventional schemes. For the number of active transmit antennas $n_S \in \{2, 3, 4, 5\}$ and total transmit antennas satisfying $n_T > n_S$, the feedback rate gain of several M-TAS/OSTBC schemes when compared to C-TAS/OSTBC, that is easily evaluated as $\Delta k(n_T, n_S) = k_{cs}(n_T, n_S) - k_{ms}(n_T, n_S)$, is shown to be positive for $\forall n_T, n_S$ and to follow in general an increasing trend for increasing values of the difference $n_T - n_S$. As seen in Figure 4.12, by setting the first $n_S - 1$ transmit antennas

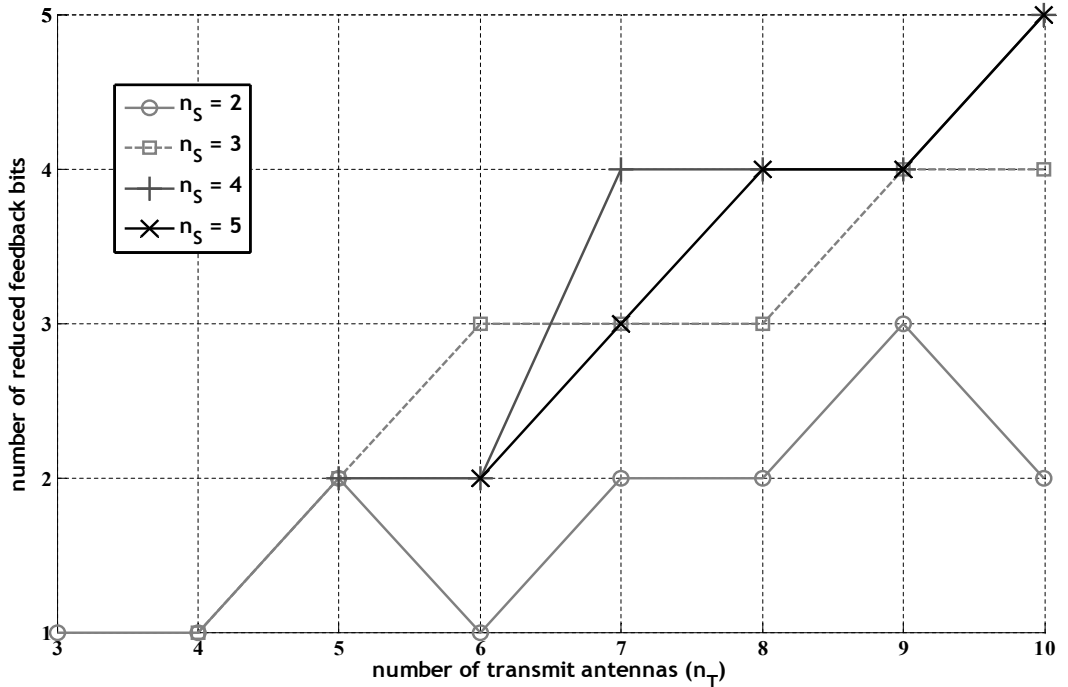


Figure 4.12: Feedback reduction gain of the M-TAS/OSTBC scheme over the C-TAS/OSTBC scheme for several values of n_S and n_T .

fixed for all transmission intervals and performing single TAS among $n_T - n_S + 1$ ones, M-TAS/OSTBC schemes reduce the feedback rate while providing full-diversity order transmission in ideal feedback cases and a robust error performance in the presence of FEs.

On the other hand, the receiver, considering that the channel coefficients $h_{i,j}$, $i = 1, 2, \dots, n_R$, $j = 1, 2, \dots, n_T$ are estimated perfectly, combines the independent signal replicas gathered from the j^{th} transmit antenna through n_R receive antennas during

the OSTBC codeword interval by using the optimal combiner (i.e., MRC) weights (i.e., w_i , $i = 1, 2, \dots, n_R$) and puts out the combined SNR: $\gamma_j = \frac{E_s}{N_0} \sum_{i=1}^{n_R} |h_{i,j}|^2$. While the envelopes $|h_{i,j}|$ are assumed to be i.i.d. Nakagami- m random variates with fading parameter m and squared mean $\Omega = E[|h_{i,j}|^2]$, γ_j can be easily shown to be a Gamma random variate with PDF and CDF given as $f(x) = (m/\bar{\gamma})^{mn_R} x^{mn_R-1} e^{-xm\bar{\gamma}} / \Gamma(mn_R)$, $x \geq 0$, and $F(x) = \psi(mn_R, xm\bar{\gamma}) / \Gamma(mn_R)$, $x \geq 0$, respectively. By comparing the instantaneous SNRs γ_j , $j = n_S, n_S + 1, \dots, n_T$, the receiver determines the transmit antenna that will be used for transmission: $\gamma_{max} = \max_{n_S \leq j \leq n_T} \{\gamma_j\}$. The index of the selected transmit antenna is fed to the transmitter back over a feedback link that is considered to have fading characteristics from a more realistic viewpoint [19], [50], [74]. Since the feedback channel is assumed to be reciprocal to the main communications channel, the fading parameter m is identical for both signaling directions. The k_{ms} -bit feedback information is sent to the transmitter followed by an L_{ms} -PSK modulation where $L_{ms} = 2^{k_{ms}}$. Due to degrading effects of the feedback channel with fading characteristics, the index information would be decoded erroneously. With the probability p_k , $k = 1, 2, \dots, n_T - n_S + 1$, the k^{th} best transmit antenna will be activated together with the fixed transmit antenna(s) while the others are kept silent. The a priori probabilities p_k can be easily defined as $p_k = \sum_{m=1}^{K_{ms}} p(s_m) P(s_m \rightarrow s_k)$, where $K_{ms} = n_T - n_S + 1$ and s_m denotes the baseband symbol belonging to L_{ms} -PSK modulation, $p(s_m)$ represents the occurrence probability of the symbol s_m and $P(s_m \rightarrow s_k)$ is the PEP related to the k^{th} and m^{th} symbols corresponding to the k^{th} and m^{th} best antennas. The PEP values have been evaluated for different modulation levels (L_{ms}) in the literature [19]. Based on feedback link assumption with fading and AWGN characteristics, the total error probability expression can be expressed as $P_s(\bar{\gamma}_{FC}) = \sum_{k=1}^{K_{ms}} p_k P_{s,k}(\bar{\gamma}_{FC})$. Here, $P_{s,k}(\bar{\gamma}_{FC})$ is the average SER related to the k^{th} best transmit antenna together with the fixed one(s) and $\bar{\gamma}_{FC}$ is the average SNR of the feedback link. The feedback message, delivering k_{ms} bits in order to represent $K_{ms} = n_T - n_S + 1$ transmit antennas, can be decoded to $L_{ms} = 2^{k_{ms}} \geq K_{ms}$ index values in the presence of FEs. This may cause the proper transmit antenna indices ($1, 2, \dots, K_{ms}$) to be de-mapped to one of the improper transmit antenna indices ($K_{ms} + 1, K_{ms} + 2, \dots, L_{ms}$) (which can be called FF). Although many re-transmission or error control coding protocols such as ARQ can be employed to overcome the FF, it is assumed that the transmitter

arbitrarily selects and activates one of the K_{ms} transmit antennas in order not to reduce the transmission rate that would be brought along with the usage of a re-signaling structure. Thus, assuming that each transmit antenna is selected equally-likely in the presence of FF as in [74], [84] the a priori probabilities would be increased by the factor $(1/K_{ms}) \sum_{j=K_{ms}+1}^{L_{ms}} p_j$ yielding $p'_k = p_k + \frac{1}{K_{ms}} \sum_{j=1+K_{ms}}^{L_{ms}} p_j, k = 1, 2, \dots, K_{ms}$. Hence, it would be possible to concern also the case of $L_{ms} > K_{ms}$ unlike [45] which considers $K_{ms} = 2^b, b \in \mathbb{Z}^+$. After switching the k^{th} best transmit antenna as a result of erroneous decoding, the resulted instantaneous post-processing SNR at the receive-combiner output will be $\gamma_{o,k} = \frac{1}{n_S} \left(\gamma_{max,k} + \sum_{j=1}^{n_S-1} \gamma_j \right)$ where $\gamma_{max,k}$ denotes the k^{th} maximum instantaneous SNR among $n_T - n_S + 1$ transmit branch SNRs and $\sum_{j=1}^{n_S-1} \gamma_j$ denotes the instantaneous SNR of the fixed transmit branches. The total transmit power is equally allocated to all active transmit antennas that yields the factor $\frac{1}{n_S}$.

By manipulating the PDF expression given in [74, (1)], the PDF expression of $\gamma_{max,k}$ can be obtained as

$$f_{(k)}(x) = a_0 \sum_{p=0}^{k-1} a_1 \sum_{q=0}^{n_T-n_S+1-k+p} a_2 \sum_{r=0}^{q(mn_R-1)} \frac{a_3 x^{r+mn_R-1}}{e^{xm(1+q)/\bar{\gamma}}} \quad (4.40)$$

Note that, (4.40) is obtained for arbitrary values of $m \geq 1/2$ that satisfy the product mn_R to be integer (i.e., $mn_R \in \mathbb{Z}^+$). In (4.40), the parameters are defined as $a_0 = (n_T - n_S + 1) \binom{n_T-n_S+1}{k-1} / \Gamma(mn_R) (m/\bar{\gamma})^{-mn_R}$, $a_1 = \binom{k-1}{p} (-1)^p$, $a_2 = \binom{n_T-n_S+1-k+p}{q} (-1)^q$ and $a_3 = \mu_r(q, mn_R)$ are the multinomial coefficients [62, (0.314)]. The MGF of the instantaneous output SNR $\gamma_{o,k}$ can be easily obtained via the product of MGFs of the instantaneous SNRs $\gamma_{max,k}$ and $\gamma_F = \sum_{j=1}^{n_S-1} \gamma_j$: $\mathcal{M}_{\gamma_{o,k}}(s) = \mathcal{M}_{\gamma_{max,k}}\left(\frac{s}{n_S}\right) \mathcal{M}_{\gamma_F}\left(\frac{s}{n_S}\right)$. By performing the MGF integral [19, (7.44)] over (4.40), $\mathcal{M}_{\gamma_{max,k}}\left(\frac{s}{n_S}\right)$ can be obtained as

$$\mathcal{M}_{\gamma_{max,k}}\left(\frac{s}{n_S}\right) = a_0 \sum_p a_1 \sum_q a_2 \sum_r \frac{\Gamma(r+mn_R)}{n_S^{-r-mn_R}} \frac{a_3}{[s + s_q]^{r+mn_R}} \quad (4.41)$$

where $s_q = (n_S m (q+1)) / \bar{\gamma}$. Also, the MGF of γ_F has been derived in [19, (6.65)] as $\mathcal{M}_{\gamma_F}\left(\frac{s}{n_S}\right) = (n_S m / \bar{\gamma})^{mn_R(n_S-1)} (s + \frac{n_S m}{\bar{\gamma}})^{-mn_R(n_S-1)}$. Thus, the MGF expression of

the output SNR can be written as

$$\mathcal{M}_{\gamma_{o,k}}(s) = \bar{a}_0 \sum_p a_1 \sum_q a_2 \sum_r \frac{a_3 a_4 [s + s_0]^{mn_R(n_S-1)}}{[s + s_q]^{r+mn_R}} \quad (4.42)$$

where $\bar{a}_0 = a_0 s_0^{mn_R(n_S-1)}$, $a_4 = \Gamma(r + mn_R) n_S^{r+mn_R}$ and $s_0 = n_S m / \bar{\gamma}$. In order to obtain a simplified and univariate expression of the CDF related to the output SNR, the inverse Laplace transform is applied over (4.42) as

$$F_{\gamma_{o,k}}(x) = \bar{a}_0 \sum_p a_1 \sum_q a_2 \sum_r a_3 a_4 \mathcal{L}^{-1} \left\{ \frac{[s + s_0]^{-c_0(n_S-1)}}{s[s + s_q]^{c_r}} \right\} \quad (4.43)$$

where $c_r = r + mn_R$. Inverse Laplace transform can be obtained by using [72, (2.1.4-8)] that is valid for $s_i \neq s_j \neq 0, \forall i \neq j$. In order to utilize this identity, the coefficients $s_k, k = 1, 2, 3, (s_1 = 0, s_2 = s_0, s_3 = s_q)$ are renamed as $\hat{s}_d, d = 1, \dots, N \leq 3$, such that all the coefficients ($\hat{s}_d$) are different from each other. The powers of equal coefficients will sum up and result in the new powers ($\hat{c}_d$) (i.e., $\hat{c}_1 = c_1 + c_2$ if $s_1 = s_2$). Thus, the CDF in (4.43) can be expressed as

$$F_{\gamma_{o,k}}(x) = \bar{a}_0 \sum_p a_1 \sum_q a_2 \sum_r a_3 a_4 \sum_{n=1}^N \sum_{l=1}^{\hat{c}_n} \frac{a_5 x^{\hat{c}_n-l}}{e^{x \hat{s}_n}} \quad (4.44)$$

where $a_5 = \frac{A_{nl}(-\hat{s}_n)}{(\hat{c}_n-l)!(l-1)!}$, $A_{nl}(\cdot)$ is defined in [72, (2.1.4-8)]. The derivation of (4.44) helps obtaining not only the exact error probability expressions but also the OP that would be handled in the following sections.

4.4.2. Outage Probability Analysis for the M-TAS/OSTBC Schemes

OP, which is defined as the probability that the instantaneous capacity is less than a given capacity (bandwidth efficiency) $\mathcal{R}$ (bit/s/Hertz) [51], is a useful performance metric since it comprises a realistic view on the system capacity. Considering the definition of the instantaneous capacity $C_{\gamma_{o,k}} = \log_2(1 + \gamma_{o,k})$, the OP can be written as $P_{out,k} = F_{\gamma_{o,k}}(2^{\mathcal{R}} - 1)$ [51], [70]. By the weighted sum of OPs related to each selection case (i.e., $P_{out,k}$) and using the a priori probabilities p'_k , the average OP can easily be evaluated: $P_{out} = \sum_{k=1}^{n_T-n_S+1} p'_k P_{out,k}$.

4.4.3. Exact Error Performance Analysis for the M-TAS/OSTBC Schemes

Based on the output SNR statistics provided in Section 4.4.1, the exact error performance expressions of the investigated M-TAS/OSTBC schemes are derived for erroneous feedback conditions and Nakagami- m fading channels. In order to provide a compact representation of different performance metrics such as MGF of the output SNR and the average BER/SER, the unified integral definitions given in [73] and [74], respectively as

$$\mathcal{J}_{\gamma_{o,k}}(\theta, \epsilon, \Phi) = \theta \int_0^\infty x^\epsilon e^{-x\Phi} F_{\gamma_{o,k}}(x) dx \quad (4.45)$$

and

$$\hat{\mathcal{J}}_{\gamma_{o,k}}(\theta, \Phi) = \theta \int_0^\infty e^{-x\Phi} F_{\gamma_{o,k}}(x) {}_1F_1\left(1; \frac{3}{2}; \frac{x\Phi}{2}\right) dx \quad (4.46)$$

are used where ${}_1F_1(\cdot, \cdot, \cdot)$ denotes the confluent hypergeometric function [62, (210.1)]. Setting different values in the parameters of both integrals given in (4.45) and (4.46) (i.e., θ, ϵ and Φ) will result in the MGF and the BER/SER of several binary/ M -ary modulations. By considering both (4.45) and (4.46), it is possible to evaluate the MGF of $\gamma_{o,k}$ by using the identity $\mathcal{M}_{\gamma_{o,k}}(s) = \mathcal{J}_{\gamma_{o,k}}(s, 0, s)$, BER of BPSK, CBFSK, NCBFSK, DBPSK, the approximate SER of M -PSK signals ($M \geq 8$) by using the identity $P_{s,k} = \mathcal{J}_{\gamma_{o,k}}\left(\frac{\lambda_3(\lambda_2^{\lambda_1})}{2\Gamma(\lambda_1)}, \lambda_1 - 1, \lambda_2\right)$, SER of M -PAM signals by using the identity $P_{s,k} = \mathcal{J}_{\gamma_{o,k}}\left(\sqrt{\frac{3(M-1)}{\pi M^2(M+1)}}, \frac{-1}{2}, \frac{3}{(M^2-1)}\right)$ and SER of QPSK and M -QAM signals by using the identity $P_{s,k} = \mathcal{J}_{\gamma_{o,k}}\left(\sqrt{\frac{\lambda_4}{8\pi}}(\lambda_5 - \lambda_6), \frac{-1}{2}, \frac{\lambda_4}{2}\right) + \hat{\mathcal{J}}_{\gamma_{o,k}}\left(\frac{\lambda_4\lambda_6}{2\pi}, \lambda_4\right)$. The modulation-dependent parameters for binary and M -ary cases are listed in Table 3.1.

Thus, deriving the exact expressions of (4.45) and (4.46) will lead us to obtain and examine the performance metrics of the M-TAS/OSTBC schemes for several modulations. By substituting (4.44) into the unified integrals in (4.45) and (4.46), and rearranging the resulting integrals in terms of the Gamma function and the Gauss hypergeometric function ${}_2F_1$ [62, (3.197-3)], respectively, the unified performance expressions can be obtained as [87]

$$\mathcal{J}_{\gamma_{o,k}}(\theta, \epsilon, \Phi) = \theta \bar{a}_0 \sum_p a_1 \sum_q a_2 \sum_r a_3 a_4 \sum_n \sum_l a_5 \frac{\Gamma(\epsilon + 1 + \hat{c}_n - l)}{(\Phi + \hat{s}_n)^{\epsilon+1+\hat{c}_n-l}} \quad (4.47)$$

and

$$\begin{aligned} \hat{\mathcal{J}}_{\gamma_{o,k}}(\theta, \epsilon, \Phi) &= \theta \bar{a}_0 \sum_p a_1 \sum_q a_2 \sum_r a_3 a_4 \sum_n \sum_l a_5 \\ &\times \frac{{}_2F_1\left(1; 2 + \hat{c}_n - l; \frac{3}{2}; \frac{\hat{s}_n}{\Phi + \hat{s}_n}\right) \Gamma(1 + \hat{c}_n - l)}{(\Phi + \hat{s}_n)^{1+\hat{c}_n-l}} \end{aligned} \quad (4.48)$$

As in the OP analysis, the average error probability performance of the M-TAS/OSTBC schemes can be obtained by simply averaging over all possible transmit antenna combinations. Note that ${}_2F_1$ is available in well known software programs (e.g., MATHEMATICA and MAPLE). Also, substituting $(\theta, \epsilon, \Phi) = (s, 0, s)$ into (4.47) provides an alternative representation of the MGF expression given in (4.42).

4.4.4. Asymptotic Diversity and Coding Gain Analysis for the M-TAS/OSTBC Schemes

The asymptotic performance analysis, by supporting and strengthening the outcomes and insights that the exact performance analysis has provided, constitutes a useful tool for examining the high SNR behavior of the proposed M-TAS/OSTBC schemes. This section provides the derivations related to the ADO and the ACG that yield to examine the relative enhancements due to C-TAS/OSTBC schemes.

In order to obtain the asymptotic performance expressions, the high-SNR approximation technique proposed in [69] is used together with the asymptotic expressions of the PDF and CDF of envelopes $|h_{i,j}|$, respectively as

$$f(x) \approx \left(\frac{m}{\bar{\gamma}}\right)^{mn_R} \frac{x^{mn_R-1}}{\Gamma(mn_R)} \quad (4.49)$$

and

$$F(x) \approx \left(\frac{m}{\bar{\gamma}}\right)^{mn_R} \frac{x^{mn_R}}{\Gamma(1 + mn_R)} \quad (4.50)$$

Repeating the analytical derivations provided for the exact SNR statistics in Section 4.4.1 by substituting (4.49) and (4.50) into the related parts of the derivation steps yields to obtain the asymptotic expression of the MGF given in (4.42) as

$$\mathcal{M}_{\gamma_{o,k}}(s) \approx \frac{mn_R(n_T - n_S + 1)\Gamma(mn_R(n_T - n_S + 2 - k))}{\left\{ \left(\frac{n_S m}{\bar{\gamma}} \right)^{-mn_R} \Gamma(1 + mn_R) \right\}^{n_T - n_S + 2 - k}} \times \binom{n_T - n_S}{k - 1} s^{-mn_R(n_T + 1 - k)} \quad (4.51)$$

By simply applying the inverse Laplace transform over (4.51) results in the asymptotic PDF expression related to the case of the selection of k^{th} best transmit antenna [87]:

$$f_{\gamma_{o,k}}(x) \approx \frac{mn_R(n_T - n_S + 1)\Gamma(mn_R(n_T - n_S + 2 - k))}{\{(n_S m)^{-mn_R} \Gamma(1 + mn_R)\}^{n_T - n_S + 2 - k}} \times \binom{n_T - n_S}{k - 1} \frac{x^{mn_R(n_T + 1 - k) - 1}}{\Gamma(mn_R(n_T + 1 - k))} \quad (4.52)$$

Finally, through the decomposition viewpoint provided in [69], the ADO and ACG of M-TAS/OSTBC schemes can easily be extracted from the asymptotic PDF in (4.52) as

$$G_{d,ms}^{(k)} = mn_R(n_T + 1 - k) \quad (4.53)$$

and

$$G_{c,ms}^{(k)} = p'_{k,ms} n_M \left\{ \frac{2^{mn_R(n_T + 1 - k) - 1} \Gamma(mn_R(n_T + 1 - k) + 0.5)}{\sqrt{\pi} \{(n_S m)^{-mn_R} \Gamma(1 + mn_R)\}^{n_T - n_S + 2 - k}} \times \binom{n_T - n_S}{k - 1} \frac{mn_R(n_T - n_S + 1)\Gamma(mn_R(n_T - n_S + 2 - k))}{\Gamma(mn_R(n_T + 1 - k))^2 \Gamma(mn_R)^{n_T - n_S}} \right\}^{-\frac{1}{mn_R(n_T + 1 - k)}} \quad (4.54)$$

respectively. Here, n_M is the modulation-dependent parameter and $p'_{k,ms}$ is the a-priori probability related to the selection of k^{th} best transmit antenna selection. Since (4.53) and (4.54) are dependent to the ordinal number k , the average asymptotic statistics would be obtained by averaging over all possible values of k and the a-priori

probabilities $p'_{k,ms}$ as in the OP and exact error probability analyses. However, considering the fact that the term corresponding to the lowest ADO (i.e., minimizing (4.53)) would dominate the overall asymptotic performance, the averaged values of ADO and ACG of M-TAS/OSTBC schemes, can be represented by $G_{d,ms}^\infty = mn_R n_S$ and $G_{c,ms}^\infty = \bar{\mathcal{P}}_{ms} n_M \left\{ \frac{2^{mn_R n_S - 1} \Gamma(mn_R n_S + 0.5) (n_T - n_S + 1)}{\sqrt{\pi} \Gamma(mn_R n_S)^2 \Gamma(mn_R)^{n_T - n_S}} (n_S m)^{mn_R} \right\}^{-\frac{1}{mn_R n_S}}$ that correspond to the dominant ordinal $k^* = n_T - n_S + 1$. The contribution of the a-priori probabilities that depend on the average feedback SNR $\bar{\gamma}_{FC}$ is applied by the term $\bar{\mathcal{P}}_{ms}$ that is simply the mean of the a-priori probabilities $p'_{k,ms}$, $k = 1, 2, \dots, K_{ms}$.

Similarly, for C-TAS/OSTBC scheme, by using the identity provided in [84, (17)], the ADO is shown to be $G_{d,cs}^\infty = mn_R n_S$ that is identical to that of M-TAS/OSTBC scheme whereas the ACG is obtained as $G_{c,cs}^\infty = \bar{\mathcal{P}}_{cs} n_M \left\{ \frac{2^{mn_R n_S - 1} \Gamma(mn_R n_S + 0.5) mn_R n_S}{\sqrt{\pi} \Gamma(mn_R n_S) \Gamma(1 + mn_R)^{n_S}} \binom{n_T}{n_S} \right\}^{-\frac{1}{mn_R n_S}}$ where $\bar{\mathcal{P}}_{cs}$ is the mean value of the a-priori probabilities $p'_{k,cs}$, $k = 1, 2, \dots, K_{cs} = n_T - n_S + 1$. Thus, the asymptotic results clearly show that both modified and conventional schemes achieve the ADO of $mn_R n_S$ in the presence of feedback impairments. Besides, it would be straightforward to obtain the asymptotic BER/SER expressions using the identity given in [69, (1)]. The ratio of the average ACG of the M-TAS/OSTBC scheme (i.e., $G_{c,ms}^\infty$) to that of the C-TAS/OSTBC scheme (i.e., $G_{c,cs}^\infty$) would simply help deriving the average SNR gain (in dB scale) obtained by employing the modified scheme instead of the conventional one [87]:

$$\begin{aligned} \eta &= 10 \log_{10} \left(\frac{G_{c,ms}^\infty}{G_{c,cs}^\infty} \right) \\ &= 10 \log_{10} \left(\frac{\bar{\mathcal{P}}_{ms}}{\bar{\mathcal{P}}_{cs}} \left\{ \frac{n_S (mn_S)^{mn_R} \Gamma(1 + mn_R)^{n_S}}{\Gamma(1 + mn_R n_S) \binom{n_T}{n_S - 1} \Gamma(mn_R)^{n_T - n_S}} \right\}^{-\frac{1}{mn_R n_S}} \right) \end{aligned} \quad (4.55)$$

that represents the relative ACG of the modified scheme to the conventional scheme under erroneous feedback channel conditions especially in the high SNR region.

4.4.5. Numerical Results for the Performance of M-TAS/OSTBC Schemes in the Presence of FEs

This section presents numerical results consisting of both theoretical performance results and Monte Carlo simulation results of the proposed and

conventional TAS/OSTBC schemes under erroneous feedback conditions (i.e., FEs). In order to examine the practical (real-world) performances of the M-TAS/OSTBC and C-TAS/OSTBC schemes and to highlight the robustness of the modified schemes to FEs, this subsection have provided miscellaneous performance results consisting of comparisons between the modified scheme and the conventional scheme for several system and channel configurations. Considering that each diversity branch has unit channel power (i.e., $\Omega = 1$) and the average energy per symbol is unity (i.e., $E_s = 1$), performance demonstration is made in Figures 4.13-4.15 by exemplifying the most popular two-branch and three-branch OSTBCs (G_2 -OSTBC and G_3 -OSTBC, respectively). Note that, in order to evaluate the average exact theoretical BER/SER results for C-TAS/OSTBC schemes, the exact derivations provided in [84, (11), (12)] have been used.

In Figure 4.13, a surface plot, that can be easily evaluated via (4.55), is presented to depict the average SNR gain of M-TAS/ G_3 -OSTBC scheme over C-TAS/ G_3 -OSTBC scheme for $m = 2$, $\bar{\gamma}_{FC} = 0$ dB and several n_T and n_R . In case of a feedback

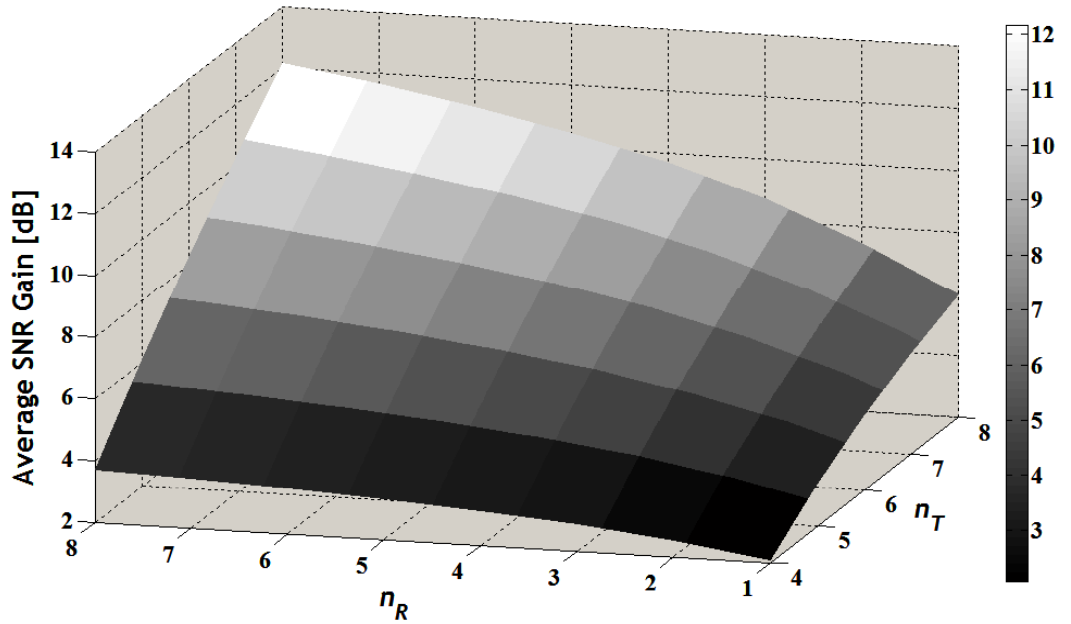


Figure 4.13: Average SNR gain of the M-TAS/ G_3 -OSTBC scheme over the C-TAS/ G_3 -OSTBC scheme for $m = 2$, $\bar{\gamma}_{FC} = 0$ dB, and several values of n_R and n_T .

channel with average SNR of $\bar{\gamma}_{FC} = 0$ dB and a main transmission channel with Nakagami fading characteristics with the parameter $m = 2$, the surface plot shows the average SNR gain achieved by employing M-TAS/ G_3 -OSTBC scheme instead of

C-TAS/G₃-OSTBC scheme. By examining the variation of the surface plot given for different values of n_T and n_R , one can realize that for a given value of n_R , the average SNR gain follows a decisively increasing trend due to increasing values of n_T . Besides, it is also important to see that the slope of the SNR gain curve due to the increasing values of n_T becomes greater with increasing values of n_R that provides the SNR advantage of M-TAS/OSTBC scheme to be dominant for greater values of n_R . Finally, it is easily seen that the modified scheme achieves considerable SNR gains over the conventional scheme in the presence of FEs. Throughout the considered values of n_T and n_R , the surface is seen to be all positive-valued that clearly mentions the robustness and superiority of M-TAS/OSTBC schemes for the entire grid. Also, the maximum SNR gain encountered through the considered n_T - n_R grid is obtained for the specific scenario of $n_T = 8$ and $n_R = 8$ as approximately 12 dB. Note that however, the SNR gain values are derived by using approximate performance metrics, a different asymptotic approach might result in slightly different values from the depicted ones but the trend of variation due to n_R and n_T would be the same.

Figure 4.14 depicts the average BER performance of both the modified and conventional TAS/OSTBC schemes for BPSK signals, $n_T = 6$, $n_R = 2$, $m = 0.5$ (i.e., One-sided Gaussian fading case), and it provides not only to examine the performance variation of both schemes due to differing values of $\bar{\gamma}_{FC} \in \{-5, 5, \infty\}$, but also to make comparisons between their performances. As seen from the BER curves related to schemes employing three-branch OSTBC transmission, both schemes face performance degradations as the average SNR of the feedback link decreases from $\bar{\gamma}_{FC} \rightarrow \infty$ (ideal feedback case) to $\bar{\gamma}_{FC} = -5$ dB. The BER curves related to the ideal feedback case shows that the conventional TAS/OSTBC scheme is superior to the modified scheme. With the help of the coding gain achieved by the optimal transmit antenna subset selection, the conventional scheme provides an SNR gain of approximately 0.6 dB at a BER of 10^{-5} . Whereas, in the presence of feedback impairments (i.e., FEs), the modified scheme achieves considerable SNR advantages when compared to the conventional one. For the average SNR values $\bar{\gamma}_{FC} = -5$ dB and $\bar{\gamma}_{FC} = 5$ dB, M-TAS/G₃-OSTBC scheme provides average SNR gains of approximately 0.95 dB and 1.75 dB for a BER of 10^{-5} , respectively when employed instead of the conventional scheme. Note that, by examining the slope of the BER curves of the ideal-feedback cases, it is easily noticed that both the modified and

conventional TAS/OSTBC schemes achieve full diversity orders (i.e., $mn_Rn_T = 6$) whereas in the presence of FEs, the ADO of both schemes would be lower-bounded by the ADO of G_3 -OSTBC (w/o TAS) scheme (i.e., $mn_Rn_S = 3$). Hence, the variation of the ADO in the absence and the presence of FEs is consistent with the outcomes of the asymptotic analysis provided in Section V. Also, the simulation results are all seen to be in perfect agreement with the theoretical ones. Figure 4.15 exposes the

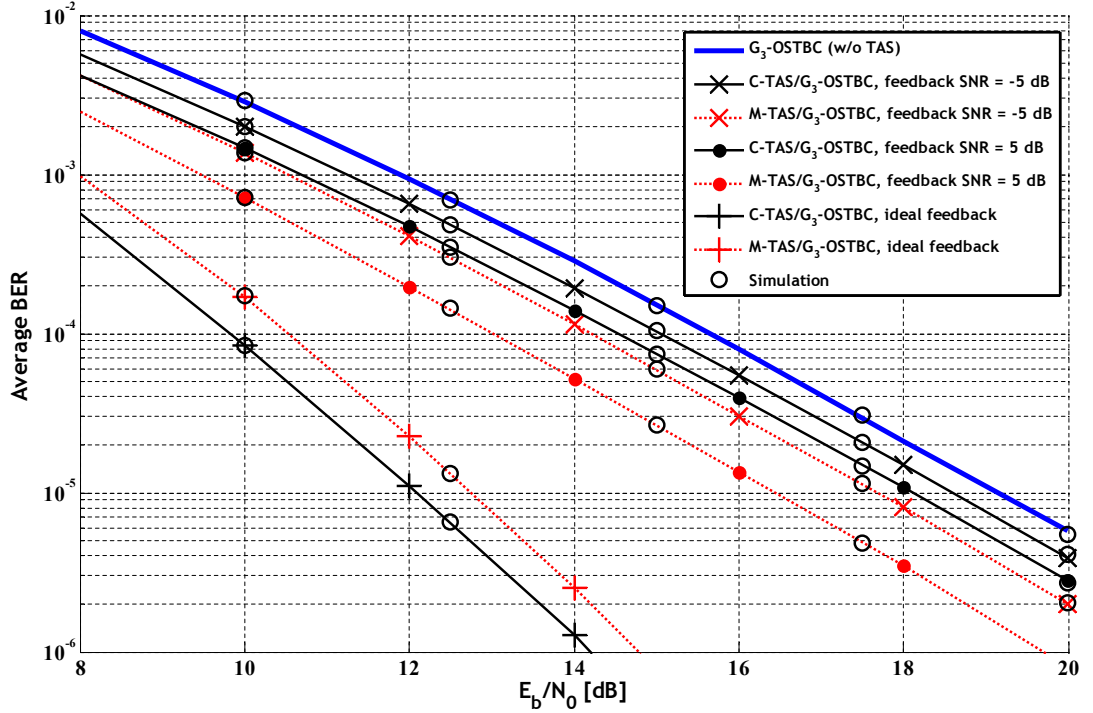


Figure 4.14: Average BER comparison of the M-TAS/ G_3 -OSTBC scheme to the C-TAS/ G_3 -OSTBC scheme for BPSK signals, $n_T = 6$, $n_R = 2$ and $m = 0.5$ (One-sided Gaussian fading case) and $\bar{\gamma}_{FC,dB} \in \{-5, 5, \infty\}$.

average SER performances of M-TAS/ G_2 -OSTBC and C-TAS/ G_2 -OSTBC schemes for 16-QAM signals, $n_T = 5$, $\bar{\gamma}_{FC} = 0$ dB and different values of m and n_R . As clearly seen from the depicted curves, for each scenario constructed by different $(n_R - m)$ pairs, the average SER curves that are related to the practical feedback conditions of both schemes are lower and upper bounded by the curves of pure G_2 -OSTBC (w/o TAS) scenario and C-TAS/ G_2 -OSTBC scenario which considers ideal feedback conditions, respectively. Similar to the outcomes of Figure 4.14, the ADO of the hybrid TAS/OSTBC schemes is seen to converge to that of the pure OSTBC (w/o TAS) scheme in the presence of FEs, as also stated as a result of the ADO analysis. Moreover, employing M-TAS/ G_2 -OSTBC scheme instead of C-TAS/ G_2 -OSTBC scheme reduces

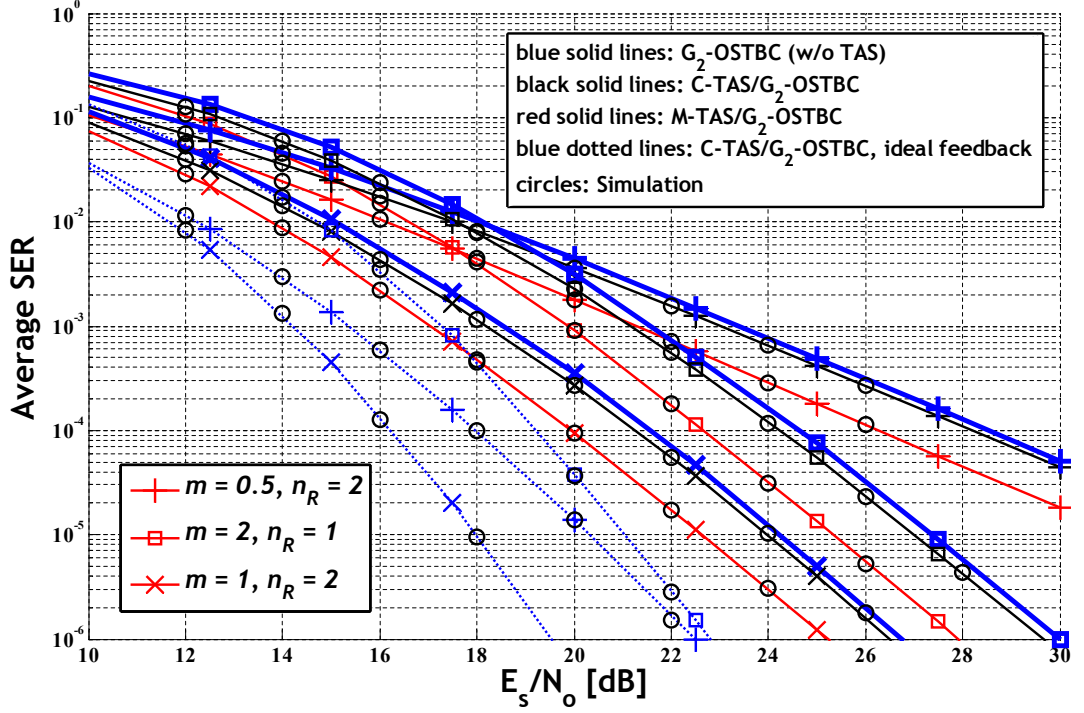


Figure 4.15: Average SER comparison of the M-TAS/ G_2 -OSTBC scheme to the C-TAS/ G_2 -OSTBC scheme for 16-QAM signals, $n_T = 5$, $\bar{\gamma}_{FC} = 0$ dB, and several values of n_R and m .

the SNR requirement to assure a SER of 10^{-4} by approximately 2 dB, 1.9 dB and 1.3 dB, respectively for three different scenarios with the configuration parameters $\{m = 0.5, n_R = 2\}$, $\{m = 2, n_R = 1\}$ and $\{m = 1, n_R = 2\}$.

4.4.6. Conclusions for the Effects of FEs on the Performance of the M-TAS/OSTBC Schemes

This section has introduced an feedback-rate efficient and FE-resistant TAS/OSTBC scheme employing a modified transmit antenna selection and switching strategy that brings along with considerable reduction in the feedback burden of the closed-loop diversity scheme and in the SNR requirement corresponding to any desired SER values. The analytical derivations and the numerical results presented within this section have exposed the robustness of the modified scheme to practical (erroneous-feedback) conditions when compared to the conventional scheme. Considering a generic fading model like Nakagami- m , statistics related to the output SNR and the performance metrics have been presented for ideal feedback case and practical impairments. SER curves have shown that both the modified and

conventional schemes achieve full diversity order (mn_Rn_T) for ideal cases, while in the presence of FEs they maintain the ADO of mn_Rn_S that is provided by the pure OSTBC (w/o TAS) transmission itself and the optimal-diversity reception provided by the MRC-like structure. Analytical and numerical performance results clearly point out the superiority of M-TAS/OSTBC scheme to C-TAS/OSTBC scheme in the presence of FEs. This superiority is provided by the compromise between the advantage of feedback rate reduction and the gain based on antenna subset selection strategy. The number of feedback bits per each index transmission limits the number of imperfect TAS cases under erroneous feedback conditions that directly degrades the error performance. Besides, the antenna selection strategy has a prepotent role on the system performance depending on the selection efficiency of most qualified transmit branches. Moreover, the numerical results have provided another important insight about the relative ACG achieved by the M-TAS/OSTBC scheme which is shown to follow an increasing trend due to the increasing values of n_T and an almost increasing trend due to the increasing values of n_R . The incremental variation of the relative ACG is based on the advantage achieved by the reduction of feedback rate which results in reduction on the modulation level (e.g., 4 (QPSK) to 2 (BPSK)) of the feedback link transmission (accordingly enhancements in the BER/SER performance of the feedback link for any $\bar{\gamma}_{FC}$). Thus, the increase in the SNR gain gathered by M-TAS/OSTBC scheme for increasing values of n_T is highly related to the increase in the reduction of feedback burden with increasing n_T as depicted in Figure 4.12. Also, since the SNR gain gathered by the optimal receive combining is almost independent from the variation of n_T and n_S , an increase in the value of n_R stretches and amplifies the SNR difference that have had obtained for a given n_T . Thus, the act of the transmit selection&switching strategy is seen to be strengthened by the increased number of optimal receive combining chains.

Consequently, the considerable feedback-rate reduction gain and the average SNR gain achieved against the conventional scheme in the presence of FEs, provide this scheme to maintain significant importance in real-world wireless communications system design.

5. CONCLUSIONS

Within the investigation provided in this thesis, the performance analyses of HDSs that are constructed by combining TAS and/or RAS techniques with other combining and beamforming schemes are examined under the effects of practical impairments such as CEE, FD and FE. By focusing on the derivation of performance metrics of HDSs in the presence of practical imperfections, the robustness and the fragility of different diversity schemes have been exposed which brings along with a useful interpretation of the feasibility and suitability of the investigated WCSs to real-world communications applications. Hence, this thesis has presented an extensive and a more realistic examination on the actual performance of many well-investigated and approved diversity schemes in erroneous channel estimation and feedback conditions. Moreover, after analyzing the differing characteristics of several HDSs in different type of practical imperfections, this thesis has proposed modified versions of TAS/OSTBC schemes which are more robust and superior than the conventional (not modified) ones in the presence of FEs. By contributing a viewpoint to construct modified HDSs that could be shown to have higher immunity to the presence of practical impairments than the conventional ones, this thesis paves the way for the designation of alternative diversity schemes for real-world wireless communications applications and reveal great amount of information about the further WCS proposals.

Within the thesis' consideration, the exact and asymptotic expressions of performance metrics such as the PDF, CDF, MGF, n^{th} order moments of the output SNR, ergodic capacity, average OP, average BERs/SERs that provide useful insights for the performance interpretation of WCSs have been derived analytically, also validated by computer simulations based on the well-known Monte Carlo method.

The performance examinations consisting of system models and derivations followed by the general definitions and descriptions related to the investigated diversity schemes and the practical impairments have been given in detail in successive chapters of the thesis.

The background information that is necessary to interpret the analyses' results presented as the outcomes of the thesis is provided in Chapter 2. The characteristics

of the wireless propagation channel, the fading effect and the counter-measures that are proposed to combat with it, and the basic descriptions of the practical channel estimation and feedback impairments (i.e., CEE, FD and FE) are introduced briefly.

In the third chapter of the thesis, which has focused on the performance investigation of mainly four HDSs; TAS/OSTBC, TAS/MRT, MRT&RAS and JTRAS in the presence of both CEEs and FD over Nakagami- m fading channels, some important results have been gathered and some worthy breakthroughs have been made which can be listed as:

- As indicated in Section 3.2 and Section 3.3, the OP and BER/SER performances can easily achieve great improvements due to higher diversity orders yielded by the increase in the number of transmit antennas and/or receive antennas available for diversity transmission/reception.
- With the help of ADO analysis provided in Section 3.3.7, it has been shown that the unified scheme covering four HDSs (TAS/OSTBC, TAS/MRT, MRT&RAS and JTRAS) achieves an ADO of N which is shown to be unity for the first two (MISO) schemes (i.e., TAS/OSTBC and TAS/MRT) and $N = n_R$ for the latter two (MIMO) schemes (i.e., MRT&RAS and JTRAS).
- For the erroneous cases, the increase in the number of transmit and/or receive antennas provides SER performance enhancements in the low SNR region. All the performance results indicate that each HDS more or less faces performance degradations caused by CEE and FD phenomena.
- Numerical results of both sections show that, for reasonable values of the parameters of practical existence such as f_D , N_d and N_{fr} , HDSs still achieve considerable outage and error probability performances (with tolerable SNR degradations). It is important to state that each HDS still provides performance improvements when compared to its non-hybrid version in an SNR region restricted by a critical SNR value as mentioned in Numerical Results section. Beyond these critical SNR values, the benefits gathered by the transmit diversity of HDSs vanish in the presence of practical imperfections while only receive diversity persists as the ADO analysis has verified.

- Also, the performance comparisons that have been made between the first two HDSs (i.e., TAS/OSTBC and TAS/MRT schemes) and the latter two HDSs (i.e., MRT&RAS and JTRAS schemes) separately have provided insights about the robustness of two groups of HDSs to the variation of the number of pilot symbols (N_d) and the maximum Doppler frequency shift (f_D), respectively. The comparisons have shown that TAS/MRT and MRT&RAS schemes have been more robust than TAS/OSTBC and JTRAS schemes, respectively in the presence of both CEEs and FD.
- As consistent to the generic results of the unified scheme considered in Section 3.3, the specific results of Section 3.2 (for only MRT&RAS scheme) also show the maintaining of the receive diversity when speaking about MRT&RAS scheme in the presence of CEEs and FD.
- The performance results given in Section 3.2 clearly point out the superiority of MRT&RAS scheme to other diversity techniques such as single TAS, TAS/OSTBC, TAS/MRT, JTRAS and pure MRT schemes that employ equal number of total antennas. This superiority is stated to be provided by the virtue of full-CSI feedback to the transmitter however this CSI requirement feature tends the MRT&RAS scheme to be accorded especially with downlink communications systems (since they can tolerate complexity at the transmitter side).

In the fourth chapter of the thesis, the performance examination of TAS/OSTBC-type HDSs in the presence of FE has been presented. Under the effects of FEs in a feedback channel that has been modeled as BSC, the performance metrics related to hybrid TAS/OSTBC&MRC and JTRAS/OSTBC schemes are exposed in the first analysis part of this chapter (Section 4.2). Sequentially, by using a realistic feedback channel model that has fading characteristics as the main transmission link, the second and third parts make HDS proposals that are constructed by the modification of the transmission strategies of TAS/OSTBC&MRC schemes. Both Section 4.3 and Section 4.4 of this chapter expose the robustness and superiority of the proposed schemes in the presence of FEs when compared to the conventional TAS/OSTBC&MRC schemes by also deriving the performance metrics related to the erroneous feedback conditions of them.

Focusing on the practical feedback imperfections (i.e., FEs) in Nakagami- m fading channels, this chapter has provided detailed analyses on the exact expressions related to the statistics of the output SNR (PDF, CDF and MGF) and the capacity and error performances (OP and BER/SER) for both schemes. Some important results that have been gathered and some worthy breakthroughs that have been made throughout the performance investigation of Chapter 4 can be listed as:

- The ADO analysis of the unified scheme considered within Section 4.2 has shown that both TAS/OSTBC&MRC and JTRAS/OSTBC schemes achieve full diversity order (i.e., mn_Rn_T) for ideal conditions, while in the presence of FEs they keep maintaining the ADO provided by the OSTBC scheme (with no TAS) itself and the receive diversity provided by MRC-like structure and RAS, respectively (i.e., mn_Rn_S).
- The robustness that is gathered on the ADO by the combined TAS&OSTBC structure employed at the transmitters of both TAS/OSTBC and JTRAS/OSTBC schemes exposes the importance of these schemes in practical system scenarios when compared to the single TAS scheme that is shown in [45] to lose the transmitter diversity and to face severe deteriorations in the presence of FEs.
- As indicated in the numerical results of Section 4.2, for the ideal and erroneous conditions of both schemes, the OP and BER/SER performances can easily achieve considerable enhancements brought along with higher diversity orders provided by the increment in the number of transmit and receive antennas for BSC BER values that are greater than a specific threshold. Also, this threshold is seen to be greater for TAS/OSTBC scheme than that of JTRAS/OSTBC scheme which clearly point out the superiority of TAS/OSTBC scheme to JTRAS/OSTBC scheme especially in the presence of FEs. This superiority is provided by the virtue of multiple RF chains employed in the receiver side of TAS/OSTBC. However the simple receiver structure of JTRAS/OSTBC (employing only a single RF chain) tends this scheme to be a more feasible technique for multi-antenna communications systems rather than TAS/OSTBC scheme.
- For reasonable average BSC BER values specified for the feedback link (such as $p_e = 0.01$), both schemes still achieve considerable performances (with tolerable

SNR degradations) that provide both schemes to maintain significant importance in real-world wireless communications system design.

- The TAS/OSTBC-type HDS proposals in Section 4.3 and Section 4.4 are shown to perform better than the conventional TAS/OSTBC schemes in the presence of FEs by also providing reduced feedback burden.
- Similar to the asymptotic performance results presented in Section 4.2, SER curves related to the modified TAS/OSTBC (M-TAS/OSTBC) schemes have shown that both the modified and conventional TAS/OSTBC schemes achieve full diversity order (mn_Rn_T) for ideal cases, while in the presence of FEs they maintain the ADO of mn_Rn_S that is provided by the pure OSTBC scheme itself (w/o TAS) and the receive diversity provided by the MRC-like structure.
- Also, the specific results given in Section 4.3 for the M-TAS/Alamouti-OSTBC scheme clearly point out the superiority of this scheme to the conventional TAS/Alamouti-OSTBC and the other reduced-feedback-rate TAS/Alamouti-OSTBC schemes (S-1, S-2 and S-3 of [78]) in the presence of FEs. This superiority is provided by the compromise between the advantage of feedback rate reduction and the gain based on antenna subset selection strategy. The number of feedback bits per each index transmission limits the number of imperfect TAS cases under erroneous feedback conditions that directly degrades the error performance. Besides, the antenna selection strategy has a prepotent role on the system performance depending on the selection efficiency of most qualified transmit branches.
- Also, for higher values of $\bar{\gamma}_{FC,dB}$, M-TAS/Alamouti-OSTBC scheme still achieves considerable error performance that, together with the reduced feedback link traffic, provides this scheme to maintain significant importance in real-world wireless communications system design.
- The numerical results presented in Section 4.4 have provided another important insight about the relative ACG achieved by the M-TAS/OSTBC scheme which is shown to follow an increasing trend due to the increasing values of n_T and an almost increasing trend due to the increasing values of n_R . The incremental variation of the relative ACG is based on the advantage achieved by the reduction

of feedback rate which results in reduction on the modulation level (e.g., 4 (QPSK) to 2 (BPSK)) of the feedback link transmission (accordingly enhancements in the BER/SER performance of the feedback link for any $\bar{\gamma}_{FC}$). Thus, the increase in the SNR gain gathered by M-TAS/OSTBC scheme for increasing values of n_T is highly related to the increase in the reduction of feedback burden with increasing n_T as depicted in Figure 4.12. Also, since the SNR gain gathered by the optimal receive combining is almost independent from the variation of n_T and n_S , an increase in the value of n_R stretches and amplifies the SNR difference that has been obtained for a given n_T . Thus, the act of the transmit selection&switching strategy is seen to be strengthened by the increased number of optimal receive combining chains.

Consequently, the extensive examination of these HDSs has provided a framework on the inspection of practical performance, feasibility, superiority, robustness and importance in physical deployment scenarios of MIMO diversity schemes in real-world wireless communications. The extensive work included in this thesis provides guidance and encouragement for the investigation of other practical impairments' effect on WCSs.

As the extensive performance investigation provided in this thesis is being concluded, it should be clearly realized that the material documented throughout the successive chapters of this thesis not only makes contributions to the current literature by providing accurate solutions for many different diversity schemes but also aims to point out the crucial need for further performance investigations on practical impairments. The literature should include detailed examinations of other type of practical impairments such as FO, CCI as mentioned in the Introduction section and some other system-specific challenges such as FQE in order to compare different types of feedback quantization methods. Moreover, the performance analyses of the investigated and pointed out practical imperfections in multi-hop or cooperative-based relay networks are worth dealing [88]-[90], especially for generic wireless channel models such as Generalized K that would also provide insights about the composite fading-shadowing characteristics.

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CURRICULUM VITAE

Ahmet Faruk Coşkun, born in Kadıköy-İstanbul in 1984, has graduated from Kadıköy Anatolian High School in 2003. After matriculating to the Electronics Engineering Department of Gebze Institute of Technology (GIT) in 2005, he received the B.S. and M.S. degrees in Electronics Engineering in 2008 and 2010, respectively, from GIT, Gebze, Turkey, where he is currently working toward the Ph.D. degree with the same department. He has worked as a research assistant at the Telecommunications Section of the Electronics Engineering Department of GIT between 2008 and 2011. He has been currently working as a researcher in The Scientific and Technological Research Council of Turkey since 2011. His research interests include wireless communications, multiple-input multiple-output systems, and diversity techniques.

APPENDICES

A1. PUBLICATIONS BASED ON THE THESIS

International Journal Papers

1. Coşkun A. F., Kucur O., (2012), "Performance analysis of maximal-ratio transmission/receive antenna selection in Nakagami-m fading channels with channel estimation errors and feedback delay", IEEE Trans. on Veh. Tech., 61 (3), 1099-1108.
2. Coşkun A. F., Kucur O., (2013), "Feedback-rate efficient transmit antenna selection/Alamouti scheme with robust error performance in the presence of feedback errors", IEEE Communications Letters, 17 (5), 908-911.
3. Coşkun A. F., Kucur O., (2013), "Analysis of Transmit Antenna Selection/Space-Time Block Coding with Receive Selection and Combining over Nakagami-m Fading Channels in the Presence of Feedback Errors", Under Review at the Journal of Wireless Communications and Mobile Computing, submitted on January 20, 2013.
4. Coşkun A. F., Kucur O., (2013), "Feedback-Error-Resistant and Reduced Feedback-Rate Orthogonal Space-Time Block Coding Scheme Employing Single Transmit Antenna Selection", Under Review at the Journal of IET Communications, submitted on August 07, 2013.
5. Coşkun A. F., Kucur O., (2013), "Analysis of Hybrid Diversity Schemes Employing Antenna Selection in the Presence of Channel Estimation Errors and Feedback Delay", Under Review at the Journal of IET Communications, submitted on August 28, 2013.

National Proceedings

1. Coşkun A. F., Kucur O., (2011), "Performance of space-time block coded systems with joint transmit and receive antenna selection in Nakagami-m fading channels", in Proc. IEEE SIU 2011, 118-121, Antalya, Turkey, 20-22 April.

2. Coşkun A. F., Kucur O., (2012), "Error performance of space-time block coded systems with joint transmit and receive antenna selection in the presence of feedback errors", IEEE SIU 2012, 1-4, Muğla, Türkiye, 18-20 April.
3. Coşkun A. F., Kucur O., (2013), "Modified transmit antenna selection/orthogonal space-time block coding scheme with reduced feedback-rate and robust performance in the presence of feedback errors", IEEE SIU 2013, 1-4, Girne, Turkish Republic of Northern Cyprus, 24-26 April 2013.