

PRE-PROCESSING BASED ADVANCED SPATIAL MODULATION
TECHNIQUES

by

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ABSTRACT

PRE-PROCESSING BASED ADVANCED SPATIAL MODULATION TECHNIQUES

Spatial Modulation (SM) extends the concept of index modulation into the spatial domain, forming a specialized type of multiple-input multiple-output (MIMO) architecture. Inspired by the advantages of the SM, this thesis deals with systems that utilize different forms of advanced SM. Each of these systems is designed with a specific purpose, among these objectives are robustness to channel and antenna correlations, reliability, energy efficiency and improved bit error rate (BER) performance. This thesis proposes different communication systems employing advanced SM techniques in six chapters as the following: i) a precoding strategy for quadrature spatial modulation (QSM), aiming to enhance robustness against the effects of Rayleigh/Rician fading and spatial correlations among the transmit antennas, ii) a dual-polarized (DP) QSM scheme and a precoding approach to make the system more robust against spatial correlation and channel fading effects, iii) the joint use of DP antennas and of active antenna dependent signature constellations, to improve performance in terms of robustness to transmit antenna correlation, particularly for Rician fading channels, iv) a novel precoding technique to address the correlation issue for multi-stream SM (MSM) in which multiple antennas are active at a time, v) an innovative backscatter communication (BC) system design based on tag-side pre-processing aided receive spatial modulation (PSM), involving zero-forcing (ZF) and minimum mean squared-error (MMSE) precoders to achieve reduced complexity, enhanced energy efficiency and superior BER performance, vi) a reconfigurable intelligent surfaces (RIS) assisted tag-side PSM based BC system is introduced to make control of the signal propagation, reception and transmission. This achieves better BER performance and energy efficiency in comparison to conventional benchmark approaches. In all proposed architectures, the BER performances are derived analytically, which are confirmed by simulations.

ÖZET

ÖN İŞLEME TABANLI İLERİ UZAYSAL MODÜLASYON TEKNİKLERİ

Uzaysal modülasyon (SM), indeks modülasyon kavramını uzaysal alana genişletip, özelleşmiş bir çoklu-giriş çoklu-çıkış (MIMO) mimarisi oluşturur. SM'nin avantajlarından esinlenerek, bu tez farklı ileri SM formlarını kullanan sistemlerle ilgilenmektedir. Bu sistemlerin her biri belirli bir amaçla tasarlanmıştır; bu amaçlar arasında kanal ve anten korelasyonlarına karşı sağlamlık, güvenilirlik, enerji verimliliği ve iyileştirilmiş bit hata oranı (BER) performansı bulunmaktadır. Bu tez, altı bölümde ileri SM tekniklerini kullanan farklı iletişim sistemlerini önermektedir: i) Dördül uzaysal modülasyon (QSM) için bir ön kodlama stratejisi, Rayleigh/Rician solma ve iletim antenleri arasındaki uzaysal korelasyon etkilerine karşı sağlamlığı artırmayı hedeflemektedir, ii) Uzaysal korelasyon ve kanal solma etkilerine karşı sistemi daha dayanıklı hale getirmek için çift kutuplanmalı (DP) bir QSM şeması ve ön kodlama yaklaşımı, iii) DP antenlerin ve aktif anten bağımlı imza sabit dizilerinin birlikte kullanımı, özellikle Rician solma kanallarında iletim anteni korelasyonuna karşı sağlamlığı artırmak amacıyla, iv) Aynı anda birden fazla antenin aktif olduğu çok akışlı SM (MSM) için korelasyon sorununu ele almak üzere yeni bir ön kodlama tekniği, v) etiket tarafı ön işlem destekli alım uzaysal modülasyonu (PSM) temelli yenilikçi bir yansıtma iletişimi (BC) sistem tasarımı, düşük karmaşıklık, artırılmış enerji verimliliği ve üstün BER performansı elde etmek için sıfır zorlama (ZF) ve minimum ortalama kare hatası (MMSE) ön kodlayıcılarını içerir, vi) Sinyal yayılımının, alımın ve iletimin kontrolünü sağlamak amacıyla yeniden yapılandırılabilir akıllı yüzeyler (RIS) destekli etiket tarafı ön işlemleri SM tabanlı BC sistem tanıtılmıştır. Bu, geleneksel referans yaklaşımlara kıyasla daha iyi BER performansı ve enerji verimliliği sağlar. Önerilen tüm mimarilerde, BER performansları analitik olarak türetilmiş ve simülasyonlarla doğrulanmıştır.

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LIST OF SYMBOLS

$\mathcal{CN}(m_x, \sigma_x^2)$	Complex Gaussian random variable with mean m_x and variance σ_x^2
$D(\cdot)$	Distance metric
$E[\cdot]$	Expectation operator
$\mathbf{H}$	Channel matrix
$\overline{\mathbf{H}}$	Deterministic matrix with all one elements
$\bar{\mathbf{H}}_w$	Fixed part of the channel $\mathbf{H}$
$\tilde{\mathbf{H}}_w$	Variable part of the channel $\mathbf{H}$
$\mathbf{h}_u$	u th column of $\mathbf{H}$
$h_{i,j}$	Channel coefficient between i th receive antenna and j th transmit antenna
$I_0(\cdot)$	Modified Bessel function of the first kind
K	Rician factor
N_0	Variance of the noise
$\mathcal{N}(m_x, \sigma_x^2)$	Gaussian random variable with mean m_x and variance σ_x^2
P_r	Average received power
$\mathbf{R}_r$	Receive rotation matrix
$\mathbf{R}_t$	Transmit rotation matrix
R	Spectral efficiency
$\text{Re}\{x\}$	Real part of a complex variable
$\text{vec}(\cdot)$	Vectorization of a matrix
X	Transmitted symbol
$\mathbf{x}$	Transmitted signal
$\mathbf{y}$	Received signal
α_r	Spatial correlation coefficient at the receiver
α_t	Spatial correlation coefficient at the transmitter
χ_f	The ratio of co-polar to cross-polar term for fixed part of the channel
χ_v	The ratio of co-polar to cross-polar term for variable part of the channel

η	Signal to noise ratio (SNR)
$\mathbf{\Gamma}_f$	Leakage matrix for fixed part of the channel
$\mathbf{\Gamma}_v$	Leakage matrix for variable part of the channel
γ_r	Polarization correlation coefficient at the receiver
γ_t	Polarization correlation coefficient at the transmitter
Δ	Estimation error matrix
μ_f	Power leakage amount from one polarization to its orthogonal counterpart for fixed part the channel
μ_v	Power leakage amount from one polarization to its orthogonal counterpart for variable part the channel
$\mathbf{\Pi}_t$	Polarization correlation matrix at the transmitter
$\mathbf{\Pi}_r$	Polarization correlation matrix at the receiver
Ψ	Diagonal precoding matrix
$\psi_{i,j}$	i, j th element of the precoding matrix
σ_e^2	Variance of the channel estimation error
$\sigma_{i,j}^r$	i, j th element of receive spatial correlation matrix
$\sigma_{i,j}^t$	i, j th element of transmit spatial correlation matrix
θ_r	Rotation angle relative to vertical direction at the receive side
θ_t	Rotation angle relative to vertical direction at the transmit side
θ_u	Rotation of the symbols transmitted from u th unique active transmit antenna and polarization pair
ν	Noise vector
Φ_r	Receive correlation matrix
Φ_t	Transmit correlation matrix
Ξ_r	Correlation matrix of the channel estimation error at the receiver
Ξ_t	Correlation matrix of the channel estimation error at the transmitter
$\{\cdot\}^*$	Complex conjugation
$ \cdot $	Determinant of a matrix
$\{\cdot\}^\dagger$	Transposition of a matrix followed by conjugation
$\{\cdot\}^H$	Hermitian of a matrix

$\{\cdot\}^T$	Transpose of a matrix
$\odot$	Element by element Hadamard multiplication
$\otimes$	Kronecker product



LIST OF ACRONYMS/ABBREVIATIONS

ABEP	Average Bit Error Probability
APEP	Average Pairwise Error Probability
APM	Amplitude Phase Modulation
AmBC	Ambient Backscatter Communication
AWGN	Additive White Gaussian Noise
BC	Backscatter Communication
BER	Bit Error Probability
BICM	Bit-Interleaved Coded Modulation
bpcu	Bit per Channel Use
BT	Backscatter Transmitter
CSI	Channel State Information
CSIT	Channel State Information at Transmitter
C-SM	Conventional Spatial Modulation
CW	Continuous Wave
DP	Dual-polarized
DP-QSM	Dual-polarized Quadrature SM
ESM	Enhanced SM
exp	Exponential Function
G-SM	Generalized Spatial Modulation
i.i.d	Independent and Identically Distributed
ICI	Inter Channel Interference
IM	Index Modulation
IoT	Internet of Things
LOS	Line of Sight
LS	Least Squares
MGF	Moment Generating Function
MIMO	Multiple Input Multiple Output
ML	Maximum Likelihood
MMSE	Minimum Mean Squared Error
MSM	Multi-stream SM

NLOS	Non Line of Sight
NOMA	Non-orthogonal Multiple Access
PDF	Probability Density Function
PEP	Pairwise Error Probability
PSM	Pre-processed Receive Spatial Modulation
QSM	Quadrature SM
RF	Radio Frequency
RFID	Radio Frequency Identification
RIS	Reconfigurable Intelligent Surfaces
SM	Spatial Modulation
SMX	Spatial Multiplexing
SNR	Signal to Noise Ratio
SSK	Space Shift Keying
SSM	Signature Spatial Modulation
STBC	Space-Time Block Coding
STSK	Space-time Shift Keying
UEP	Unequal-error Protection
UP	Uni Polarized
VH	Vertical-Horizontal
XPC	Polarization Correlation
XPB	Cross-Polar Discrimination
XPI	Cross-Polar Interference
ZF	Zero-Forcing

1. INTRODUCTION

1.1. Related Background

Wireless communication technology is currently dealing with several challenges, including the need for enhanced data rates, greater system capacity and improved service performance. To address these issues, multiple-input multiple-output (MIMO) communication has emerged as a promising solution. MIMO technology, which involves the use of multiple antennas at both the transmitter and receiver ends, introduces an additional spatial dimension to communication. This extra dimension provides a degree-of-freedom gain, leading to increased capacity. Moreover, the use of this new dimension significantly boosts reliability and data rates. The presence of multiple antennas in a system creates various propagation paths. When identical data packets are transmitted from different active antennas, they may undergo varying levels of fading. This variability enhances the probability of the transmitted data being accurately received, thereby increasing system reliability. This benefit is known as spatial diversity or diversity gain. Another advantage of MIMO systems is the multiplexing gain, which is achieved by transmitting independent information symbols through different active antennas. Multiplexing gain, often referred to as degrees of freedom relative to the signal space constellation, allows for an increase in data rate proportional to the minimum number of transmitters and receivers, without necessitating an increase in spectrum usage or transmit power.

A primary challenge in MIMO systems is the complexity of implementation, which increases with the number of antennas. Spatial Modulation (SM) [1], [2], has recently emerged as a popular transmission technology within MIMO communication systems, which significantly simplifies this complexity, particularly on the receiver side, and often delivers superior multiplexing and, in some instances, diversity gains compared to other MIMO technologies with equivalent spectral efficiency. The core concept of SM is straightforward: Information is transmitted not only through digital modulation within the signal space but also through the spatial positioning of the antennas. In this method, the information bits are divided into two parts: the first part selects a

single transmit antenna to activate, and the second part selects a modulation symbol from the available constellation. SM improves the error performance significantly, particularly at high spectral efficiencies, which results from increased minimum distances between constellation points since this technology assigns several bits to select the antenna index, thus reducing the size of the constellation compared to other MIMO configurations. In the conventional SM (C-SM) setup, where only one antenna is active on the transmitter side, inter-channel interference (ICI) is entirely avoided, which simplifies the receiver design compared to other spatial multiplexing methods. C-SM represents a specific case of generalized SM (G-SM) [3–6], where the initial bit segment activates multiple transmit antennas simultaneously. G-SM can transmit the same symbols from active antennas to achieve diversity gains, or different symbols can be transmitted from each active antenna to attain multiplexing gains. Overall, due to its rate improvements, offering greater robustness against channel correlation and channel estimation errors, reduced complexity in implementation, and enhanced performance, SM-based wireless communication systems have become widely used and actively researched areas in recent years. In SM systems, the enhancement in data rate is directly proportional to the base-two logarithm of the number of antennas on the transmitter side. This implies that to achieve a 1 bit/s/Hz increase in spectral efficiency, it is necessary to either double the number of transmit antennas or the constellation size, which may not be costly effective. This increases the interest to boost the data rate in SM systems recently.

Quadrature SM (QSM) [7], a modulation scheme proposed recently, improves the overall spectral efficiency of SM systems by decomposing an M-ary symbol into its real (in-phase, I) and imaginary (quadrature, Q) components, which are transmitted independently using selected antennas. To be more specific, information bits are divided into three groups: the first two groups are used to select indices of the two active antennas and the remaining part is used to select one modulation symbol to be transmitted. The computational complexity of QSM is comparable to that of traditional SM, and the no-ICI structure of SM is maintained in QSM, as two mutually orthogonal carriers are utilized to transmit information. While the conventional SM and QSM systems offer several advantages, their simultaneous use of antenna and signal spaces to transmit information presents a significant drawback. Especially when there is significant

transmit antenna correlation, the channels from various transmit antennas to each receive antenna become nearly identical, which substantially makes the detection of the group of bits responsible for selecting the active antenna impossible. Consequently, this results in the unreliable detection of bits assigned to the antenna index, leading to a considerable increase in the overall average bit-error rate (ABER) of the SM/QSM transceiver. Recently introduced approaches for solving this problem rely on the use of either channel coding to obtain unequal error protection (UEP) as in [8–10], where an encoder-decoder pair needs to be added to the transceiver, which increases the latency and the computational complexity significantly. To address this issue, the precoding approach for SM system was designed in [11] which is based on phase-rotation and/or amplitude scaling of the transmitted symbols according to the active transmit antenna(s) and has been adapted for time-varying fading channels to enhance system robustness against both the Rayleigh/Rician fading effects and also spatial correlations among transmit antennas. The zero ICI structure remains intact in this model since the precoding operation simply involves multiplying the transmitted signal vector by a diagonal matrix, eliminating the need for explicit channel state information (CSI). Additionally, this method does not alter the average transmit power budget and can be implemented with only a slight increase in transmitter circuitry complexity.

Precoding methods have demonstrated notable enhancements for simpler constellations like BPSK and QPSK, where there is a substantially large minimum Euclidean distance between the rotated constellations. However, as the complexity of the constellation increases as M -ary QAM for $M = 16, 64, \dots$ the minimum distance between constellations reduces, and the benefits from precoding become less significant. Additionally, the complexity involved in optimizing precoding escalates markedly with an increase in the number of transmit antennas and the modulation order. As a result, while precoding-based strategies are effective for smaller-scale MIMO systems using simple modulation schemes, they prove to be less practical for high spectral efficiency applications where addressing channel and transmit antenna correlation effects is critical. To overcome this limitation, signature spatial modulation (SSM) has been developed [12], where transmitted symbols are selected from various constellations, each representing the signature of one of the N_t transmit antennas. These constellations, first introduced in earlier studies [13] and [14], evolve from one another through se-

quential geometric interpolations, starting from the primary constellation plane. They are designed to optimize the minimum Euclidean distance between constellations. Significant performance improvements have been achieved with SSM, particularly with higher-order modulations, surpassing both conventional SM and precoded SM in scenarios with strong transmit antenna correlations. As mentioned previously, SM can be significantly impacted by spatial correlation, which is influenced by antenna spacing. The performance deterioration due to spatial correlation necessitates a large antenna separation, which can be impractical in scenarios involving small devices due to space constraints. To address the challenges posed by limited space, the use of dual-polarized (DP) antennas has been proposed. These antennas transmit data streams through two orthogonal polarizations simultaneously, effectively introducing a polarization dimension that mitigates spatial correlation and reduces spatial requirements by replacing a single DP antenna with two uni-polarized (UP) antennas [15–20]. Additionally, the use of the polarization domain offers polarization diversity, enhancing channel capacity [19], [21] and [22]. Studies have shown that DP systems generally perform better than their UP counterparts under non-ideal conditions. Despite the benefits, using DP antennas introduces two types of correlations: cross-polar discrimination (XPD) and polarization correlation (XPC) [16], [23] and [18]. XPD, a measure of polarization mismatch defined as the ratio between the average received power on the same and orthogonal polarizations, varies with different channel environments and can result in a loss of diversity if its value is high [24–26]. XPC results from polarization imperfections and typically has a minimal practical impact, often being overlooked.

SM was initially developed for MIMO systems, characterized by having fewer transmit power amplifiers and radio frequency (RF) chains than antennas [27], [28]. This configuration is especially beneficial for the uplink in cellular systems where the cost and power constraints make it impractical to have as many RF chains as antennas, as might be required for optimal downlink performance. In SM, the number of active transmit antennas at any moment is limited to the available RF chains, restricting the uplink data streams. However, by not activating all antennas simultaneously, additional data can be conveyed through the indices of the antennas that are transmitting. This principle underlies SM, termed single-stream SM when only one antenna is active, and multi-stream SM (MSM) when more than one is active [6].

A significant challenge in SM occurs with highly correlated channels. Apart from decoding the transmitted symbols, the receiver must also identify which antennas were used for transmission. High correlation between channels complicates this identification, adversely affecting the BER of SM. Addressing this issue has been the focus of substantial research, resulting in various strategies to mitigate antenna and channel correlation in single-stream SM [11], [12, 29–34]. The most effective approach involves linear precoding at the transmitter, which creates distinct signals from each antenna, making them recognizable by the receiver even if the channel responses are similar. Popular linear precoding methods include rotation-based precoding [11], [29–33], suitable for simpler constellations like BPSK and QPSK, and signature constellation-based precoding [34], [12], which offers greater efficiency, particularly with more complex constellations like 16-QAM and 64-QAM.

Backscatter communications (BC), an innovative ultra-low-power communication approach, has been introduced as a simple yet effective method that involves small tags capable of harvesting energy from a transmitter to receive, modulate, and reflect RF signals. This technology, which potentially eliminates the need for batteries and maximizes spectrum efficiency, is becoming increasingly important for internet of things (IoT) applications and continues to draw significant research interest.

BC systems are divided into three main categories: monostatic, like radio frequency identification (RFID) systems; bistatic; and ambient backscatter communication (AmBC) systems. In each type, the backscatter tag modifies its impedance to switch between reflective and absorptive states, thereby transmitting information without generating its own RF signals. This feature dramatically reduces power consumption, making these systems ideal for battery-powered IoT devices by extending their operational life and reducing the need for frequent battery replacements. In bistatic architecture, the RF source and receiver are separated, offering improved performance and cost-effectiveness compared to its monostatic counterpart. In AmBC systems, the transmitter and backscatter receiver are also separated, but there is no dedicated RF source; instead, it utilizes ambient RF sources like cell towers and TV stations. Although this optimizes spectrum utilization, it can cause interference issues that affect detection performance and reduce transmission range. Bistatic BC system,

has simpler, cost-effective design which also reduces interference from ambient signals, unlike the AmBC system which suffers from this issue.

Backscatter tags are categorized into passive, semi-active, and active types based on their energy needs. Semi-active tags, which, like passive tags, rely on RF signal excitation but also include a power source to enhance the functionality of sensors or memory, are improving reliability and accessibility related to active transmitter tags. This allows for the development of more sophisticated applications at a higher cost and with limited lifespan compared to passive tags.

The drive for improved reliability, spectral efficiency, and transmission range in IoT networks is pushing the development of advanced BC configurations. Integrating MIMO technology into BC is expected to be crucial for enhancing these aspects in Beyond-5G systems. Further enhancement is possible by incorporating index modulation, a specialized MIMO technique, into SM strategies, where information is transmitted using both signal and antenna dimensions. This approach is being explored extensively and is being applied in both SM and GSM based AmBC systems to utilize the transmitter tag index as an additional information dimension. Advanced transmission techniques like these are the technologies for higher data rates and more efficient communication in BC systems.

The management of wireless channels is gaining increasing interest, especially in the context of enabling data transfer between smart devices without relying on network resources. This capability is essential for the effective operation of intelligent systems in advanced 5G and upcoming 6G technologies, as highlighted in [35]. This surge in attention is largely due to the innovative use of reconfigurable intelligent surfaces (RIS), which allow for the dynamic modification of signals [36–40]. RIS technology, also referred to as software-defined meta-surfaces, smartly controls electromagnetic waves through passive elements that can individually alter the signal's amplitude and/or phase. This facilitates precise three-dimensional reflection beamforming without the need for active radio frequency (RF) signal generation.

Despite the implementation of meta-surfaces, the standout feature of RIS technology in terms of energy efficiency lies in its ability to amplify and forward incoming signals without needing a power amplifier. This is achieved through meticulously designed phase shifts by each reflecting element, which effectively combine reflected signals. Moreover, the compact design of RIS structures makes them easy to integrate into various communication environments, allowing for straightforward installations in places like factory ceilings, indoor spaces, laptop cases, and even within clothing [41]. RIS incorporates numerous low-cost, nearly passive elements that not only create additional indirect links between transmitters and receivers but also significantly reduce energy consumption by boosting signal strength, enabling devices to operate at lower power levels.

RIS also shows considerable potential in meeting the increasing demands for higher data rates, broader coverage, and improved energy efficiency in wireless networks. By dynamically adapting to changing channel conditions and user requirements, RIS enhances system robustness, effectively mitigates multi-path fading, improves received signal quality, and boosts overall reliability [40], [42], [43].

Additionally, the integration of energy harvesting capabilities into devices is emerging as a crucial development in various wireless applications. In this context, BC [44–47] represents a groundbreaking approach to energy harvesting. This technology utilizes small tags that capture energy emitted by transmitters to perform functions such as receiving, modulating, and backscattering RF signals. The potential for battery-free, spectrum-efficient communication in a low-power consumption mode makes BC a viable option for powering IoT applications, attracting significant and growing research interest. This shift toward sustainable power sourcing signifies a fundamental transformation in how energy is obtained and utilized. Unlike BC, which uses low-cost passive tags to reflect and modulate RF signals while harvesting energy, RIS primarily enhances signal quality between the source and destination without transmitting its own information [48].

A recent series of studies have introduced the use of RIS to boost the performance of BC systems [49–54]. For example, an analysis of AmBC systems utilizing RIS

is conducted in [51], where the BER is calculated when RIS is integrated into the source-to-reader link, the source-to-tag link, or both. The findings indicate that adding more RIS elements to the source-to-reader link tends to worsen performance, whereas enhancements in the source-to-tag link lead to improved outcomes. A novel approach to incorporating RIS in a bistatic BC system, which achieves considerable savings in transmit power for both single-tag and multi-tag configurations, is introduced in [55], even with only moderate sizes of RIS. Further, in [56], there is an exploration of signaling design and phase shift optimization for an IRS-assisted BC system within indoor environments aimed at enhancing detection performance. The study in [54] models an RIS-assisted backscatter communication setup, where there is no direct link between the tag and reader, relying solely on RIS elements, analyzing system performance amid hardware imperfections. An experimental evaluation, the first of its kind, is performed in [57], showcasing how RIS can enhance the functionality of ambient backscatter systems. Additionally, RIS is applied to cooperative BC in [52], where RIS elements alternate between reflecting and energy-harvesting modes. This research proposes a transmission design aimed at minimizing RIS power consumption. The incorporation of RIS into backscatter communication systems to enable 6G IoT networks is detailed in [58], highlighting its potential in advancing future connectivity solutions.

1.2. Motivation for This Thesis

Conventional SM/QSM technology, in which there is only a single transmit antenna activated at the signalling instant, removes ICI completely, relaxes inter-antenna synchronization requirements, decreases the complexity of the receiver and allows the use of a single RF chain at the transmitter so that reduces the average power consumption. However, there are some limitations with the SM technology, motivating researchers to enhance the systems in order to overcome these limitations. The main problem of the SM comes from the joint utilization of both the antenna and signal spaces to convey information which leads to effecting the detection of antenna and signal bits unequally and increases the overall average bit error probability. Especially in the case of channel correlations and small antenna spacing distances deteriorates the reliability of the detection. The precoding approach introduced in the second chapter

of this thesis for QSM system is designed such that to increase the robustness of the QSM system against channel correlations and LOS channel component.

As mentioned previously, antenna spacing is also an important factor to avoid channel correlations, especially UP antennas need higher spacing to avoid the similarity between channels, which can not be feasible in applications operating in small spaces. This motivates us to add polarization dimension to the conventional QSM system and introduce a DP-QSM technology in the third chapter of this thesis. In addition a precoding approach is proposed in this chapter to increase the minimum Euclidean distance among the employed constellation by rotating and amplitude scaling of the constellation. This precoding approach is done to increase the reliability of the detection of the bits associated to the antenna index.

The precoding approaches which are based on the constellation rotation or amplitude scaling are operating well in the case of small antenna number and small order constellations. This limits the application of precoding scenario in higher order systems. To achieve robustness against channel and spatial correlations in the case of DP-SM systems operating in high order schemes, we introduce a SSM technique in the Chapter 4 of this thesis.

The mentioned solutions to the problem of the sensitivity of SM systems to channel and antenna correlations and the LOS channel component are performing adequately for the case of single stream SM systems, in which there is only one active transmitting antenna. For the case of multi-stream SM system, where there are more than one active antenna at the transmitter side, these approaches do not add significant improvement in terms of BER performance. This problem motivates us to introduce a novel precoding approach to counteract antenna and channel correlations in multi-stream SM systems in Chapter 5.

Another technique in the domain of SM involves performing a pre-processing at the transmitter side, in order to send the majority of the transmitted signal power to a desired receiver. This is named as pre-processing aided receive spatial modulation (PSM) which conveys extra information bits by strategical selection of the receive

antenna indices instead of the transmit antenna indices as in SM. More explicitly, under the assumption of channel state information at the transmitter (CSIT), the transmitter possesses the capability to precisely direct the transmit power to a single receive antenna at the receiver through the use of precoding. This means that in PSM, information bits are conveyed by the index of the received antenna and the symbol drawn from a conventional modulation constellation. This approach, mitigates channel effects through pre-processing and offers superior detection performances compared to conventional SM employed at the transmitter. Considering the mentioned benefits of the PSM scheme, the main idea behind Chapter 6. of this thesis, is to extend this transmission technique to the bistatic BC system. The objective of this work is to leverage the advantages of a bistatic backscatter architecture by employing tag-side pre-processing assisted SM transmission, aiming to establish a comprehensive mathematical framework and introduce a system model adaptable to a majority of next-generation IoT applications.

The increasing demand to control the wireless channels, including signal propagation, reception, and transmission, leads to the adoption of RIS elements, which can manipulate the phase, amplitude, and frequency of incoming waves through passive elements. Previous works employing RIS elements through SM, SSK and other index modulation systems, are the result of the improvement in terms of energy consumption and better BER performance. This motivates us to apply RIS elements between transmitter tags and reader node to the system model described in Chapter 6. to control over the signal which is backscattered and received by a targeted receiver.

1.3. Contributions of This Thesis

This thesis studies advanced SM techniques in communication systems and introduces different architectures to make these systems perform better in terms of robustness to channel and spatial correlations, line-of-side (LOS) effect, energy efficiency, power consumption and BER performance. To summarize, the contributions of this thesis are as follows.

- (i) In Chapter 2, a precoding architecture is considered for QSM system, to enhance the deteriorating performance of QSM with correlation and Rician channel characteristics. The primary benefit of the proposed approach is its ability to enhance performance significantly while only slightly increasing the complexity of the transmitter-receiver system, without requiring detailed channel knowledge at the transmitter side. The only necessary information is the channel's fixed component, the Rician factor, and the correlation among the transmitter antennas. With this fundamental setup, a substantial improvement in performance can be achieved, particularly under Rician fading conditions.
- (ii) In Chapter 3, the DP-QSM system is presented, exploring how the QSM architecture can be expanded through the addition of polarization dimensions. A precoding method for DP-QSM systems is examined, specifically designed for slow-fading correlated channels, to enhance the system's resilience against channel disruptions and correlation effects.
- (iii) In Chapter 4, the benefits of SSM and dual-polarized transmission are integrated to develop a robust SM-MIMO transceiver architecture that resists the impacts of transmit antenna and channel correlations. Both theoretical analysis and comprehensive simulations demonstrate that the proposed method can provide considerable performance improvements over traditional DP-SM and equivalent UP-SM systems.
- (iv) In chapter 5, a new precoding strategy is introduced to mitigate the impact of transmit antenna and channel correlations in multi-stream SM. This technique utilizes signature constellations; however, rather than allocating a constellation to each antenna, it assigns a specific set of signature constellations to each combination of active antennas. Additionally, phase rotation and/or amplitude scaling are applied to address any singularities, even under conditions of complete correlation.
- (v) In Chapter 6, a tag-side pre-processing assisted SM strategy is detailed, tailored for bistatic BC systems that function under the CSIT mode with semi-active transmitter tags. This work analyzes the optimization of pre-processing and signal detection within this framework. Both MMSE and ZF architectures are introduced, complete with theoretical error performance and complexity analyses.

These theoretical derivations are substantiated by Monte-Carlo simulations conducted across various configurations.

- (vi) In Chapter 7, an innovative communication framework known as the backscatter communication (BC), utilizing pre-processed receive spatial modulation (PSM) technique for data transfer assisted by RIS elements is introduced. This architecture is done in response to the critical need for optimizing the spectral and energy efficiency of these devices, and in accordance with the evolving standards of advanced 5G and potential 6G networks. This chapter involves the signaling design and phase shift optimization of an RIS-assisted PSM based BC system in addition to the comprehensive evaluation of these strategies concerning the BER performance.

2. PRECODED QUADRATURE SPATIAL MODULATION

2.1. Introduction

Index Modulation (IM) leverages the indices of transmit antennas, receive antennas, linear block codes, time slots, or other mediums to transmit additional information without needing extra resources [59–61]. SM [2], a variant of IM applied in the spatial domain, is recognized as a specialized form of the MIMO architecture. Variants of SM like space shift keying (SSK) and space-time shift keying (STSK) are detailed in [62] and [63] respectively. These techniques utilize the dimension of transmit antennas as spatial constellation points to carry information bits. In SM, activating a single transmit antenna per transmission instant eradicates inter-channel interference (ICI), eases inter-antenna synchronization, simplifies the receiver design, and reduces average power consumption by using a single RF chain [64], [65]. SM also shows greater robustness against spatial channel correlation and channel estimation errors, improving error performance over traditional MIMO methods. The simplicity, improved performance, and rate enhancements makes SM-based systems an important point of contemporary wireless communication research. Comprehensive insights into recent developments, design principles, practical implementations, and potential applications of SM are available in [27] and [65–67].

In SM, information bits are divided into index bits and modulation bits, where the index bits are transmitted using the index of the active transmit antenna. In SSK, a derivative of SM, information is transmitted solely through the index of the activated transmit antenna, yielding higher energy efficiency and reduced detection complexity. Additionally, the SM/SSK scheme has been expanded into the space-time domain to improve its bit error rate (BER) performance, with STSK [63].

While in traditional MIMO technologies spectral efficiency increases linearly with the number of transmit antennas, these systems often suffer from high ICI at the receiver. In contrast, SM's data rate enhancement is logarithmic relative to the number of antennas, necessitating a doubling of the number of transmit antennas to increase

spectral efficiency by 1 bit/s/Hz, an impractical strategy. Consequently, enhancing data rates in SM systems is a significant area of focus, with various methodologies proposed in recent literature [3–6], [13], [68–70]. Generalized SM (GSM) [68] was introduced to enhance the overall spectral efficiency, in which the mentioned scheme maps each index to the antenna bits addressing a group of simultaneously activated antennas, where the active antennas may transmit the same symbol and benefit from diversity gain or each active antenna may transmit different symbols and thus achieve multiplexing gain. So in GSM, the spectral efficiency is enhanced at the cost of multiple RF chains. However, the performance degradation of these systems in comparison to SM systems was reported. An extended constellation design for GSM is proposed in [69] and [70], in which the number of activated antennas is different from symbol to symbol, this makes the system have a higher number of antenna combinations which results in better error performance in addition to the enhanced spectral efficiency. A multiple active-SM (MA-SM) scheme is proposed in [6], combining the traditional SM architecture with spatial multiplexing to improve the data rate significantly at the cost of increasing ICI. Another scheme is presented in [13] as enhanced SM (ESM), which introduces multiple signal constellations for active antennas in order to send the information bits. In this methodology, the secondary constellation is derived by the geometric interpolation from the primary constellation and it is seen from the results that ESM achieves higher throughput and in the case of the same spectral efficiency as SM, it reaches better error performance.

Quadrature SM (QSM) [7], a new modulation scheme, expands the spatial constellation domain into a new dimension for signal transmission and surpasses the performance of generalized SM (GSM). In QSM, selected antennas transmit the real (in-phase, I) and imaginary (quadrature, Q) components of an M-ary symbol independently, maintaining the no-ICI structure of SM with identical computational complexity. QSM demonstrates equivalent error performance with 3 dB less signal power compared to SM at the same spectral efficiency.

QSM technology, a promising technology for wireless communications, has been the subject of research focusing on enhancing throughput of conventional systems [31] and [71–79]. Innovative approaches for improving QSM's system performance or in-

creasing spectral efficiency further include a generalized framework for QSM (GQSM) [72], improved QSM (IQSM) [80], and parallel QSM for reliable transmission in massive MIMO systems [81]. These developments showcase QSM's potential for high-throughput secure transmissions and its applicability in multi-user MIMO transmissions.

Despite these advancements, using antenna and signal spaces to jointly convey information introduces a significant challenge, notably the unreliable detection of bits assigned to the antenna index, leading to increased overall average bit-error rate (ABER) in SM/QSM systems. This issue, termed unequal error vulnerability of symbol bits and antenna against channel fading effects, has prompted a variety of solutions aimed at enhancing robustness against channel fading and spatial correlations in SM systems [8–10], [29], [30], [32], [82], [83]. These include unequal-error protection (UEP) schemes and approaches based on bit-interleaved coded modulation (BICM) and block-Markov superposition coding. Additionally, a diversity-achieving QSM (DA-QSM) transmission scheme has been introduced [84], providing higher spectral efficiency than space-time block coded SM (STBC-SM) [85] and spatially modulated orthogonal STBC, with better BER performance.

The precoding method presented in this chapter of the thesis is designed around constellation shaping and has been adapted for time-varying fading channels to enhance system robustness against adverse channel effects. This method is particularly effective for the QSM system, aiming to boost its performance, which is a central theme of this chapter. The technique involves phase rotation and/or amplitude scaling of transmitted symbols, maintaining the zero ICI structure, as the precoding process is analogous to multiplying the transmitted signal vector with a diagonal matrix. Moreover, this approach does not require explicit CSI, does not alter the average transmit power, and involves only a slight increase in transmitter circuitry complexity. The optimal complex constellation scaling coefficients are determined by minimizing the asymptotic ABEP at high signal-to-noise ratio (SNR) values.

This work conducts a thorough analysis of QSM system performance, comparing it to conventional SM systems under fading channel conditions. It establishes a tight

upper bound on the overall average ABEP and calculates the pairwise error probability (PEP) of QSM in a definitive form. Additionally, a new precoding strategy is introduced to improve the QSM system's resilience against channel fading effects and spatial correlations among transmit antennas. This precoding technique relies on phase rotation and amplitude scaling of the transmitted symbols according to the active antennas. Alternatively, the precoding could solely involve rotating the constellation based on the active antenna to simplify optimization and implementation. In both scenarios, the transmitter is assumed to know only the fixed component of the channel and possibly its spatial correlation structure, but not the specific fading channel details. The optimization process aims to minimize the Average Bit Error Rate (ABER) at high SNR, with appropriate selection of rotation and amplitude coefficients. Theoretical analysis and simulations indicate that the precoded QSM (PQSM) system could significantly enhance robustness against spatial correlations at the transmitter side and Rician fading scenarios compared to standard SM and QSM systems. Thus, integrating the benefits of QSM with precoding techniques might facilitate the development of QSM-MIMO systems that are more resistant to Rician fading and spatial correlations at the transmitter side.

2.2. System Model

2.2.1. Quadrature Spatial Modulation

Consider an $N_t \times N_r$ QSM system, where N_t and N_r denote the number of transmit and receive antennas, respectively. For simplicity, N_t is assumed to be an integer power of 2 ($n = \log_2(N_t)$). The system employs an M -ary constellation, where m bits are assigned to a constellation $\mathcal{X} = \{X_1, \dots, X_k, \dots, X_M\}$ containing M symbols, thus $m = \log_2 M$. At each transmission instant, $2n + m$ bits are divided into three groups: two groups of n bits each to select the indices of the two active antennas, u_1 and u_2 where $u_1, u_2 = 1, \dots, N_t$, and the remaining m bits to select a modulation symbol to be transmitted. Suppose the selected symbol is X_v for $v = 1, \dots, M$, then the u_1 -th antenna transmits the real part of X_v whereas u_2 -th antenna sends imaginary part of X_v . Notice that the spectral efficiency of the system is $R = 2n + m$ b/s/Hz which is n b/s/Hz higher than that of SM with the same transmission parameters.

The transmitted signal vector is $\mathbf{x} = [0, \dots, x_{u_1}, \dots, x_{u_2}, \dots, 0]^T$ where only one (if the selected antennas are the same) or two entries are non-zero. In other words, the locations of the non-zero elements indicate which antennas are active, and the values of these elements represent the real and imaginary components of the transmitted symbol, i.e., $x_{u_1} = \text{Re}(X_v)$ and $x_{u_2} = \text{Im}(X_v)$. We assume a power constraint of unity, i.e., $E_{\mathbf{x}} = [\mathbf{x}^H \mathbf{x} = 1]$. Accordingly, the received signal model is

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \boldsymbol{\nu}, \quad (2.1)$$

where $\mathbf{y}$ and $\boldsymbol{\nu}$ are $N_r \times 1$ received signal and noise vectors respectively, $\mathbf{H}$ is $N_r \times N_t$ channel matrix and $\mathbf{x}$ is a $N_t \times 1$ transmitted signal vector. $\boldsymbol{\nu}$ has independent and identically distributed (i.i.d.) complex Gaussian random variables with zero mean and variance of N_0 , i.e., $\nu_k \sim \mathcal{CN}(0, N_0)$ for $k = 1, \dots, N_r$, thus the average SNR at the transmitter side is described as $\eta = \frac{E_s}{N_0} = \frac{1}{N_0}$.

A slow-fading MIMO channel is assumed which includes a fixed (line of sight) and a variable (non-line of sight) part. The general channel model can be shown as

$$\mathbf{H} = \sqrt{\frac{K}{K+1}} \bar{\mathbf{H}} + \sqrt{\frac{1}{K+1}} \tilde{\mathbf{H}}, \quad (2.2)$$

where the square-root terms are the normalization weights with K being the Rician factor. $\bar{\mathbf{H}}$ and $\tilde{\mathbf{H}}$ are the fixed and variable components, respectively. The fixed part $\bar{\mathbf{H}}$ is such that $[\bar{\mathbf{H}}]_{p,q} = \bar{h}_{p,q}$, for $p = 1, \dots, N_r$ and $q = 1, \dots, N_t$. The variable component of the channel matrix, $\tilde{\mathbf{H}}$, consists of possibly correlated complex coefficients that belong to Gaussian random distribution and defined

$$\tilde{\mathbf{H}} = \boldsymbol{\Sigma}_{\mathbf{r}}^{\frac{1}{2}} \mathbf{H}_w \boldsymbol{\Sigma}_{\mathbf{t}}^{\frac{1}{2}T}, \quad (2.3)$$

where $\mathbf{H}_w$ is the independent Rayleigh fading channel matrix whose elements are described as i.i.d. complex Gaussian random variables, i.e., $[\mathbf{H}_w]_{p,q} = h_{wp,q}$, for $p = 1, \dots, N_r$ and $q = 1, \dots, N_t$. $\boldsymbol{\Sigma}_{\mathbf{t}}$ and $\boldsymbol{\Sigma}_{\mathbf{r}}$ are the real-valued and Hermitian symmetric transmit and receive correlation matrices, respectively and defined as $[\boldsymbol{\Sigma}_{\mathbf{t}}]_{i,j} = \sigma_{i,j}^t = \alpha_t^{|i-j|}$ for $i, j = 1, \dots, N_t$ and $[\boldsymbol{\Sigma}_{\mathbf{r}}]_{i,j} = \sigma_{i,j}^r = \alpha_r^{|i-j|}$ for $i, j = 1, \dots, N_r$, with $|\alpha_t|, |\alpha_r| < 1$ being the transmit and receive correlation coefficients respectively.

If we assume that the v -th component, X_v for $v = 1, \dots, M$, from the constellation alphabet, is transmitted, also the activated antenna indices are given as (u_1, u_2) , (2.1) can also be written as

$$\mathbf{y} = \mathbf{h}_{u_1} \text{Re}(\mathbf{X}_v) + \mathbf{h}_{u_2} \text{Im}(\mathbf{X}_v) + \boldsymbol{\nu}, \quad (2.4)$$

where $\mathbf{h}_{u_1}$ and $\mathbf{h}_{u_2}$ are the u_1 -th and u_2 -th columns of the channel matrix of $\mathbf{H}$, respectively, for $u_1, u_2 = 1, \dots, N_t$. Given the signal model, the receiver uses the maximum likelihood (ML) detection rule to detect the transmit antennas and the chosen symbol from the constellation.

2.2.2. Precoded Quadrature Spatial Modulation

This section introduces a precoding technique for the QSM-MIMO system to enhance its robustness against channel-induced distortions, such as LOS components and spatial correlations among transmit antennas. This approach involves multiplying the transmitted symbol vector $\mathbf{x}$, as defined previously, by a diagonal precoding matrix, given by

$$\boldsymbol{\Psi} = \text{Diag} \begin{bmatrix} \psi_1 & \psi_2 & \dots & \psi_{N_t} \end{bmatrix} = \text{Diag} \begin{bmatrix} a_1 e^{j\theta_1} & a_2 e^{j\theta_2} & \dots & a_{N_t} e^{j\theta_{N_t}} \end{bmatrix}, \quad (2.5)$$

where for non-modified average power budget, the coefficients in the precoder matrix must fulfill the following situation

$$\frac{1}{N_t} \sum_{\ell=1}^{N_t} |\psi_\ell|^2 = \frac{1}{N_t} \sum_{\ell=1}^{N_t} a_\ell^2 = 1. \quad (2.6)$$

The design of the precoder is grounded in identifying the optimal $N_t - 1$ phase rotation values, under the assumption that symbols transmitted from the first antenna experience no rotation. Additionally, N_t optimal amplitude coefficients are determined. There are also more straightforward implementations, such as using only phase rotations and setting $a_l = 1$ for all $l = 1, \dots, N_t$. This approach reduces the search space for optimizing the precoding matrix and prevents power fluctuations by avoiding amplitude scaling. Consequently, the received signal is

$$\mathbf{y} = \mathbf{H}\boldsymbol{\Psi}\mathbf{x} + \boldsymbol{\nu}. \quad (2.7)$$

If it is supposed that the v -th component of the constellation, X_v for $v = 1, \dots, M$, is transmitted as before, Equation (2.4) can be rewritten as

$$\begin{aligned} \mathbf{y} &= \mathbf{h}_{u_1} \operatorname{Re}(X_v \psi_{u_1}) + \mathbf{h}_{u_2} \operatorname{Im}(X_v \psi_{u_2}) + \nu \\ &= \mathbf{h}_{u_1} a_{u_1} \operatorname{Re}(X_v) e^{j\theta_{u_1}} + \mathbf{h}_{u_2} a_{u_2} \operatorname{Im}(X_v) e^{j\theta_{u_2}} + \nu, \end{aligned} \quad (2.8)$$

where a_{u_1} and a_{u_2} are the amplitude scaling coefficients and θ_{u_1} and θ_{u_2} are the rotation of the symbols for the transmitter antenna indices of u_1 and u_2 respectively. The receiver uses the ML detection rule to estimate the most probable channels and symbol index pair $(u_1, u_2, v)_{\text{ML}}$ given as

$$(u_{1\text{ML}}, u_{2\text{ML}}, v_{\text{ML}}) = \arg \min_{u_1, u_2, v} D(\mathbf{y} | \mathbf{h}_{u_1}, \mathbf{h}_{u_2}, X_v, \Psi), \quad (2.9)$$

where $D(\mathbf{y} | \mathbf{h}_{u_1}, \mathbf{h}_{u_2}, X_v, \Psi) = \|\mathbf{y} - \mathbf{h}_{u_1} a_{u_1} \operatorname{Re}(X_v) e^{j\theta_{u_1}} - \mathbf{h}_{u_2} a_{u_2} \operatorname{Im}(X_v) e^{j\theta_{u_2}}\|^2$, is the distance metric, which the receiver tries to find the indices which make this metric minimum.

2.3. Performance Analysis

The primary objective in designing the precoding matrix Ψ is to determine the optimal values of ψ_ℓ or, equivalently, the rotation angles θ_ℓ 's for $\ell = 2, 3, \dots, N_t$ and the amplitude coefficients a_ℓ for $\ell = 1, 2, 3, \dots, N_t$. We consider a time-varying fading channel with no channel state information at the transmitter. Therefore, our design focuses on minimizing the ABEP, which is optimized for high SNR values.

In this section, we calculate the pairwise error probability (PEP) to approximate the ABEP using the well-known upper bound described in [86]

$$\bar{P}_b \leq \frac{1}{N_t^2 M} \sum_{u_1=1}^{N_t} \sum_{\hat{u}_1=1}^{N_t} \sum_{u_2=1}^{N_t} \sum_{\hat{u}_2=1}^{N_t} \sum_{v=1}^M \sum_{\hat{v}=1}^M \frac{N(u_1, \hat{u}_1, u_2, \hat{u}_2, v, \hat{v})}{\log_2(N_t^2 M)} \bar{P}_s, \quad (2.10)$$

where $N(u_1, \hat{u}_1, u_2, \hat{u}_2, v, \hat{v})$ represents the number of bit errors between the respective channel and symbol index pairs $(\mathbf{h}_{u_1}, \mathbf{h}_{u_2}, X_v)$ and $(\mathbf{h}_{\hat{u}_1}, \mathbf{h}_{\hat{u}_2}, X_{\hat{v}})$. The term $\log_2(N_t^2 M)$ denotes the total number of bits used to identify the active antennas and the symbol. $\bar{P}_s$ is the average pairwise error probability (APEP) for detecting the channel and symbol pair $(\mathbf{h}_{\hat{u}_1}, \mathbf{h}_{\hat{u}_2}, X_{\hat{v}})$ instead of the correct pair $(\mathbf{h}_{u_1}, \mathbf{h}_{u_2}, X_v)$.

To compute the ABEP upper bound, it is essential to calculate the PEPs to determine the APEP, as

$$\overline{P}_s = E_{\mathbf{H}} \left[Q \left(\sqrt{\frac{\|\mathbf{z}\|^2}{2}} \right) \right], \quad (2.11)$$

where $\mathbf{z}$ is a $N_r \times 1$ vector as described in [2], with its mean vector $\mathbf{m}_{\mathbf{z}}$, calculated as

$$\mathbf{m}_{\mathbf{z}} = \sqrt{\frac{K}{N_0(K+1)}} (\overline{\mathbf{h}}_{u_1} a_{u_1} \text{Re}(X_v) e^{j\theta_{u_1}} + \overline{\mathbf{h}}_{u_2} a_{u_2} \text{Im}(X_v) e^{j\theta_{u_2}} - \overline{\mathbf{h}}_{\hat{u}_1} a_{\hat{u}_1} \text{Re}(X_{\hat{v}}) e^{j\theta_{\hat{u}_1}} - \overline{\mathbf{h}}_{\hat{u}_2} a_{\hat{u}_2} \text{Im}(X_{\hat{v}}) e^{j\theta_{\hat{u}_2}}), \quad (2.12)$$

and the covariance matrix of the $\mathbf{z}$ vector, is determined by $\mathbf{\Lambda}_{\mathbf{z}}$ and calculated as

$$\mathbf{\Lambda}_{\mathbf{z}} = \frac{1}{N_0(K+1)} \left(|\text{Re}(X_v \psi_{u_1})|^2 + |\text{Im}(X_v \psi_{u_2})|^2 + |\text{Re}(X_{\hat{v}} \psi_{\hat{u}_1})|^2 + |\text{Im}(X_{\hat{v}} \psi_{\hat{u}_2})|^2 - 2(\sigma_{u_1, \hat{u}_1}^t \text{Re}(X_v \psi_{u_1}) \text{Re}(X_{\hat{v}} \psi_{\hat{u}_1}) - \sigma_{u_2, \hat{u}_2}^t \text{Im}(X_v \psi_{u_2}) \text{Im}(X_{\hat{v}} \psi_{\hat{u}_2})) \right) \mathbf{\Sigma}_{\mathbf{r}}, \quad (2.13)$$

where $\sigma_{i,j}^t$ is the (i,j) -th entry of the transmit correlation matrix $\mathbf{\Sigma}_{\mathbf{t}}$. By using the moment generating function, the APEP term can be defined as

$$\overline{P}_s = \frac{1}{\pi} \int_0^{\frac{\pi}{2}} \frac{\exp(-\mathbf{m}_{\mathbf{z}}^\dagger [\mathbf{\Lambda}_{\mathbf{z}} + 4 \sin^2 \theta \mathbf{I}]^{-1} \mathbf{m}_{\mathbf{z}})}{|\frac{\mathbf{\Lambda}_{\mathbf{z}}}{4 \sin^2 \theta} + \mathbf{I}|} d\theta. \quad (2.14)$$

In the case of NLOS channels where $K = 0$, the exponential term in (3.18) vanishes and the computation is possible. When the channel has a LOS component we should follow the upper bound computation for APEP, this equation becomes a single definite integral that can be evaluated numerically for any given mean vector and covariance matrix. This makes the computation of the APEP possible for any constellation and configuration without resorting to Monte Carlo methods. Once the APEPs for all channel and symbol pairs have been calculated, the tight ABEP upper bound can be obtained as shown in (2.10). This process is relatively straightforward and allows for the determination of optimal precoding coefficients through numerical search methods.

2.4. Precoder Optimization

As mentioned previously, because the transmitter knows only the channel statistics and the correlation structure among the transmit antennas and does not have any knowledge about the complete information of the slow-fading and time-varying channel, channel-aware optimization of the precoder is not possible. However, a feasible approach can be minimizing the ABEP upper bound at enough high SNR values. Our attempt for the proposed work is to find the optimum phase rotation values and amplitude coefficients utilizing numerical search methods on the spaces of phase coefficients and amplitude values. By considering slow-fading channels with constant statistical properties during the transmission, the amplitude and phase coefficients are optimized so as to minimize the ABEP only once given the channel statistics and the transmit correlation structure. So we are not able to obtain analytical closed-form solutions for the optimum precoding coefficients and that is why the approach in this study is to compute the precoding coefficients using numerical search algorithms over discretized phase and amplitude coefficient spaces. To understand how to obtain the optimum values for minimizing the asymptotic ABEP, we first start with 2×2 QSM-MIMO system where QPSK symbols are transmitted. In order to make minimization with respect to only phase, we assume $\theta_1 = 0$ and iterate θ_2 with some intervals and obtain the optimum θ_2 value that yields the best error performance. For the minimization with respect to amplitudes for the same system, the amplitudes should satisfy the power constraint equation as $a_1^2 + a_2^2 = 2$. This equation implies that a_1 or a_2 can take $\sqrt{2}$ as the maximum amplitude scaling factor. For 4×4 QSM system the search space and the complexity of the optimization problem will increase, which is why in this scenario only phase rotation precoding is done to simplify the model and prevent optimization complexity. The optimum precoding coefficients are measured at specific high SNR values because theoretical results are very close to the simulation results in high SNR regions. Different θ and a values are obtained for various α_t coefficients in Rayleigh and Rician fading channels. After finding the optimum rotation and amplitude scaling values, it is time for comparing the performances of PQSM systems and conventional systems which is presented in the next section.

2.5. Simulation Results

This section presents simulation results demonstrating the close alignment between theoretical ABEP upper bounds and simulated performances, as well as the performance improvements provided by the proposed technique over benchmark approaches. The results are shown for a conventional and precoded QSM-MIMO system under correlated and uncorrelated Rayleigh and Rician fading scenarios, compared to an SM system with the same spectral efficiency to highlight the gains achieved by the proposed methodology. The fixed part of the channel, $\bar{\mathbf{H}}$, is assumed to be an all-one matrix, while $\mathbf{H}_w$ represents the independent Rayleigh fading channel matrix, with elements described as i.i.d complex Gaussian random variables. The Rician factor is set to $K = 3$, and natural binary code is used in all simulations. There is no assumed correlation at the receiver side $\alpha_r = 0$. For correlated scenarios, a high correlation at the transmitter side is assumed $\alpha_t = 0.9$ to better illustrate the improvements from the precoding approach. Additionally, the 4-QAM used in this chapter for QSM is actually a rotated version of QPSK by $\pi/4$, as QPSK cannot be directly used in QSM. This is because the transmitted symbols must always have non-zero real and imaginary parts to detect the transmitting antenna, as sending zero provides no information for antenna detection. The results show that the theoretical upper bounds are in close agreement with the simulated ABER performances, particularly at moderate to high SNR values across all channel scenarios.

Figures 2.1 and 2.2 illustrate the performance of the 2×2 QSM system and compare it with the precoded results obtained through independent phase rotations and amplitude scaling. In Figure 2.1, which depicts the results for the Rayleigh channel scenario, precoded results for the uncorrelated transmit antenna case are not shown, as the precoding approach does not provide significant improvement in these systems. From the figure, it is evident that a correlation coefficient of $\alpha_t = 0.9$ degrades performance by approximately 6 dB at an ABER level of 1×10^{-5} . While the performance improvement from precoding does not match the uncorrelated case, there is a 3-4 dB enhancement with both amplitude scaling and phase rotation. The performance difference between phase rotation and amplitude scaling is minimal, suggesting that phase rotation alone can be applied for simplicity in implementation.

Figure 2.2 presents the performance comparison for the same 2×2 QSM system over a Rician fading channel. The Rician channel characteristics degrade the performance of SM, so we can also expect performance improvements in the uncorrelated case by diversifying the antennas. With amplitude scaling, the performance significantly improves, even surpassing the performance of the amplitude-scaled uncorrelated case. The amplitude-scaled QSM system enhances the conventional system's performance by approximately 16 dB in highly correlated channel conditions. Phase rotation also yields substantial improvements, especially with a 0.9 correlation, though it is less effective than amplitude scaling. Essentially, correlation becomes advantageous with amplitude scaling and phase rotation precoding. In the Rayleigh case, phase rotation was preferred for its ease of implementation since the performance difference between phase rotation and amplitude scaling was minimal. However, this is not the case for the Rician scenario. Even in the uncorrelated case, amplitude scaling results in an approximately 5 dB improvement compared to the conventional and uncorrelated systems.

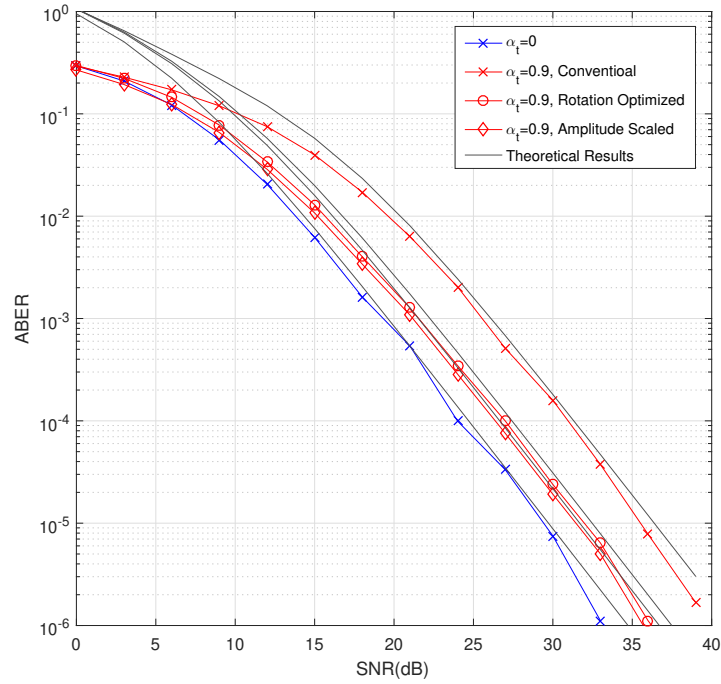


Figure 2.1. ABER of standard and precoded 2×2 QSM-MIMO systems using QPSK modulation over Rayleigh channel.

Figures 2.3 and 2.4 illustrate the performances of precoded and conventional 4×4 QSM-MIMO systems using a 4-QAM constellation (labeled as QPSK in the legends). These results are compared with conventional and precoded SM systems with the same spectral efficiency of $R = 6$. The precoded results, marked by diamond symbols, represent rotation-based optimized performances. The solid gray lines indicate the theoretical results described in the system models, while the colored lines depict the outcomes from Monte Carlo simulations. To simplify the optimization problem and clarify the results, amplitude-scaled optimization is not included in this analysis.

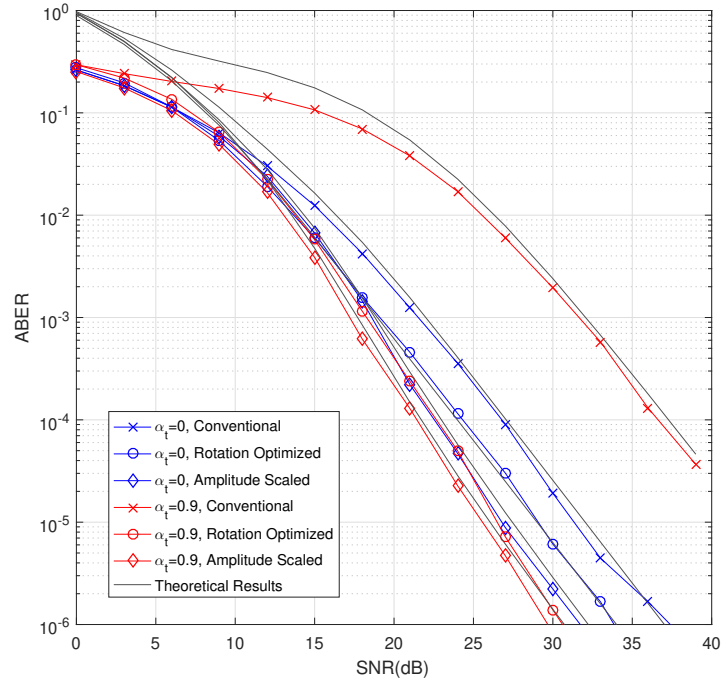


Figure 2.2. ABER of conventional and precoded 2×2 QSM-MIMO systems using QPSK over Rician channel with $K = 3$.

In Figure 2.3, which shows the results for the Rayleigh channel scenario, the QSM system performs 3 dB better than the SM system at an ABER level of 1×10^{-5} for the uncorrelated transmit antenna scenario. For highly correlated Rayleigh channels, the conventional SM system lags 1.5 dB behind the QSM-MIMO transceiver, with this improvement increasing to 3 dB when the constellations are optimally rotated. Precoded results are not shown for the uncorrelated Rayleigh case as the precoding approach does not significantly enhance these systems.

Figure 2.4 presents the performance improvements achieved by transmitting the real and imaginary parts of the symbols independently and using the precoding approach in the Rician fading channel. For the uncorrelated Rician channel, the conventional SM system's performance is 1 dB worse than the conventional QSM system, while its precoded performance is 3 dB worse than the PQSM scenario. In cases of highly correlated transmit antennas, the benefits of the constellation rotation-based precoding approach become more pronounced. In this scenario, the performance gap between conventional and precoded SM and QSM systems is approximately 6 dB at an ABER level of 1×10^{-5} . In summary, QSM already outperforms SM at high SNR levels even without precoding, and precoding further amplifies the performance differences between QSM and SM.

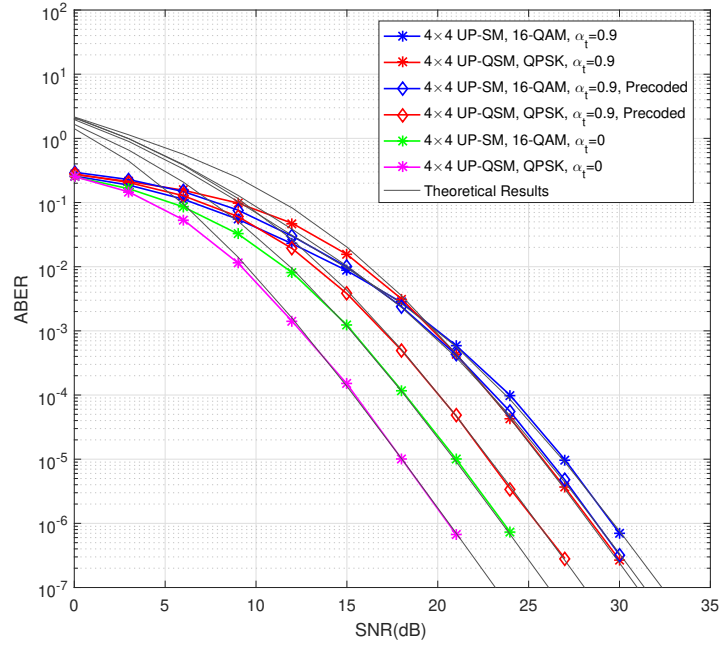


Figure 2.3. ABER of conventional and precoded QSM with $R = 6$ b/s/Hz, over Rayleigh channel.

2.6. Conclusion and Future Work

In summary, the QSM system surpasses the SM system in uncorrelated Rayleigh channels when both have equal spectral efficiency. However, as the channel characteristics become more Rician (with higher K values), the QSM schemes lose their advantage,

especially at low SNR levels. Moreover, in highly correlated or deterministic channels, prioritizing modulation bits over spatial bits becomes more beneficial. Precoding significantly enhances the performance of QSM systems in correlated and Rician channels. The main advantage of the proposed method is that it improves performance with only a minimal increase in transmitter-receiver complexity, without needing explicit channel information at the transmitter. The required information includes the channel's fixed component, the Rician factor, and the correlation between transmit antennas. This straightforward implementation can lead to substantial performance gains, particularly in Rician fading scenarios.

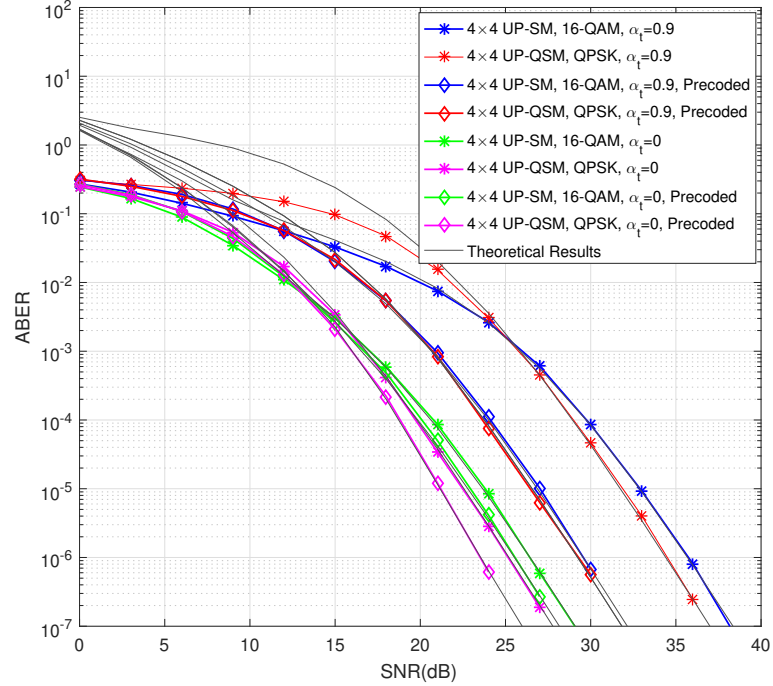


Figure 2.4. Performance of conventional and PQSM with $R = 6$ b/s/Hz, over Rician channels with $K = 3$.

Additionally, in specific amplitude scaling situations, inducing correlation can further boost performance. For simplicity and ease of optimization, using phase rotation alone instead of amplitude scaling can be considered, with a performance trade-off of about 1 dB in Rayleigh channels and some uncorrelated Rician channels. However, it is generally observed that the performance difference between amplitude scaling and phase rotation increases as correlation rises in Rician channels.

3. PRECODING FOR DUAL-POLARIZED QUADRATURE SPATIAL MODULATION

3.1. Introduction

Index modulation techniques such as SM, SSK and QSM can be significantly impacted by spatial correlation effects, which depend on antenna separation [87]. To mitigate performance degradation due to spatial correlation, adequate separation between antennas is required, but this can be challenging in compact devices due to space constraints. To address space limitations, the implementation of dual-polarized (DP) antennas has been explored [15–20]. DP antennas transmit data streams using two orthogonal polarizations simultaneously, reducing the need for large antenna separations by replacing two uni-polarized (UP) antennas with one DP antenna. This not only circumvents spatial correlation but also conserves space. Additionally, employing polarization diversity enhances channel capacity, with studies showing DP systems outperform their UP counterparts under non-ideal conditions [19], [21], [22]. Despite these benefits, DP antennas introduce cross-polar discrimination (XPD) and polarization correlation (XPC). XPD arises from polarization mismatches and varies with channel environments, potentially affecting diversity [24–26]. XPC results from polarization imperfections, typically has minimal practical impact, and is often overlooked [16], [18], [23]. Further discussion on advanced MIMO schemes using DP antennas can be found in [88–90].

Leveraging the benefits of QSM and polarization, this study proposes a new transmission scheme that extends QSM to the polarization domain using DP antennas. In DP-QSM, two transmit antenna indices (two polarization indices) are selected at each signaling instant, one for the real part and one for the imaginary part of the modulation symbol. This method enhances spectral efficiency by $\log_2(2M)$, where M is the number of DP transmit antennas, compared to traditional DP-SM. The use of orthogonal carriers ensures the preservation of the no-ICI structure in DP-QSM. For signal detection, the ML method is used, and an upper bound for the ABEP of the DP-QSM

system is derived and validated through Monte Carlo simulations. Results indicate that DP-QSM not only improves spectral efficiency but also system performance, reducing the need for large inter-antenna separation, particularly in line-of-sight (LOS) channel scenarios. However, the combination of antenna and signal spaces for information transmission introduces a significant drawback in conventional SM and QSM systems: the unreliable detection of bits assigned to the antenna index, leading to a substantial increase in the overall ABER of the SM/QSM transceiver. This issue, known as the unequal error vulnerability of symbol and antenna bits against channel fading and spatial correlations, has been extensively studied to enhance system robustness [8–10]. These methods typically increase system complexity and latency due to added encoder-decoder pairs. Subsequent research has proposed a precoding approach that does not require precise channel coefficients at the transmitter, suitable for time-varying channels, and preserves the average power budget [29], [30], [32], [33], [82], [83]. This method, involving phase rotation and/or amplitude scaling of transmitted symbols, is designed to minimize ABER in high SNR conditions, significantly enhancing system performance against spatial correlations and fading effects. Previously, QSM techniques predominantly utilized UP antennas. This study extends the previously discussed precoding approach to a DP-QSM system, developing a novel scheme that effectively compensates for the unequal error vulnerability of the antenna index bits, particularly under conditions of Rician fading and/or high inter-antenna correlations at the transmitter. This method removes the need for complex coding/decoding techniques to achieve unequal error protection. The no ICI structure is maintained in this model because the precoding operation simply involves multiplying the transmitted signal vector by a diagonal matrix, eliminating the need for explicit CSI. Additionally, this approach does not alter the average transmit power budget and requires only a slight modification to the transmitter circuitry. The optimization strategy focuses on minimizing the ABER in high SNR conditions, where optimal rotation values are determined. Both theoretical analysis and simulation results suggest that the precoded DP-QSM system provides considerable robustness against spatial correlations and the Rician fading scenario compared to the standard DP-QSM system. Thus, integrating the benefits of QSM, polarization dimension, and precoding techniques could lead to the design of QSM-MIMO systems that are more resilient to Rician fading and spatial correlations at the transmitter side.

3.2. System Model

3.2.1. Conventional DP-QSM

We consider a QSM system using M and N DP antennas at the transmitter and receiver sides respectively. The number of transmit antennas M is assumed to be an integer power of 2 for simplicity, so $M = 2^m$, also we consider an L -ary constellation, with ℓ bits assigning to constellation $\mathcal{X} = \{X_1, \dots, X_k, \dots, X_L\}$ which consists of L symbols so $\ell = \log_2 L$. At each transmission instant, $2(m+1) + \ell$ bits are divided into three groups: the first two groups consist of $m+1$ bits which are used to select indices of the two active polarizations of antennas, u_1 and u_2 where $u_1, u_2 = 1, \dots, 2M$. The remaining ℓ bits are used to select the one modulation symbol to be transmitted. Suppose the selected symbol is X_v for $v = 1, \dots, L$, then the u_1 -th antenna transmits the real part of X_v whereas u_2 -th antenna sends imaginary part of X_v . Notice that the spectral efficiency of the system is $R = 2(m+1) + \ell = \log_2((2M)^2 L)$ b/s/Hz which is $m+1$ b/sec/Hz higher than that of DP-SM with the same transmission parameters. The transmitted signal vector is $\mathbf{x} = [0, \dots, x_{u_1}, \dots, x_{u_2}, \dots, 0]^T$, where only one or two entries are non-zero, based on the selected antenna polarization dimension. In other words, the position of the non-zero elements denotes the active antenna indices and their values indicate the real and imaginary parts of the transmitted symbol, i.e., $x_{u_1} = \text{Re}(X_v)$ and $x_{u_2} = \text{Im}(X_v)$, and a power constraint of unity, i.e., $E_{\mathbf{x}} = [\mathbf{x}^H \mathbf{x} = 1]$ is assumed.

Accordingly, the received signal model is

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \nu, \quad (3.1)$$

where $\mathbf{y}$ and ν are $2N \times 1$ received signal and noise vectors respectively, $\mathbf{H}$ is the channel matrix, and $\mathbf{x}$ is a $2M \times 1$ transmitted signal vector. ν has independent and identically distributed (i.i.d.) complex Gaussian random variables with zero mean and variance of N_0 , i.e., $\nu_k \sim \mathcal{CN}(0, N_0)$ for $k = 1, \dots, 2N$, thus the average SNR at the transmitter side is described as $\eta = \frac{E_s}{N_0} = \frac{1}{N_0}$.

Channel matrix $\mathbf{H}$ can be defined as a $2N \times 2M$ dimensional matrix as

$$\mathbf{H} = \begin{bmatrix} h_{1,1} & h_{1,2} & \cdots & h_{1,(2M)} \\ h_{2,1} & h_{2,2} & \cdots & h_{2,(2M)} \\ h_{3,1} & h_{3,2} & \cdots & h_{3,(2M)} \\ h_{4,1} & h_{4,2} & \cdots & h_{4,(2M)} \\ \vdots & \vdots & \vdots & \vdots \\ h_{(2N-1),1} & h_{(2N-1),2} & \cdots & h_{(2N-1),(2M)} \\ h_{(2N),1} & h_{(2N),2} & \cdots & h_{(2N),(2M)} \end{bmatrix}, \quad (3.2)$$

where in this model, the polarization direction and antenna indices are combined into one channel index for ease of notation. We consider a slow-fading MIMO channel that comprises both a fixed (line-of-sight) component and a variable (non-line-of-sight) component. The general model for this channel can be expressed as

$$\mathbf{H} = \sqrt{\frac{K}{K+1}} \bar{\mathbf{H}} + \sqrt{\frac{1}{K+1}} \tilde{\mathbf{H}}, \quad (3.3)$$

where the square-root terms represent the normalization weights, with K being the Rician factor. $\bar{\mathbf{H}}$ and $\tilde{\mathbf{H}}$ are the fixed and variable components, respectively. The fixed part $\bar{\mathbf{H}}$ is such that $[\bar{\mathbf{H}}]_{p,q} = \bar{h}_{p,q}$, for $p = 1, \dots, 2N$ and $q = 1, \dots, 2M$. The variable component of the channel matrix, $\tilde{\mathbf{H}}$, consists of potentially correlated complex coefficients that follow a Gaussian random distribution and is defined

$$\tilde{\mathbf{H}} = \Sigma_{\mathbf{r}}^{\frac{1}{2}} \mathbf{H}_w \Sigma_{\mathbf{t}}^{\frac{1}{2}T}, \quad (3.4)$$

where $\mathbf{H}_w$ is the independent Rayleigh fading channel matrix whose elements are described as i.i.d. complex Gaussian random variables, i.e., $[\mathbf{H}_w]_{p,q} = h_{w,p,q}$, for $p = 1, \dots, 2N$ and $q = 1, \dots, 2M$. $\Sigma_{\mathbf{t}}$ and $\Sigma_{\mathbf{r}}$ are the real-valued and Hermitian symmetric transmit and receive correlation matrices, respectively, and defined as $[\Sigma_{\mathbf{t}}]_{i,j} = \sigma_{i,j}^t = \alpha_t^{|i-j|}$ for $i, j = 1, \dots, 2M$ and $[\Sigma_{\mathbf{r}}]_{i,j} = \sigma_{i,j}^r = \alpha_r^{|i-j|}$ for $i, j = 1, \dots, 2N$, with α_t and α_r being the transmit and receive correlation coefficients respectively. In addition by assuming the polarization correlations which are coming from the polarization mismatch the general channel model becomes as

$$\mathbf{H} = \sqrt{\frac{K}{1+K}} (\mathbf{1}_{N \times M} \otimes \mathbf{\Gamma}_f) \odot \bar{\bar{\mathbf{H}}} + \sqrt{\frac{1}{1+K}} (\mathbf{1}_{N \times M} \otimes \mathbf{\Gamma}_v) \odot \left(\Phi_{\mathbf{r}}^{\frac{1}{2}} \mathbf{H}_w \Phi_{\mathbf{t}}^{\frac{1}{2}} \right), \quad (3.5)$$

where $\bar{\bar{\mathbf{H}}}$ is the fixed part of the channel, where its elements are assumed to be all one for simplicity. $\mathbf{1}_{N \times M}$ is an $N \times M$ matrix with all one elements and $\otimes$ stands

for the Kronecker product. $\Phi_{\mathbf{t}}$ and $\Phi_{\mathbf{r}}$ are $2M \times 2M$ and $2N \times 2N$ real-valued and Hermitian symmetric transmit and receive correlation matrices respectively, defined as $\Phi_{\mathbf{r}} = \Sigma_{\mathbf{r}} \otimes \Pi_{\mathbf{r}}$, $\Phi_{\mathbf{t}} = \Sigma_{\mathbf{t}} \otimes \Pi_{\mathbf{t}}$, with $\Pi_{\mathbf{r}}$ and $\Pi_{\mathbf{t}}$ being the polarization correlation matrices at the receiver and the transmitter sides respectively. If we assume that the v -th component of X_v for $v = 1, \dots, L$, from the constellation alphabet, is transmitted, also the activated antenna indices are given as (u_1, u_2) , (4.1) can also be written as

$$\mathbf{y} = \mathbf{h}_{u_1} \text{Re}(X_v) + \mathbf{h}_{u_2} \text{Im}(X_v) + \nu, \quad (3.6)$$

where $\mathbf{h}_{u_1}$ and $\mathbf{h}_{u_2}$ are the u_1 -th and u_2 -th columns of the channel matrix of $\mathbf{H}$, respectively, for $u_1, u_2 = 1, \dots, 2M$ and X_v is the v -th element of the constellation X for $v = 1, \dots, L$. Given the signal model, the receiver uses the ML detection rule to detect the transmit antennas and polarization directions and the chosen symbol from the constellation. Receiver estimates the most probable channels and symbol index pair of $(u_1, u_2, v)_{\text{ML}}$ given as

$$(u_{1\text{ML}}, u_{2\text{ML}}, v_{\text{ML}}) = \arg \min_{u_1, u_2, v} D(\mathbf{y} | \mathbf{h}_{u_1}, \mathbf{h}_{u_2}, X_v), \quad (3.7)$$

where $D(\mathbf{y} | \mathbf{h}_{u_1}, \mathbf{h}_{u_2}, X_v)$, is the distance metric, which the receiver tries to find the indices which make this metric minimum and defined as

$$D(\mathbf{y} | \mathbf{h}_{u_1}, \mathbf{h}_{u_2}, X_v) = \|\mathbf{y} - \mathbf{h}_{u_1} \text{Re}(X_v) - \mathbf{h}_{u_2} \text{Im}(X_v)\|^2. \quad (3.8)$$

Once the most likely channel indices are found it is easy to detect the indices of the most likely antennas and polarization directions by the use of a mapping plan in (3.2).

3.2.2. Precoded DP-QSM

The precoding approach is based on rotating the symbol constellations according to the selected polarization and the active transmit antennas for signal transmission. This means multiplying the transmitted symbol vector $\mathbf{x}$ as defined in the previous part with a diagonal precoding matrix defined as

$$\Psi = \mathbf{Diag} \begin{bmatrix} \psi_1 & \psi_2 & \dots & \psi_{2M} \end{bmatrix} = \mathbf{Diag} \begin{bmatrix} e^{j\theta_1} & e^{j\theta_2} & \dots & e^{j\theta_{2M}} \end{bmatrix}, \quad (3.9)$$

where θ values are the rotations of the symbol constellation transmitted from antennas. Precoder design in this work is based on the determination of the $2M - 1$ optimum values of the phase rotations because the assumption is that there is no rotation for

symbols transmitted from the first polarization of the first antenna. The corresponding received signal is written as

$$\mathbf{y} = \mathbf{H}\Psi\mathbf{x} + \nu. \quad (3.10)$$

If we suppose that the v -th component of the constellation, X_v for $v = 1, \dots, L$, is transmitted as before, Equation (3.10) can be rewritten as

$$\mathbf{y} = \mathbf{h}_{u_1} \text{Re}(X_v)e^{j\theta_{u_1}} + \mathbf{h}_{u_2} \text{Im}(X_v)e^{j\theta_{u_2}} + \nu, \quad (3.11)$$

where $\mathbf{h}_{u_1}$ and $\mathbf{h}_{u_2}$ identify the u_1 -th and u_2 -th columns of $\mathbf{H}$ respectively. Given the signal model in this part, the receiver uses the ML detection rule to estimates the most probable channels and symbol index pair of $(u_1, u_2, v)_{\text{ML}}$ given as

$$(u_{1\text{ML}}, u_{2\text{ML}}, v_{\text{ML}}) = \arg \min_{u_1, u_2, v} D(\mathbf{y}|\mathbf{h}_{u_1}, \mathbf{h}_{u_2}, X_v, \Psi), \quad (3.12)$$

where $D(\mathbf{y}|\mathbf{h}_{u_1}, \mathbf{h}_{u_2}, X_v, \Psi)$ is the distance metric used in the detection and defined as

$$D(\mathbf{y}|\mathbf{h}_{u_1}, \mathbf{h}_{u_2}, X_v, \Psi) = \|\mathbf{y} - \mathbf{h}_{u_1} \text{Re}(X_v)e^{j\theta_{u_1}} - \mathbf{h}_{u_2} \text{Im}(X_v)e^{j\theta_{u_2}}\|^2. \quad (3.13)$$

3.3. Precoder Design and Performance Analysis

The main goal in the design of the precoding matrix Ψ is to calculate the optimum values of θ_u 's which are the rotation values for $u = 2, 3, \dots, 2M$. A time-varying fading channel is considered where there is no available information about the channel at the transmitter, so our design is based on the minimization of the ABEP at high SNR values. The ABEP for SM systems is not available in closed form so it can be approximated by using the well-known upper-bound expression introduced in [91]. This equation can be extended to a DP-QSM system as following

$$\bar{P}_b \leq \frac{1}{(2M)^2 L} \sum_{u_1=1}^{2M} \sum_{\hat{u}_1=1}^{2M} \sum_{u_2=1}^{2M} \sum_{\hat{u}_2=1}^{2M} \sum_{v=1}^L \sum_{\hat{v}=1}^L \frac{N(u_1, \hat{u}_1, u_2, \hat{u}_2, v, \hat{v})}{\log_2((2M)^2 L)} \bar{P}_s, \quad (3.14)$$

where $N(u_1, \hat{u}_1, u_2, \hat{u}_2, v, \hat{v})$ is the number of bits in error between respective channel and symbol index pairs of $(\mathbf{h}_{u_1}, \mathbf{h}_{u_2}, X_v)$ and $(\mathbf{h}_{\hat{u}_1}, \mathbf{h}_{\hat{u}_2}, X_{\hat{v}})$ and $\log_2((2M)^2 L)$ identifies the entire number of bits utilized for detecting the direction of the polarization, active antenna, and symbol. $\bar{P}_s$ is APEP. The necessary part of computing the ABEP

upper bound is to calculate the PEPs and the expected values as shown in the following. PEP identifies the probability of error for detecting the channel and symbol pair $(\mathbf{h}_{\hat{u}_1}, \mathbf{h}_{\hat{u}_2}, X_{\hat{v}})$ instead of the exact pair of $(\mathbf{h}_{u_1}, \mathbf{h}_{u_2}, X_v)$. APEP can be calculated as

$$\bar{P}_s = E \left[Q \left(\sqrt{\frac{\|\mathbf{z}\|^2}{2}} \right) \right], \quad (3.15)$$

where $\mathbf{z}$ is a $2N \times 1$ vector as described in [2] and defined as

$$\begin{aligned} \mathbf{z} &= \frac{1}{\sqrt{N_0}} (\mathbf{h}_{u_1} \text{Re}(X_v) e^{j\theta_{u_1}} + \mathbf{h}_{u_2} \text{Im}(X_v) e^{j\theta_{u_2}} \\ &\quad - \mathbf{h}_{\hat{u}_1} \text{Re}(X_{\hat{v}}) e^{j\theta_{\hat{u}_1}} - \mathbf{h}_{\hat{u}_2} \text{Im}(X_{\hat{v}}) e^{j\theta_{\hat{u}_2}}). \end{aligned} \quad (3.16)$$

The mean of the $\mathbf{z}$ vector is denoted by $\mathbf{m}_z$ and computed as

$$\begin{aligned} \mathbf{m}_z &= \sqrt{\frac{K}{N_0(K+1)}} (\bar{\mathbf{h}}_{u_1} \text{Re}(X_v) e^{j\theta_{u_1}} + \bar{\mathbf{h}}_{u_2} \text{Im}(X_v) e^{j\theta_{u_2}} - \\ &\quad \bar{\mathbf{h}}_{\hat{u}_1} \text{Re}(X_{\hat{v}}) e^{j\theta_{\hat{u}_1}} - \bar{\mathbf{h}}_{\hat{u}_2} \text{Im}(X_{\hat{v}}) e^{j\theta_{\hat{u}_2}}), \end{aligned}$$

and the covariance matrix of the $\mathbf{z}$ vector which is denoted by Λ_z , computed as

$$\begin{aligned} \Lambda_z &= \frac{1}{N_0(K+1)} ((\Phi_{u_1, u_1}^t | \text{Re}(X_v e^{j\theta_{u_1}})|^2 + \Phi_{u_2, u_2}^t | \text{Im}(X_v e^{j\theta_{u_2}})|^2 + \\ &\quad \Phi_{\hat{u}_1, \hat{u}_1}^t | \text{Re}(X_{\hat{v}} e^{j\theta_{\hat{u}_1}})|^2 + \Phi_{\hat{u}_2, \hat{u}_2}^t | \text{Im}(X_{\hat{v}} e^{j\theta_{\hat{u}_2}})|^2 - \\ &\quad 2(\text{Re}(\Phi_{u_1, \hat{u}_1}^t e^{j(\theta_{u_1} - \theta_{\hat{u}_1})} \text{Re}(X_v) \text{Re}(X_{\hat{v}}^*)) - \\ &\quad 2(\text{Im}(\Phi_{u_2, \hat{u}_2}^t e^{j(\theta_{u_2} - \theta_{\hat{u}_2})} \text{Im}(X_v) \text{Im}(X_{\hat{v}}^*)) \Phi_r), \end{aligned} \quad (3.17)$$

where $\Phi_{i,j}^t$ is the (i,j) -th entry of the transmit correlation matrix Φ_t . By using the Moment generating function, $\bar{P}_s$ in (3.15) becomes as

$$\bar{P}_s = \frac{1}{\pi} \int_0^{\frac{\pi}{2}} \frac{\exp(-\mathbf{m}_z^\dagger [\Lambda_z + 4 \sin^2 \theta \mathbf{I}]^{-1} \mathbf{m}_z)}{|\frac{\Lambda_z}{4 \sin^2 \theta} + \mathbf{I}|} d\theta. \quad (3.18)$$

In NLOS channels where the Rician K factor is zero, the exponential term in the APEP formula disappears, allowing for feasible computation. For channels with a LOS component, it's necessary to compute the APEP using an upper bound method, which involves a definite integral that can be numerically evaluated using any given mean vector and covariance matrix. This approach enables the calculation of the APEP for any constellation and configuration without the need for Monte Carlo simulations. After calculating the APEPs between all combinations of channels and symbols, a tight

upper bound on the ABEP can be derived, facilitating the determination of optimal precoding coefficients through numerical search methods.

To demonstrate how the optimal phase rotation values —derived from minimizing the ABEP upper bound— are determined and to illustrate their effect on simulated ABEP performance over uncorrelated/correlated Rayleigh and Rician channels, we begin by considering a 2×2 DP-QSM MIMO system employing 4-QAM constellations. In this system, symbols are transmitted from the first polarization of the first antenna without rotation ($\theta_1 = 0$), while symbols from other polarizations are rotated using the three optimal values of θ_l for $l = 2, 3, 4$, which are selected from the range of $[0, \pi/2]$. It is assumed that $\Sigma_{\mathbf{r}} = \mathbf{I}$ in all scenarios, as receiver-side correlation does not impact our precoder design. The optimal phase rotation coefficients are determined at high SNR values because theoretical results closely match simulation results in high SNR regions. After identifying the optimal rotation values, we compare the performances of the precoded DP-QSM system with conventional systems, which is discussed in the next section.

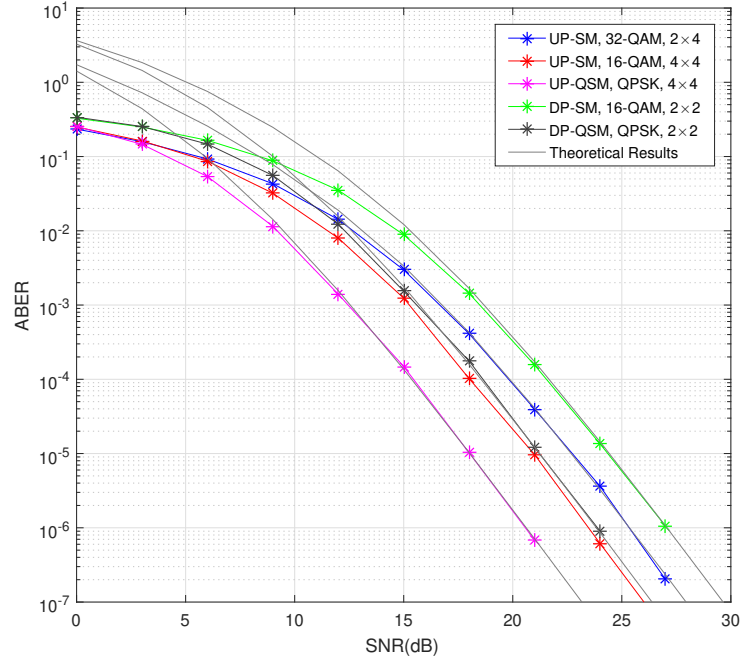


Figure 3.1. Performance of conventional and precoded systems with $R = 6$ b/s/Hz over uncorrelated Rayleigh channel.

3.4. Simulation Results

In this section, we present simulation results showing both the tightness of the theoretical ABEP upper bounds and the efficiency of the proposed precoding technique to compensate for transmit and channel correlations in DP-QSM MIMO systems. As mentioned before $\overline{\mathbf{H}}$ is the fixed part of the channel and it is assumed to be all one matrix and $\mathbf{H}_w$ is the independent Rayleigh fading channel matrix, with elements described as i.i.d complex Gaussian random variables. $[\mathbf{H}_w]_{p,q} = h_{w_{p,q}} \sim \mathcal{CN}(0, 1)$ for $p = 1 \dots 2N$ and $q = 1 \dots 2M$. Regarding simulation settings, the Rician factor is assumed as $K = 3$. Natural binary code is used in all simulations. For all results, we assumed that there is no correlation at the receiver side which means that $\alpha_r = 0$, as the correlation at the receiver side does not affect the precoding design. For correlated scenarios, it is assumed that there is a high correlation at the transmitter side because the improvements of the precoding approach are more obvious in highly correlated channels, for this reason, $\alpha_t = 0.9$ is supposed for correlated cases.

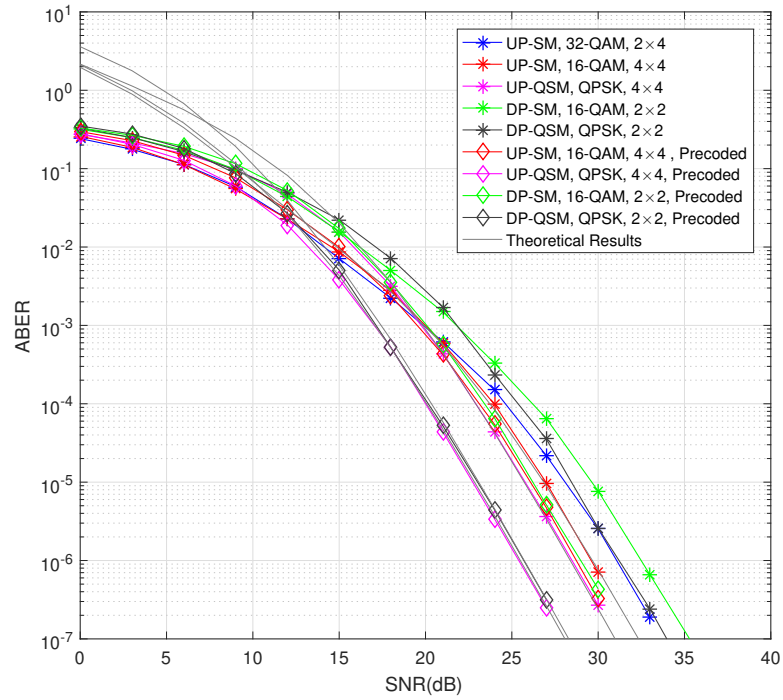


Figure 3.2. Performance of conventional and precoded systems with $R = 6$ b/s/Hz over correlated Rayleigh channel with $\alpha_t = 0.9$.

In all derivations, DP antennas are assumed to have the same polarization characteristics, so that the XPD factors which are defined as the power ratios of the co-polar and cross-polar terms and expressed in fixed and variable parts as χ_f and χ_v respectively, are set to be $\chi_f = 15$ dB and $\chi_v = 5$ dB. The details of XPD parameters are not shown in this work but they have been discussed completely in [92] and [93]. In addition, 4-QAM mentioned in this work for DP-QSM is actually the rotated version of QPSK with $\pi/4$. This scheme is used in this work because QPSK can not be used directly in QSM. In other words, the sent symbols should always have some value other than zero for both the real and imaginary parts to detect the antenna because it is not possible to get some information by sending nothing for the real or imaginary part which is used to detect the transmitter.

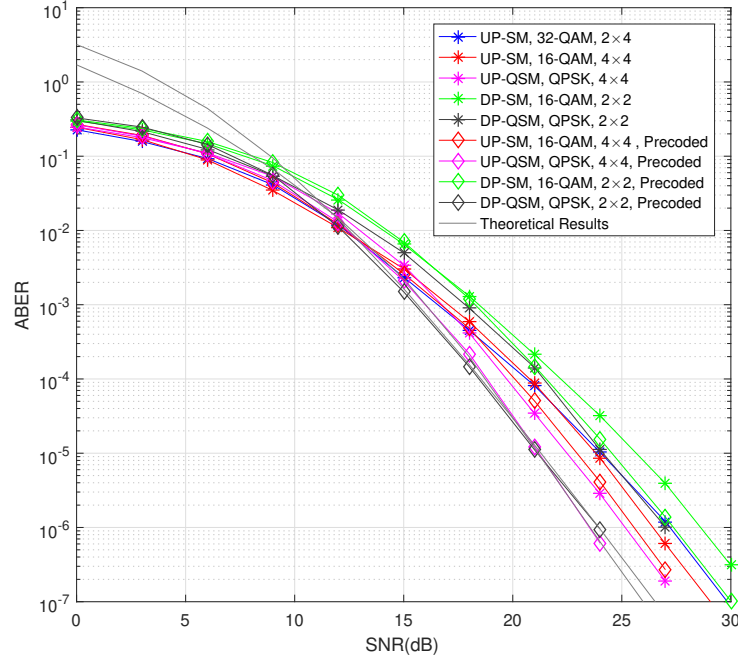


Figure 3.3. Performance of conventional and precoded DP-SM, DP-QSM, UP-SM and UP-QSM with $R = 6$ b/s/Hz over uncorrelated Rician channel with $K = 3$.

The results are performed for precoded and conventional 2×2 DP-QSM systems employing 4-QAM (shown as QPSK in legends) and 16-QAM constellations. The spectral efficiency for these systems are 6 and 8 bpcu respectively and it is compared with the conventional and precoded DP-SM systems and also conventional and precoded 2×4 and 4×4 UP-SM systems with the same spectral efficiency. The results are

shown only for the perfect channel estimation cases. As can be seen from the figures, the theoretical upper bounds are in tight agreement with the simulated ABER performances especially at moderate to high SNR values for all channel scenarios. The solid gray lines in graphs are used to show the theoretical results explained in the system models and only a few of them are represented for the clarity of the results. The advantages of precoded DP-QSM are more clear in highly correlated channels which is why only this scenario is assumed in this work due to the space limitations. We first present the comparison for the spectral efficiency of 6 bpcu, for both uncorrelated and correlated Rayleigh fading scenarios.

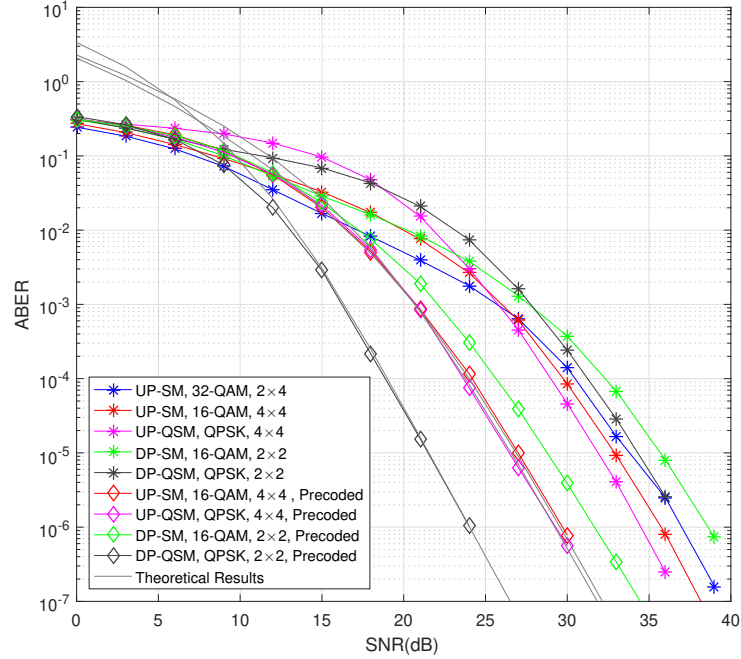


Figure 3.4. Performance of conventional and precoded DP-SM, DP-QSM, UP-SM and UP-QSM with $R = 6$ b/s/Hz over correlated Rician channel with $K = 3$ and $\alpha_t = 0.9$.

From Figure 3.1, it is obvious that DP-QSM performs 3 dB better than conventional DP-SM at 10^{-5} ABER level and it outperforms the equivalent 2×4 UP-SM systems employing 32-QAM. There are no precoded results shown for this case because precoding does not add any benefit to the performance in Rayleigh fading and uncorrelated channel scenarios. The performance improvement by using the precoded system is more obvious in correlated cases. In Figure 3.2, it can be seen that the precoded DP-QSM system outperforms its equivalent conventional DP-SM system by

approximately 6 dB at 1×10^{-5} ABER level and it has the same performance as the precoded UP-QSM system, while it occupies half as much space as this UP system. Also, it is obvious that our proposed system performs better than the precoded DP-SM system with the same spectral efficiency by 3 dB.

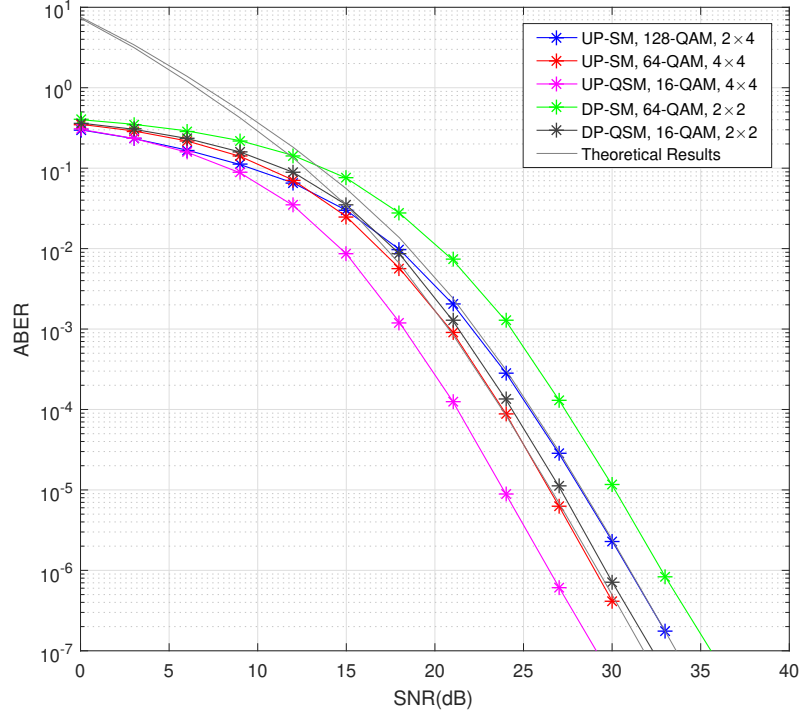


Figure 3.5. Performance of conventional and precoded systems with $R = 8$ b/s/Hz over uncorrelated Rayleigh channel.

For the Rician fading channel, from Figure 3.3, when there is no correlation at the transmitter side the results are almost the same as the correlated Rayleigh channel and the precoded DP-QSM system achieves the same performance as the precoded UP-QSM system. In the case of highly correlated transmit antennas in addition to the line-of-sight channel component, the benefits of the proposed technology are more significant. As we can see from Figure 3.4, the 2×2 precoded DP-QSM system employing 4-QAM outperforms 4×4 precoded UP-QSM system employing 4-QAM by approximately 6 dB and 2×2 precoded DP-SM using 16-QAM constellation by 8 dB, at 1×10^{-6} ABER level. In the next figures the results are shown for higher spectral efficiency of $R = 8$ bpcu and the comparisons for different channel scenarios are presented. In the uncorrelated Rayleigh channel fading case, Figure 3.5, the results

are the same as the results for a spectral efficiency of 6 bpcu. For the Rayleigh faded channel with the correlation coefficient of $\alpha_t = 0.9$ at the transmitter side in Figure 3.6, the precoded DP-QSM achieves the same performance as precoded UP-SM and precoded UP-QSM systems with 4×4 antenna numbers and employing 64-QAM and 16-QAM constellations respectively. It also surpasses the precoded DP-SM system by approximately 3 dB.

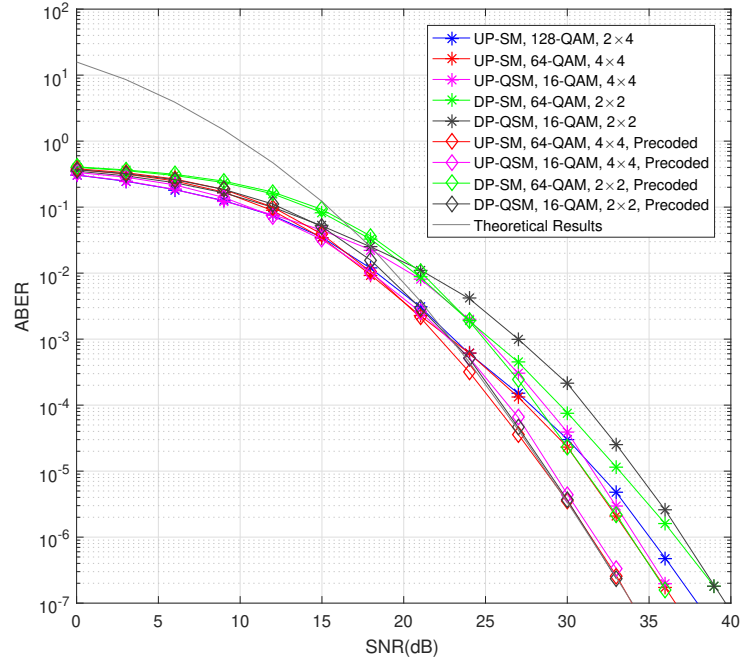


Figure 3.6. Performance of conventional and precoded systems with $R = 8$ b/s/Hz over correlated Rayleigh channel with $\alpha_t = 0.9$.

Figure 3.7 shows the ABER performance comparisons for the previous systems over uncorrelated/correlated Rician channel case where both line-of-sight (LOS) channel component degrades the performance significantly and this degradation of the performance is compensated by precoding. In Figure 3.8, especially when there is a high correlation at the transmitter side it can result that the precoded 2×2 DP-QSM system improves the performance of conventional DP-SM system by 9 dB at 1×10^{-5} ABER level and outperforms its conventional 4×4 UP-SM and 4×4 UP-QSM counterpart by 7 dB. Furthermore, the precoded DP-QSM system outperforms the precoded UP-QSM, and precoded UP-SM systems, while the DP system occupies half as much space as its UP counterpart system.

3.5. Conclusion

In this work, DP-QSM system is introduced, to investigate the effects of extending QSM architecture by polarization dimension and a precoding approach for DP-QSM systems over slow-fading correlated channels is considered to make this structure more robust against channel corruptions and correlation effects. The optimum rotation values are calculated by minimization of ABEP upper bound at large SNR values. Both theoretical analysis and extensive simulations show that the proposed technique has the potential to offer significant performance gains in comparison to conventional DP-SM, DP-QSM, and the equivalent UP-SM systems. In general, the correlated Rician channel model is presented, where uncorrelated and Rayleigh channels are the special cases of this model, by using the well-known Kronecker model.

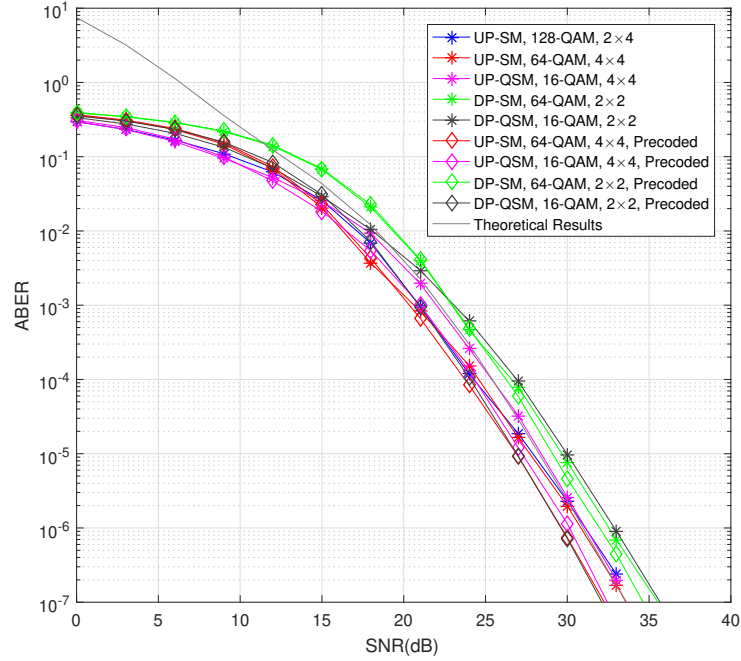


Figure 3.7. Performance of conventional and precoded DP-SM, DP-QSM, UP-SM and UP-QSM with $R = 8$ b/s/Hz over uncorrelated Rician channel with $K = 3$.

By assuming perfect channel estimation, PEP, APEP and ABEP are derived and the optimum values of the precoding coefficients which minimize ABEP have been calculated. Then by using the optimum values we compare the error performance of the conventional systems with the proposed precoded DP-QSM system.

The advantages of the precoding scheme are more obvious for highly correlated systems in addition to Rician fading effects. From the results, it is shown that for $R = 6$ and $R = 8$ b/s/Hz schemes, DP-QSM outperforms DP-SM in Rayleigh uncorrelated fading channel. In other words, if the number of transmit antennas is the same for conventional DP-SM and conventional DP-QSM systems, DP-QSM outperforms DP-SM in uncorrelated Rayleigh channels without precoding with the same spectral efficiency. However, by assuming the line-of-sight channel component, the conventional DP-QSM scheme does not add a significant advantage against conventional DP-SM schemes, especially in low-SNR values where in some cases it is even performing worse than the conventional DP-SM system.

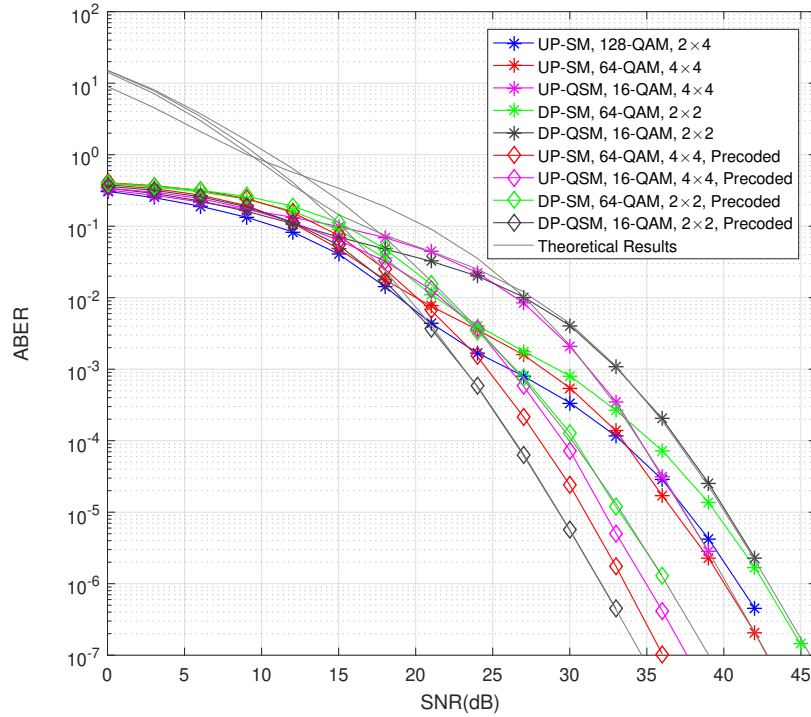


Figure 3.8. Performance of conventional and precoded DP-SM, DP-QSM, UP-SM and UP-QSM with $R = 8$ b/s/Hz over correlated Rician channel with $K = 3$ and $\alpha_t = 0.9$.

By implementing the precoding approach, which is based on rotating the constellation, huge performance improvement can be obtained especially in correlated Rician fading cases. The improvement can be up to 15 dB in the performance as in Figure 3.4. It can be concluded from the results that based on the spectral efficiency and the channel characteristics we can choose which method to use to achieve better perfor-

mances. Both theoretical analysis and simulations show that the precoded DP-QSM system has the potential to offer significant robustness against spatial correlations at the transmitter side and also the Rician fading scenario in comparison to the equivalent DP-SM and UP-SM systems, where the latter occupies twice as much antenna space. Therefore combining the advantages of dual-polarization and precoding techniques may enable the design of QSM-MIMO systems that are more robust to Rician fading and spatial correlations at the transmitter side.



4. DUAL-POLARIZED SIGNATURE SPATIAL MODULATION

4.1. Introduction

SM is an innovative MIMO technique that utilizes both the antenna and signal spaces to transmit information. A strong correlation between transmit antennas can make the channels from different antennas to each receive antenna nearly indistinguishable, which significantly hampers the detection performance of the bits responsible for selecting the active antenna. This issue leads to a substantial rise in the overall ABER of the SM transceiver.

To address this problem, recent strategies have involved the use of channel coding to provide unequal error protection (UEP) as seen in [8–10], or precoding techniques that involve antenna-dependent rotation or a combination of rotation and amplitude scaling of the signal constellations at the transmitter, as suggested in [11]. UEP techniques require the integration of an encoder-decoder pair into the transceiver, significantly increasing latency and computational complexity. On the other hand, precoding, which is based on the multiplication of the transmitted signal vector with a diagonal scaling matrix, has been effective in improving performance for lower-level constellations like BPSK and QPSK, where it significantly increases the minimum Euclidean distance between rotated constellations [11]. However, as the order of the constellation increases, such as in M -ary QAM for $M = 16, 64, \dots$, etc., the minimum distance between constellations decreases and the benefits of precoding become less pronounced. Additionally, the complexity of optimizing precoding grows considerably with the increase in the number of transmit antennas and/or the modulation order. Consequently, while precoding methods are beneficial for small-scale MIMO systems and simpler modulation schemes, they are less effective for applications requiring high spectral efficiency or in situations with significant channel and transmit antenna correlation effects. For this reason, a new approach called signature spatial modulation (SSM) has been developed, as presented in [12]. In this method, the transmitted symbols are selected from

multiple constellations, each uniquely associated with one of the N_t transmit antennas. These constellations, first introduced in [13] and [14], are generated through successive geometric interpolations starting from the primary constellation plane. They are designed to maximize the minimum Euclidean distance between constellations. The study in [12] demonstrates that SSM offers considerable performance improvements, particularly with higher-order modulations, compared to both traditional SM and precoded SM, even under conditions of severe transmit antenna correlations.

Another method to combat antenna or channel correlations involves the use of DP antennas, as their orthogonal polarizations typically provide significant channel separation through substantial decorrelation at both the transmit and receive ends. Studies such as those documented in [17] have demonstrated that DP-SM performs better than uni-polarized systems over correlated Rayleigh and Rician channels. In [94], a precoding-aided and dual-polarized SM approach is introduced, showing further enhancements in the performance of DP-SM. However, similar to conventional SM with uni-polarized antennas, the additional gains from precoding are generally limited to smaller constellations.

To overcome the limitations associated with precoding-based SM, this work integrates the benefits of antenna-dependent signature constellation design with dual-polarization, introducing a signature constellation design technique for SM-MIMO systems that utilize DP antennas. In this proposed technique, called DP-SSM, transmitted symbols are not selected from a single constellation as in conventional DP-SM. Instead, they are chosen from multiple constellations, each marked by the signature of a selected transmit antenna and its polarization. With the addition of the polarization dimension, the signature constellations, which are also derived through successive geometric interpolations starting from the primary constellation, expand into the third dimension. We outline a general methodology for constructing these three-dimensional (3D) signature constellations and provide theoretical upper bounds for the ABEP. Both simulation and analytical results demonstrate that DP-SSM achieves significant performance improvements compared to conventional and precoded DP-SM, even under severe correlation scenarios and in the presence of a LOS channel component. Compared to equivalent UP systems, DP-SSM offers comparable or enhanced performance while utilizing only

half the number of antennas and thus requiring half the space. This strategy does not alter the average power budget and avoids the optimization complexity associated with precoding, being implementable with only a slight increase in transmitter memory.

4.2. Signature Constellations and System Model

4.2.1. DP-SM with Signature Constellations

We examine an $N_t \times N_r$ dimensional DP-SSM system utilizing dual-polarized antennas, where the number of transmit antennas N_t is an integer power of 2, hence $N_t = 2^n$. The modulated symbols are derived from an M -ary constellation, thus each symbol carries $m = \log_2 M$ bits. Each group of incoming $n + m + 1$ bits is divided into n bits to select the active antenna, m bits to choose one of the M possible complex constellation points defined by the chosen signature constellation, and 1 bit to select the polarization dimension.

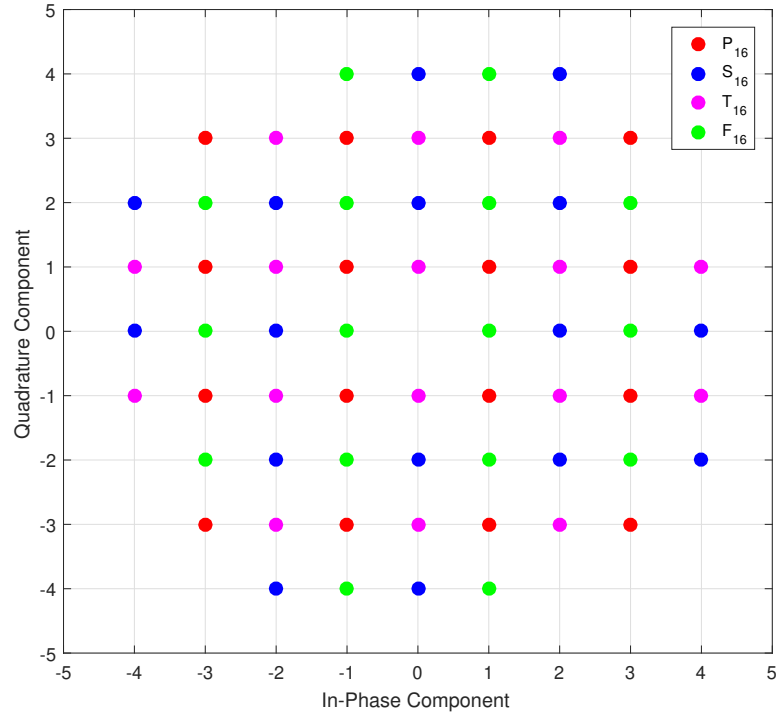


Figure 4.1. The signature constellation P_M , S_M , T_M and F_M used for $N_t = 2$ and $M = 16$, 2-Dimensional representation.

The signature constellation alphabet is denoted by $\mathcal{X}^{(u)} = \{X_1^{(u)}, \dots, X_k^{(u)}, \dots, X_M^{(u)}\}$ for $u = 1, 2, \dots, 2N_t$. Unlike traditional SM, where each active antenna polarization dimension transmits a symbol from a single constellation, in SSM, symbols are chosen from an antenna-dependent constellation at each time instant, as described in [12].

Assuming M -ary QAM constellations, the corresponding signature constellations suitable for DP-SSM are constructed as follows. The signature constellation for the vertical polarization of the first antenna corresponds to the original M -QAM signal constellation, denoted by P_M . The vertical polarization signature constellation for the second antenna, S_M is derived using first-order geometric interpolation within the primary constellation plane. For the horizontal polarization of both the first and second antennas, the third and fourth constellations, T_M and F_M , are generated via second-step geometric interpolation in the primary constellation plane.

Example signature constellations for 16-QAM are illustrated in Figure 4.1. With dual-polarized antennas, constellations used by orthogonal polarizations are positioned on orthogonal planes. In our system, the constellations used by the p polarization (vertical) of the first and second antennas are located in the X-Y plane, while the constellations for the p' polarization (horizontal) of both antennas are situated in the X-Z plane, as shown in Figure 4.2. The geometric interpolation technique employed to derive the secondary, third, and fourth constellations is optimal in maximizing the minimum Euclidean distance between constellations. Specifically, if δ_0 represents the minimum distance of the primary signal constellation, then:

- All signature constellations have the same minimum distance δ_0 as the primary constellation.
- The minimum distances between P_M and S_M and between T_M and F_M are both $\delta_0/\sqrt{2}$.
- The minimum distance between (P_M) and (T_M) , (P_M) and (F_M) and (S_M) and (F_M) in the 3D plane are also $\delta_0\sqrt{2}$.

This approach can easily be extended to higher numbers of antennas by using the same geometric interpretations and to higher-order of modulations such as 64-QAM.

It guarantees that the constellation symbols transmitted from different antennas or polarization dimensions are never the same, but in fact they are separated from each other by a minimum distance. This fact eliminates one of the fundamental error events of conventional SM and thus serves as a factor to improve the error performance of SM against the channel and antenna correlations. In fact, given an intra-constellation minimum distance of δ_0 , the minimum inter-constellation minimum distance for DP-SSM is always $\delta_0/\sqrt{N_t/2}$ whereas for conventional SM this distance is 0 when $N_t > 1$. On the other hand, in the constellation rotation-based precoding approach when maximizing the minimum inter-constellation Euclidean distance rule, it can be shown that the minimum inter-constellation distance is $\delta_0\sqrt{2}\sin(\frac{\pi}{4N_t})$ which is a smaller quantity than the one for conventional SM and SSM.

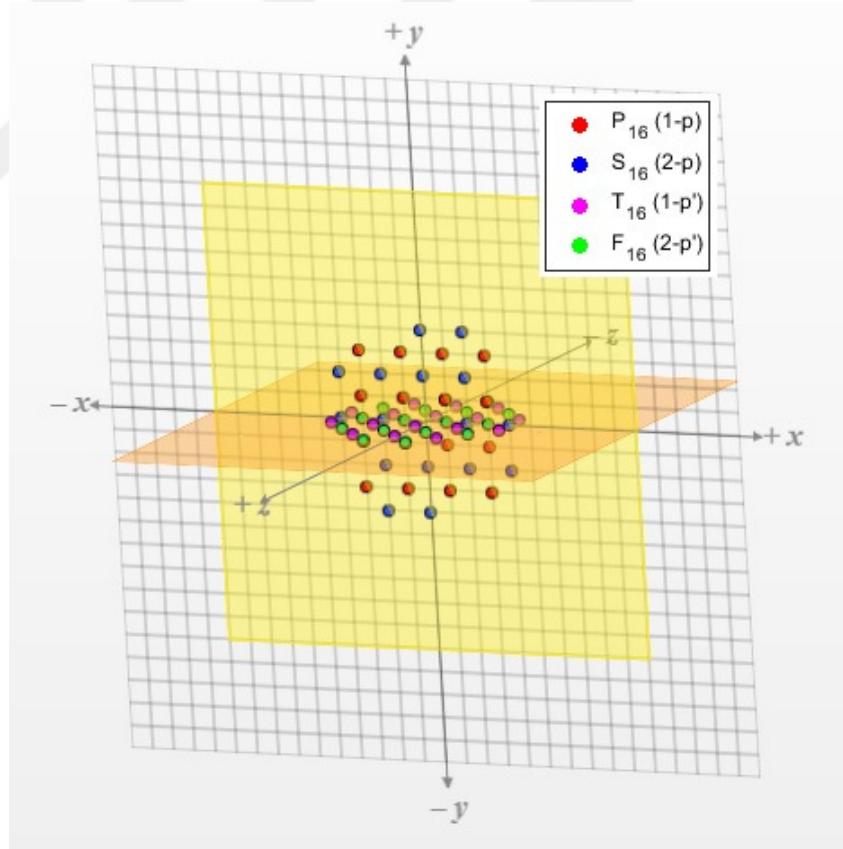


Figure 4.2. The signature constellation P_M , S_M , T_M and F_M used for $N_t = 2$ and $M = 16$, 3-Dimensional representation.

4.2.2. Signal and Channel Model

The modulated signal is transmitted over a channel and the corresponding received signal can be written as

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \boldsymbol{\nu}, \quad (4.1)$$

where $\mathbf{H}$ represents the $2N_r \times 2N_t$ channel matrix, $\mathbf{x}$ is $2N_t \times 1$ transmitted signal vector, $\mathbf{y}$ and $\boldsymbol{\nu}$ are $2N_r \times 1$ received signal and noise vectors respectively. The elements of $\boldsymbol{\nu}$ are independent and identically distributed (i.i.d.) complex Gaussian variables with zero-mean and variance N_0 , i.e., $\nu_k \sim \mathcal{CN}(0, N_0)$ for $k = 1, \dots, 2N_r$. Notice that there is only one active transmit antenna and polarization state at each signalling instant, which means that $\mathbf{x}$ has only one non-zero element. If the u -th antenna is active and the v -th symbol in $X^{(u)}$ is transmitted, the equation in (4.1) can be rewritten as

$$\mathbf{y} = \mathbf{h}_u \mathbf{X}_v^{(u)} + \boldsymbol{\nu}, \quad (4.2)$$

where $\mathbf{h}_u$ represents the u -th column of $\mathbf{H}$ with u corresponding to a unique active transmit antenna and polarization direction pair (u', j) for $u' = 1, 2, \dots, N_t$ and $j = p, p'$. In ideal antenna cases, with a particular polarization, the received cross-polar signal amount has to be zero. However in real-world systems it is not the case and some part of the received signal comes from the polarization which is orthogonal to the desired polarization. This causes power leakages from one polarization direction to its orthogonal counterpart in the fixed and variable components of the channel. These leakage amounts are defined by μ_f and μ_v for fixed and variable components respectively, also the co-polar terms can be defined by assuming symmetry between different polarizations by $1 - \mu_f$ and $1 - \mu_v$ for fixed and random components respectively. The channel can de-polarized the transmitted signal, when $\mu_f = \mu_v = 0$ but for cases other than this condition, cross-polar and co-polar signals are going to be present. In this case the ratio of co-polar term to cross-polar term is characterized by XPD (cross-polar discrimination). The XPD factors for fixed and variable components for channel are

$$\chi_f = \frac{1 - \mu_f}{\mu_f} \quad \text{and} \quad \chi_v = \frac{1 - \mu_v}{\mu_v}. \quad (4.3)$$

Larger XPD values, exhibit small cross-polar value which is the result of higher capability of the channel to de-polarize the orthogonal polarization directions. In addition

smaller XPD represents more considerable cross-polar interference (XPI). The amounts of power leakages from one polarization to the other can be collected in leakage matrices for fixed and variable components of the channel as $\mathbf{\Gamma}_f$ and $\mathbf{\Gamma}_v$ respectively. Furthermore, a second impact of the dual-polarized antennas is polarization correlation (XPC) between orthogonal polarization directions which comes when the scattering of the channel is poor de-polarization of signal will become a problem. This causes polarization correlation coefficients at transmitter and receiver. These coefficients are assumed to be the same for all dual-polarized antennas. $\mathbf{\Pi}_t$ and $\mathbf{\Pi}_r$ are the polarization correlation matrices. We assume a slow fading channel which is the summation of an average and a variable component. The channel matrix $\mathbf{H}$ is described as

$$\mathbf{H} = \sqrt{\frac{K}{K+1}} \bar{\mathbf{H}} + \sqrt{\frac{1}{K+1}} \tilde{\mathbf{H}}, \quad (4.4)$$

where $\bar{\mathbf{H}}$ and $\tilde{\mathbf{H}}$ are the fixed and variable components, respectively. The fixed part $\bar{\mathbf{H}}$ is such that $[\bar{\mathbf{H}}]_{p,q} = \bar{h}_{p,q}$, for $p = 1, \dots, 2N_r$ and $q = 1, \dots, 2N_t$. The variable component of the channel matrix, $\tilde{\mathbf{H}}$, consists of possibly correlated complex coefficients which belong to Gaussian random distribution. The square-root terms are the normalization weights with K being the Rician factor. The correlation between the entries of $\mathbf{H}$ using dual polarized antennas is due to the both polarization effects within each antenna polarization direction and also insufficient spacing between antenna pairs. These effects can be combined in matrices at the transmitter and receiver sides as $\mathbf{\Phi}_t$ and $\mathbf{\Phi}_r$ which are defined as the Kronecker product of the spatial and polarization correlation effects as $\mathbf{\Phi}_t = \mathbf{\Sigma}_t \otimes \mathbf{\Pi}_t$ and $\mathbf{\Phi}_r = \mathbf{\Sigma}_r \otimes \mathbf{\Pi}_r$. The transmit and receive correlation effects due to insufficient space allocation can be formed using the exponential model in [95], as $\mathbf{\Sigma}_t$ and $\mathbf{\Sigma}_r$, respectively, where $[\mathbf{\Sigma}_t]_{i,j} = \alpha_t^{|i-j|}$ for $i, j = 1, 2, \dots, N_t$ and $[\mathbf{\Sigma}_r]_{i,j} = \alpha_r^{|i-j|}$ for $i, j = 1, 2, \dots, N_r$, with transmit and receive correlation coefficients $|\alpha_t|, |\alpha_r| < 1$. $\mathbf{\Pi}_t$ and $\mathbf{\Pi}_r$ are polarization correlation effects expressed in block form at the transmitter and receiver as given in [92], [96], [97].

Accordingly, the general channel model can be rewritten as

$$\mathbf{H} = \sqrt{\frac{K}{1+K}} (\mathbf{1}_{N_r \times N_t} \otimes \mathbf{\Gamma}_f) \odot \bar{\bar{\mathbf{H}}} + \sqrt{\frac{1}{1+K}} (\mathbf{1}_{N_r \times N_t} \otimes \mathbf{\Gamma}_v) \odot (\mathbf{\Phi}_r^{\frac{1}{2}} \mathbf{H}_w \mathbf{\Phi}_t^{\frac{1}{2}}), \quad (4.5)$$

where the elements of $\bar{\bar{\mathbf{H}}}$ are all one and the elements of $\mathbf{H}_w$ are independent and

identically distributed circularly symmetric zero-mean complex Gaussian variables with unit variance. $\mathbf{1}_{N_r \times N_t}$ is an all one $N_r \times N_t$ matrix. $\mathbf{\Gamma}_v$ and $\mathbf{\Gamma}_f$ are defined as matrices corresponding to the polarization correlations for the variable and the fixed parts of the channel which are described in [17] in detail.

4.3. Average Bit Error Analysis of DP-SSM

Given the received signal model in (4.2) and the channel model in (4.4), the receiver employs maximum likelihood (ML) detection to estimate the active antenna, its polarization direction and the transmitted symbol. Based on the signal model in (4.2), the ML optimum antenna and symbol index pair $(u, v)_{ML}$ is

$$(u, v)_{ML} = \arg \max_{u, v} f_{\mathbf{Y}}(\mathbf{y} | \mathbf{x}, \mathbf{H}) = \arg \min_{u, v} D(\mathbf{y} | \mathbf{h}_u, X_v^{(u)}), \quad (4.6)$$

where $D(\mathbf{y} | \mathbf{h}_u, X_v^{(u)}) = \|\mathbf{y} - \mathbf{h}_u X_v^{(u)}\|^2$ is the distance metric. The ABEP with optimum ML detection can be computed by using the well-known upper bound technique in [86],

$$\bar{P}_b \leq \frac{1}{2N_t M} \sum_{u=1}^{2N_t} \sum_{\hat{u}=1}^{2N_t} \sum_{v=1}^M \sum_{\hat{v}=1}^M \frac{N(u, \hat{u}, v, \hat{v})}{\log_2(2N_t M)} \bar{P}_s(u, \hat{u}, v, \hat{v}). \quad (4.7)$$

The number of bits in error between respective channel and symbol index pairs is represented by $N(u, \hat{u}, v, \hat{v})$ and $\bar{P}_s(u, \hat{u}, v, \hat{v})$ is the average pairwise symbol error probability (APEP). The $\log_2(2N_t M)$ term denotes the total number of antenna and symbol bits. $\bar{P}_s(u, \hat{u}, v, \hat{v})$ can be computed using the approaches in [11] and [17] as

$$\bar{P}_s(u, \hat{u}, v, \hat{v}) = \mathbb{E} \left[Q \left(\sqrt{\frac{\|\mathbf{z}\|^2}{2}} \right) \right], \quad (4.8)$$

with $\mathbb{E}[\cdot]$ corresponding to the expectation operation and $\mathbf{z}$ is a $2N_r \times 1$ vector as

$$\mathbf{z} = \frac{1}{\sqrt{N_0}} (\mathbf{h}_u X_v^{(u)} - \mathbf{h}_{\hat{u}} X_{\hat{v}}^{(\hat{u})}). \quad (4.9)$$

The mean vector of $\mathbf{z}$ is computed as

$$\mathbf{m}_z = \sqrt{\frac{K}{N_0(K+1)}}(\bar{\mathbf{h}}_u X_v^{(u)} - \bar{\mathbf{h}}_{\hat{u}} X_{\hat{v}}^{(\hat{u})}), \quad (4.10)$$

and the the covariance matrix of $\mathbf{z}$ vector is calculated as

$$\mathbf{\Lambda}_z = \frac{\phi_{u,u}^t |X_v^{(u)}|^2 + \phi_{\hat{u},\hat{u}}^t |X_{\hat{v}}^{(\hat{u})}|^2 - 2\text{Re}\{\phi_{u,\hat{u}}^t X_v^{(u)} (X_{\hat{v}}^{(\hat{u})})^*\}}{N_0(K+1)} \mathbf{\Phi}_r, \quad (4.11)$$

where $\phi_{i,j}^t$ is the (i,j) -th entry of the transmit correlation matrix $\mathbf{\Phi}_t$. By using the moment generating function we can find an upper bound for APEP as

$$\bar{P}_s(u, \hat{u}, v, \hat{v}) = \frac{1}{\pi} \int_0^{\frac{\pi}{2}} \frac{\exp(-\mathbf{m}_z^\dagger [\mathbf{\Lambda}_z + 4 \sin^2 \theta \mathbf{I}]^{-1} \mathbf{m}_z)}{|\frac{\mathbf{\Lambda}_z}{4 \sin^2 \theta} + \mathbf{I}|} d\theta. \quad (4.12)$$

Once the APEP between all channel and symbol pairs is computed, it can be used to obtain the ABEP upper bound in Equation (4.7) in both correlated Rayleigh and Rician fading scenarios.

4.4. Simulation Results

This section presents simulation results that illustrate the close alignment between theoretical ABEP upper bounds and simulated performance, as well as the performance improvements offered by the proposed technique over benchmark methods. The simulations are conducted for a 2×2 DP system utilizing 16-point and 64-point QAM constellations, which correspond to spectral efficiencies of 6 and 8 bits per channel use (bpcu), respectively. Performance comparisons include conventional and precoded DP-SM systems, as well as conventional 2×2 and 4×4 uni-polarized SM systems with equivalent spectral efficiencies.

The focus in this work is on highly correlated transmit antenna systems, where the proposed architecture's improvements are more pronounced. Consequently, all simulations assume a correlation coefficient of $\alpha_t = 0.9$, between adjacent transmit antennas, with $\mathbf{\Sigma}_r = \mathbf{I}$, indicating no correlation at the receiver side to isolate the impact of transmit correlations. The fixed part of the channel matrix $\bar{\mathbf{H}}$ is set as an all-one matrix for simplicity, and the Rician factor K is chosen as $K = 3$. DP antennas

are assumed to have identical polarization characteristics, with XPD factors χ_f and χ_v respectively, set to $\chi_f = 15$ dB and $\chi_v = 5$ dB, representing the power ratios of the co-polar and cross-polar terms in both the fixed and variable parts. Figures show that theoretical upper bounds closely match simulated ABER performances, especially at moderate to high SNR values across all channel scenarios.

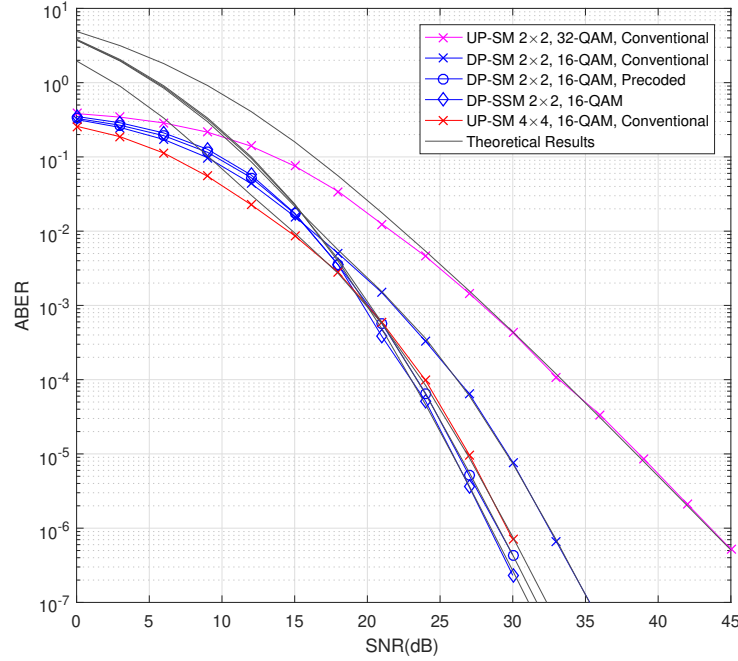


Figure 4.3. Performance of DP-SSM with spectral efficiency of $R = 6$ b/s/Hz over correlated Rayleigh channel.

Figures 4.3 and 4.4 compare ABER for systems operating at 6 bpcu spectral efficiency. Figure 4.3 demonstrates that DP-SSM outperforms conventional DP-SM and precoded DP-SM by 4 dB and approximately 0.5 dB, respectively, at the 10^{-5} ABER level. Notably, while precoded DP-SM requires complex precoder design, DP-SSM achieves superior performance without such optimization. Additionally, the 2×2 DP-SSM system surpasses the conventional 4×4 UP-SM system by 1.5 dB, despite using only half the spatial resources. For correlated Rician fading channels, the proposed approach yields even larger gains, as illustrated in Figure 4.4. Performance degradation in SM, due to spatial correlations and the LOS component, is mitigated by precoding and signature constellations. The rotation-optimized precoding approach enhances performance by approximately 6.5 dB at the 10^{-5} ABER level compared to conventional

DP-SM. DP-SSM provides an additional 4 dB gain over precoded DP-SM, without the need for transmitter optimization. Furthermore, DP-SSM outperforms the equivalent 4×4 UP-SM system by 8 dB at 10^{-5} ABER while it occupies half as much space as its uni-polarized counterpart.

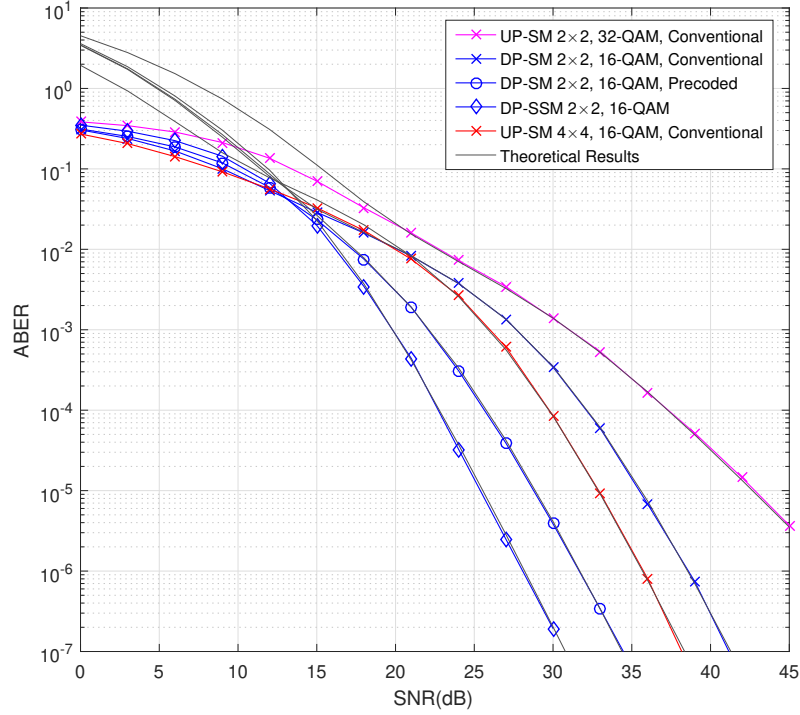


Figure 4.4. Performance of DP-SSM with DP-SM, UP-SM and rotation optimized precoded DP-SM with spectral efficiency of $R = 6$ b/s/Hz over correlated Rician channel with $\alpha_t = 0.9$ and $K = 3$.

In Figure 4.5 and 4.6, the ABER performance comparisons obtained for 8 bpcu spectral efficiency are presented. The 2×2 DP-SM system employs 64-QAM for which the precoding approach does not provide any additional performance improvements and thus the precoded performance curves are not shown for this system. For correlated Rayleigh channel fading, as shown in Figure 4.5, DP-SSM system provides a performance improvements of 3 dB and approximately 1 dB at 10^{-6} ABER level compared to the conventional DP-SM system and the conventional 4×4 UP-SM system, respectively. In addition, in the presence of LOS component, as shown in Figure 4.6, 2×2 DP-SSM system improves the performance of conventional DP-SM system by 9 dB at 10^{-6} ABER level and outperforms the 4×4 UP-SM system by 6 dB. In other

words, when higher-order constellations are employed, using 2×2 DP-SSM system is better than using 4×4 UP-SM system.

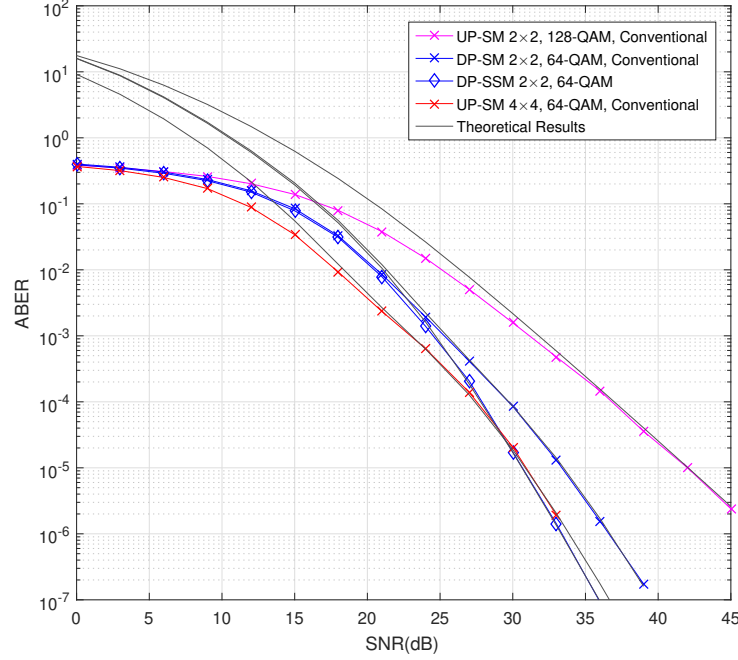


Figure 4.5. Performance of DP-SSM with $R = 8$ b/s/Hz over correlated Rayleigh channel.

Further results for higher-order modulations and larger numbers of dual-polarized antennas show similar trends, though not presented here due to space constraints. These findings confirm that extending SSM to include polarization forms a robust SM-MIMO architecture resilient to antenna and channel correlations. Even higher-order systems with more transmit antennas or larger constellation sizes would yield comparable gains, as demonstrated with 16-QAM and 64-QAM, highlighting that signature spatial modulation system gains are independent of the QAM constellation size.

4.5. Conclusion

In this work, the advantages of SSM and dual-polarized transmission are combined so as to obtain a robust SM-MIMO transceiver architecture against the effects of transmit antenna and channel correlations. Both theoretical analysis and extensive simulations show that the proposed technique has the potential to offer significant

performance gains in comparison to conventional DP-SM and the equivalent UP-SM systems. Also for DP-SM with high spectral efficiency using modulations such as 16-QAM or 64-QAM which are used commonly in the state-of-the-art approaches, our proposed system provides significant performance improvements, especially on Rician channels, with respect to the previously introduced techniques involving precoded DP-SM, without involving any complexity and transmitter optimization. The observations from the results reveal the fact that DP-SSM represents a robust DP-SM MIMO technique and introduces a reliable transmission at high spectral efficiency applications over correlated Rayleigh and Rician fading channels.

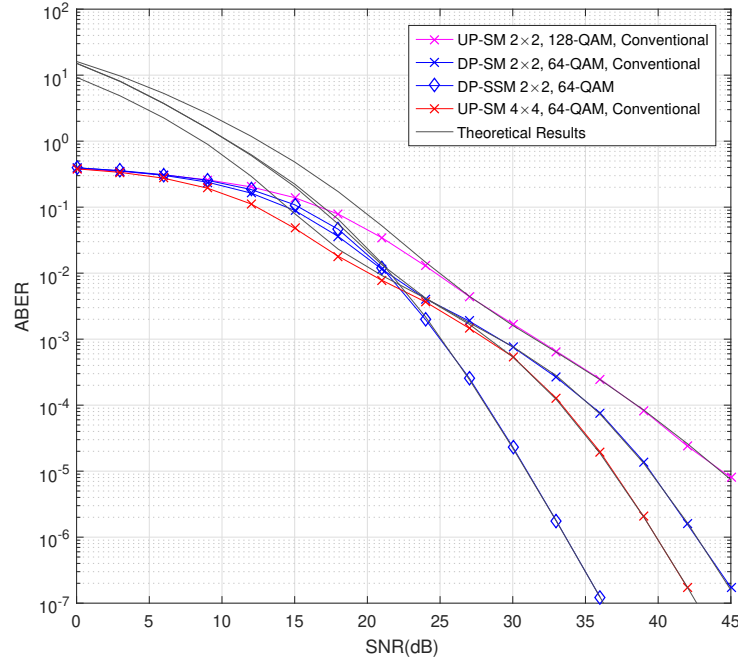


Figure 4.6. Performance of DP-SSM with DP-SM and UP-SM with $R = 8$ b/s/Hz over correlated Rician channel with $\alpha_t = 0.9$ and $K = 3$.

5. PRECODED MULTI-STREAM SPATIAL MODULATION

5.1. Introduction

SM, initially developed for MIMO systems, where the number of radio frequency (RF) chains is fewer than the number of antennas, which reduces the number of transmit power amplifiers needed [27], [28]. SM is especially appropriate for cellular system uplinks due to constraints on cost and power consumption, which often prevent having as many RF chains as antennas required by downlink performance needs. The number of active antennas at any moment is limited by the available RF chains, constraining the uplink data streams. Yet, by not activating all antennas simultaneously, additional data can be conveyed via the indices of the chosen antennas for transmission. When only one antenna is active, this is referred to as single-stream SM, and when more than one is active, it is called multi-stream SM (MSM) [6]. A significant challenge in SM arises when the channel correlations are high. The receiver not only needs to decode the transmitted symbols but also identify which antennas were active. High correlation among channels complicates identifying the active antennas, adversely affecting the BER performance of SM. To address this, extensive research has focused on overcoming issues of antenna and channel correlation in SM, resulting in various techniques, particularly for single-stream SM [11], [12] and [29–34]. The most effective of these is linear precoding at the transmitter, which enhances signal distinctiveness at the receiver, even with nearly identical channel responses. Linear precoding includes traditional rotation-based methods [11], [29–33] and more recently developed signature constellation-based precoding [12], [34]. While rotation-based precoding fits lower-level constellations like BPSK and QPSK, its effectiveness decreases with higher constellations such as 16-QAM and 64-QAM, where signature constellation-based precoding proves to be more efficient.

5.2. System Model

5.2.1. Multi-Stream SM

SM transmits data by utilizing the indices of the active antennas along with the symbols transmitted from these antennas. Let's consider a system with N_t transmit antennas, where only N_a of them are active at any given time. The active antennas transmit data symbols from an M-QAM signal constellation, where each symbol encodes $m = \log_2(M)$ bits. The active antennas transmit $N_a m$ bits, and the selection of active antennas adds an additional $\lfloor \log_2 C_{N_t}^{N_a} \rfloor$ bits, where C_y^x represents the number of combinations for selecting x out of y elements, and $\lfloor x \rfloor$ denotes the integer part of x . Therefore, the total number of information bits per channel use

$$\mathbf{n} = N_a m + \lfloor \log_2 C_{N_t}^{N_a} \rfloor. \quad (5.1)$$

Considering an MSM system with N_r receive antennas, the input – output relationship of the channel can be written as

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \nu, \quad (5.2)$$

where $\mathbf{H}$ represents the $N_r \times N_t$ -dimensional channel matrix, $\mathbf{x}$ denotes the N_t -dimensional transmitted signal vector, $\mathbf{y}$ is the N_r -dimensional received signal vector, and ν is the N_r -dimensional AWGN vector. The N_t -dimensional transmitted signal vector $\mathbf{x}$ has only N_a non-zero components, all other components being 0. At the receiver, it must estimate the transmitted symbols and determine which antennas transmitted these symbols. Using the notation from (5.2), the receiver must identify the positions of the non-zero components of $\mathbf{x}$ and the values of these components. In MSM, determining the positions of the non-zero components is challenging when the channels are highly correlated. Precoding introduces differences between the signals transmitted by different antennas, thereby compensating for the loss of independence between the channel responses.

5.2.2. Antenna and Channel Correlations

We employ a slow fading general channel model characterized by the sum of a fixed (LOS) component and a random component. When the LOS component is zero, this model effectively represents Rayleigh fading channels, commonly used for large-cell cellular systems. Conversely, with a non-zero LOS component, it models Rician fading, which is suitable for small-cell radio channels. According to this model, the $N_r \times N_t$ channel matrix is expressed

$$\mathbf{H} = \sqrt{\frac{K}{K+1}} \bar{\mathbf{H}} + \sqrt{\frac{1}{K+1}} \tilde{\mathbf{H}}, \quad (5.3)$$

where $\bar{\mathbf{H}}$ and $\tilde{\mathbf{H}}$ represent the fixed and random components, respectively. The parameter K in this expression is the Rician factor, where the quantities inside the square roots indicate the energy distribution between these two components. When $K = 0$, the model simplifies to Rayleigh fading. Typical K values for Rician fading range from 3 to 5. The random component of the channel matrix consists of complex Gaussian variables. Assuming that the real and imaginary parts have identical auto-correlation and are uncorrelated with each other, we use the well-known Kronecker model to express $\tilde{\mathbf{H}}$ as

$$\tilde{\mathbf{H}} = \mathbf{\Sigma}_r^{\frac{1}{2}} \check{\mathbf{H}} \mathbf{\Sigma}_t^{\frac{1}{2}T}, \quad (5.4)$$

where $\mathbf{\Sigma}_r$ and $\mathbf{\Sigma}_t$ are real-valued and Hermitian symmetric receive and transmit correlation matrices, respectively, and where $\check{\mathbf{H}}$ is the independent Rayleigh fading channel matrix whose elements are i.i.d. complex Gaussian random variables of zero mean and unity variance. The superscript T in (5.4) denotes matrix transpose. Combining (5.3) and (5.4), we obtain

$$\mathbf{H} = \sqrt{\frac{K}{K+1}} \bar{\mathbf{H}} + \sqrt{\frac{1}{K+1}} \mathbf{\Sigma}_r^{\frac{1}{2}} \check{\mathbf{H}} \mathbf{\Sigma}_t^{\frac{1}{2}T}. \quad (5.5)$$

By appropriately selecting the $\mathbf{\Sigma}_r$ and $\mathbf{\Sigma}_t$ matrices and the K parameter in (5.5), this model can be used to represent uncorrelated or correlated Rayleigh and Rician fading channels. In the following section, we address precoder design to compensate for transmit antenna correlations characterized by $\mathbf{\Sigma}_t$. Note that precoding does not

help against receive antenna correlations represented by Σ_r . As such, precoding is best suited for the uplink, because the small size of user devices leads to transmit antenna correlations, while additional degrees of freedom exist to reduce the receive antenna correlation at the base station.

5.3. Precoder Design

The most widely recognized precoding technique for mitigating transmit antenna and channel correlations involves using phase-rotated constellations. In this approach, the signal constellation is rotated differently at each transmit antenna, and the receiver must determine both the transmitted constellation point and the rotation applied. While effective for SM with low-level constellations such as BPSK and QPSK and a small number of transmit antennas, this method encounters significant limitations with high-level constellations or a large number of antennas.

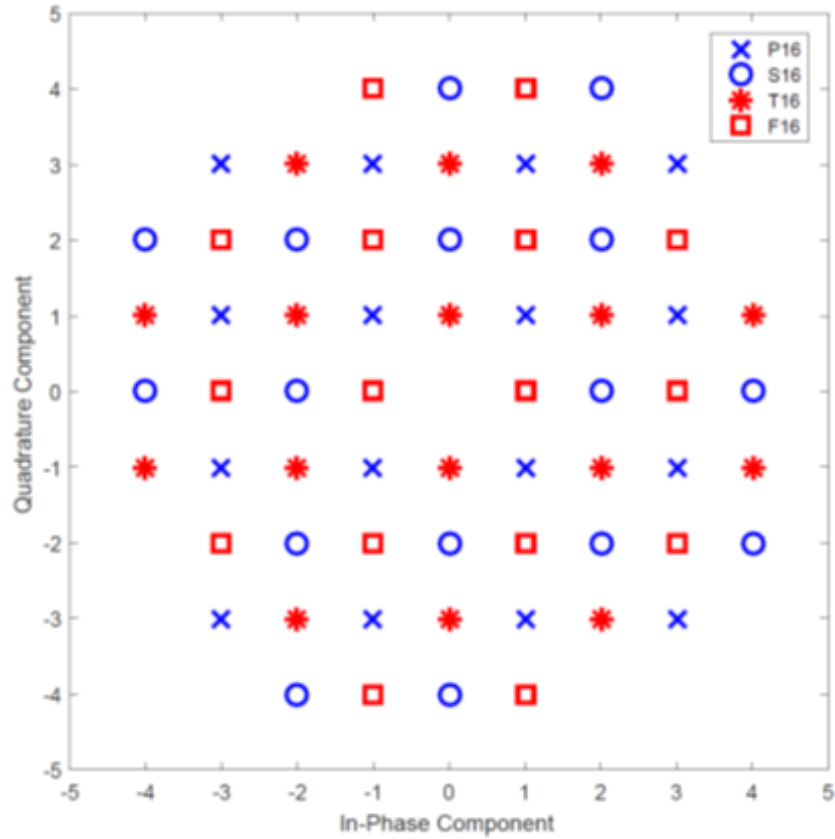


Figure 5.1. The 16-point signature constellations used for SM with $N_t = 4$.

For higher-order QAM constellations, a new precoding technique was recently introduced, utilizing different types of signature constellations. This method, originally developed for designing enhanced spatial modulation (ESM) [13], [14], provides an alternative to phase rotation. Figure 5.1 illustrates the signature constellations used for $N_t = 4$ and $M = 16$. In this figure, the blue crosses denote the P_{16} constellation, which is used as the signature for the first antenna. The blue circles represent the S_{16} constellation for the second antenna, the red stars indicate the T_{16} constellation for the third antenna, and the red squares mark the F_{16} constellation for the fourth antenna. The P_{16} constellation corresponds to the standard 16-QAM signal constellation. The S_{16} constellation is derived by performing a simple geometric interpolation between the points of P_{16} and removing the point at the origin. Additional points are then added at the constellation boundary using the corresponding infinite lattice, selecting the signal points with minimum energy. The T_{16} and F_{16} constellations are determined through a second step of geometric interpolation. Importantly, all these constellations maintain the same minimum Euclidean distance. This can be expressed as

$$d_{\min}(P_M) = d_{\min}(S_M) = d_{\min}(T_M) = d_{\min}(F_M) = \delta_0 = 2. \quad (5.6)$$

The inter-constellation distance properties are as follows

$$d_{\min}(P_M, S_M) = d_{\min}(T_M, F_M) = \delta_0/\sqrt{2} = \sqrt{2}, \quad (5.7)$$

and

$$d_{\min}(P_M, T_M) = d_{\min}(P_M, F_M) = d_{\min}(S_M, T_M) = d_{\min}(S_M, F_M) = \delta_0/2. \quad (5.8)$$

In single-stream SM, where $N_a = 1$, this set of constellations ensures a sufficient minimum inter-constellation distance to accurately estimate all information bits, even under conditions of full transmitter correlation. Unlike phase rotation-based precoding, this approach's minimum distance is independent of the constellation size $\delta_0 = 2$ representing the minimum distance of the primary constellation, the minimum distance in the signal space is $\delta_0/2 = 1$ for all M . Studies in [12] and [34] have shown that for single-stream SM, this technique offers a significant improvement over conventional rotation-based precoding.

Next, we consider the design of precoders for MSM with $N_t = 4$ and $N_a = 2$. To do this, we recall that the signal space in conventional MSM can be represented as

$$\chi = \{(P_M, P_M, 0, 0), (0, 0, P_M, P_M), (P_M, 0, 0, P_M), (0, P_M, P_M, 0)\}. \quad (5.9)$$

In this signal space, the first term indicates the scenario where the first and second antennas are active, while the second term represents the scenario where the third and fourth antennas are active, and so forth. In each scenario, the symbols transmitted by the two active antennas are selected from the same constellation P_M . MSM with precoding, based on constellation rotations, can then be described as follows

$$\begin{aligned} \chi' = \{ & (P_M, e^{j\theta_1} P_M, 0, 0), (0, 0, e^{j\theta_2} P_M, e^{j\theta_3} P_M), \\ & (P_M, 0, 0, e^{j\theta_3} P_M), (0, e^{j\theta_1} P_M, e^{j\theta_2} P_M, 0)\}, \end{aligned} \quad (5.10)$$

where the phases $\theta_i, i = 1, 2, 3$ are optimized to maximize minimum Euclidian distance in the signal space. Note that the signal constellation is used without any phase rotation at the first antenna. Proceeding further, direct extension to MSM of the precoding technique based on signature constellations that is used in [34], [12] for single-stream SM would lead to the following signal space

$$\begin{aligned} \chi_1 = \{ & (P_M, S_M, 0, 0), (0, 0, T_M, F_M), \\ & (P_M, 0, 0, F_M), (0, S_M, T_M, 0)\}. \end{aligned} \quad (5.11)$$

To illustrate the limitation of this approach, assume we are dealing with full correlation, meaning all channel responses are identical. In this case, the noiseless signal received by an antenna, up to a complex coefficient, will be derived from $P_M + S_M$, or $T_M + F_M$, or $P_M + F_M$, or $S_M + T_M$, depending on which pair of antennas is active. A simple inspection using Figure 5.1 reveals that these constellations overlap, making it impossible to uniquely determine the transmitting antenna pair and the corresponding information bits. For example, consider the complex signal point $1 + j$ on Figure 5.1 which belongs to constellation P_M and the point 2 from constellation S_M . Their vector sum is $3 + j$, which is also in constellation P_M . Similarly, the point j which belongs to constellation T_M and the signal point 3 which belongs to constellation F_M . The sum of these two signal points too is $3 + j$ belonging to constellation P_M . Therefore, upon receiving the noiseless signal point $3 + j$, we cannot determine whether the active

antennas pair is Antenna 1 and Antenna 2 or Antenna 3 and Antenna 4. This example demonstrates that the precoding technique described by (5.11) is inadequate for MSM. To address this ambiguity, we first consider a precoder design based on the following signal space

$$\chi_2 = \{(P_M, P_M, 0, 0), (0, 0, S_M, F_M), (F_M, 0, 0, T_M), (0, S_M, T_M, 0)\}. \quad (5.12)$$

In the case of full correlation, the received noiseless signal takes its values from $P_M + P_M$ when the activated antennas are the first and the second, from $S_M + F_M$ when the activated antennas are the third and the fourth, from $T_M + F_M$ when the activated antennas are the first and the fourth, and from $S_M + T_M$ when the activated antennas are the second and the third. A simple inspection of Figure 5.1 reveals that with this precoding the 4 constellations of the received noiseless signal in the full correlation case are completely disjoint, meaning that the information bits assigned to antenna selection can be uniquely determined. However, the information bits associated to the transmitted symbols cannot be uniquely determined. To view this, consider the case where the active antennas are the second and the third, antenna 2 is transmitting a symbol $x \in S_M$, antenna 3 is transmitting a symbol $y \in T_M$, and the received and properly scaled noiseless signal is $r = x + y \in S_M + T_M$. Consider now another case, where antenna 2 is transmitting a symbol $x' = x + \delta_0$ and antenna 3 is transmitting a symbol $y' = y - \delta_0$. The received noiseless signal in this case is $r' = x' + y' = x + y = r$, showing that the transmitted symbols cannot be uniquely determined. To solve this issue, we go one step further and introduce a phase rotation between the constellations used by the two activated antennas. The signal space becomes as follows

$$\chi_3 = \{(P_M, e^{j\theta} P_M, 0, 0), (0, 0, S_M, e^{j\theta} F_M), (e^{j\theta} F_M, 0, 0, T_M), (0, S_M, e^{j\theta} T_M, 0)\}. \quad (5.13)$$

It can be easily verified that this signal space does not contain any ambiguity in the case of full correlation, and therefore it can be expected to provide a significant improvement compared to MSM on highly correlated channels. The phase parameter θ must be optimized to maximize the minimum distance in the signal space and get the best performance. Another solution to resolve the ambiguity issue is to use an amplitude

mismatch between the constellations used by the two active antennas instead of phase rotation. The corresponding signal space can be written as

$$\begin{aligned} \chi_4 = \{ & (aP_M, bP_M, 0, 0), (0, 0, aS_M, bF_M), \\ & (bF_M, 0, 0, aT_M), (0, aS_M, bT_M, 0) \}, \end{aligned} \quad (5.14)$$

with $a^2 + b^2 = 2$ in order to keep the transmitted signal energy constant. As the phase parameter in the previous technique, the scalars a and b in this technique must be optimized to maximize minimum distance. Proceeding further, we can jointly use the phase rotation and amplitude scaling and write the signal space as

$$\begin{aligned} \chi_5 = \{ & (aP_M, be^{j\theta}P_M, 0, 0), (0, 0, aS_M, be^{j\theta}F_M), \\ & (be^{j\theta}F_M, 0, 0, aT_M), (0, aS_M, be^{j\theta}T_M, 0) \}. \end{aligned} \quad (5.15)$$

The signal spaces in Equations (5.13) – (5.15) represent 3 different variants of the proposed precoding scheme.

5.4. Simulation Results

Using upper bounding techniques as in [33], the ABER performance of the proposed technique was evaluated and compared to that of the original MSM without precoding as well as to the performance of rotation based precoding. The performance evaluation was carried out using $N_t = 4$, $N_a = 2$, and 16-point constellations as described in the previous section, and 4 receive antennas $N_r = 4$. The transmit and receive correlation matrices were based on the exponential Kroenecker model [95], where the entries are of the form $[\Sigma_{(p,q)}] = \alpha^{|p-q|}$, α being a fixed correlation coefficient between adjacent antennas. Regarding the channel, we used both Rayleigh fading and Rician fading with $K = 3$. It was assumed that the receive antennas have zero correlation, i.e., the receive correlation parameter was $\alpha_r = 0$, because as indicated previously, precoding does not help against receive antenna correlation. For the transmit antenna correlation, we used $\alpha_t = 0.99$ and $\alpha_t = 0.9$.

The ABER performance results are depicted in Figures. 5.2 – 5.5 for the following MSM schemes,

- Original MSM without precoding as represented by Equation (5.9).
- MSM with rotation-based precoding represented by Equation (5.10).
- Signature constellation based MSM, denoted SSM-MSM, represented by Equation (5.12).
- SSM-MSM with rotation-based precoding, represented by Equation (5.13).
- SSM-MSM with amplitude scaling-based precoding, represented by Equation (5.14).
- SSM-MSM with joint rotation and amplitude scaling-based precoding, represented by Equation (5.15)

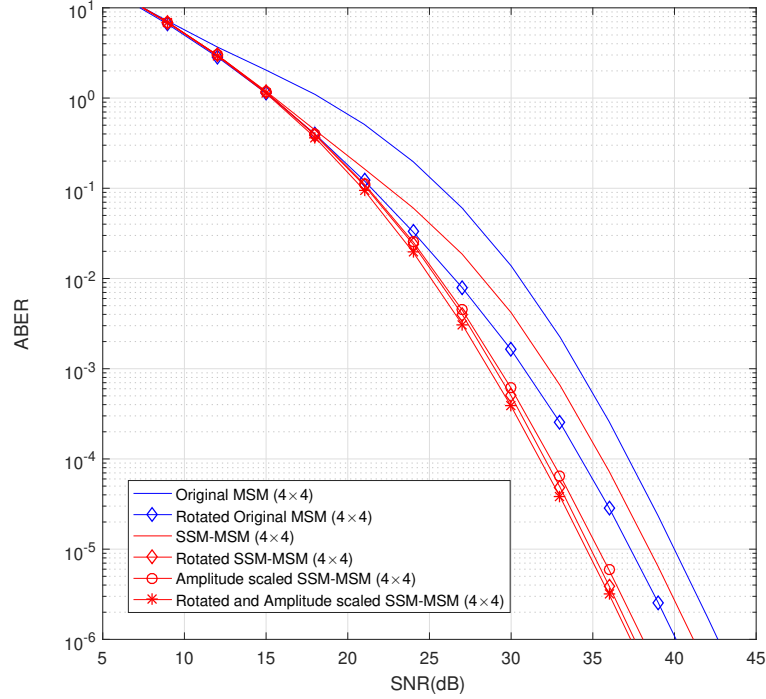


Figure 5.2. Performance of Precoded SSM-MSM over correlated Rayleigh channels with $\alpha_t = 0.99$, $N_t = 4$, $N_a = 2$, $N_r = 4$, and $M = 16$.

Figure 5.2 illustrates the results for Rayleigh fading channels with a high transmit correlation parameter of $\alpha_t = 0.99$. At an ABER of 10^{-4} , conventional rotation-based precoding achieves a 3 dB gain compared to the original MSM. At this BER level, SSM-

MSM provides a 2 dB improvement over conventional MSM, but with rotation-based precoding, the gain increases to 5 dB over conventional MSM. Under these simulation conditions, similar results were obtained when SSM-MSM used amplitude scaling-based precoding or joint phase rotation and amplitude scaling-based precoding.

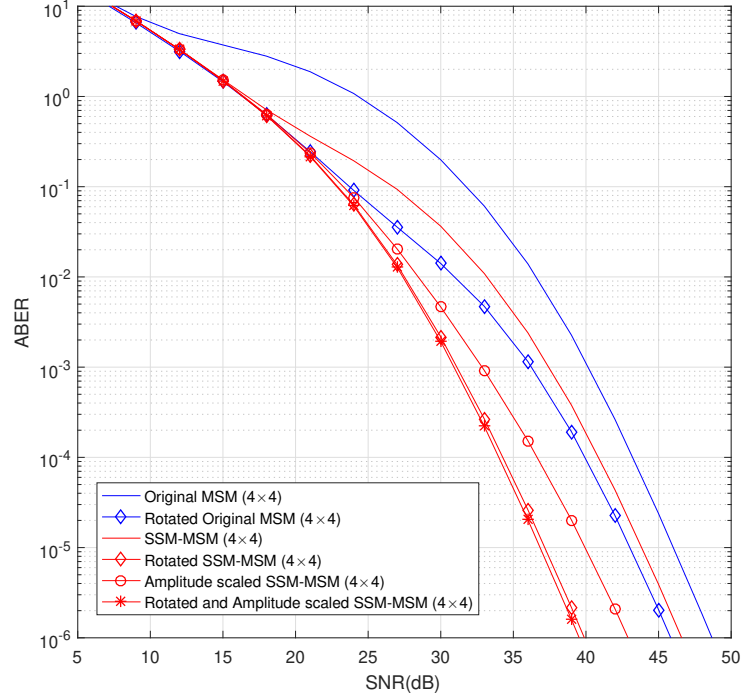


Figure 5.3. Performance of Precoded SSM-MSM over correlated Rician channels with $\alpha_t = 0.99$, $N_t = 4$, $N_a = 2$, $N_r = 4$, and $M = 16$.

Next, Figure 5.3 presents results for Rician fading channels (with $K = 3$) and a strong transmit correlation parameter $\alpha_t = 0.99$. In comparison to Rayleigh fading channels, the precoding gain is significantly higher in this scenario. The gain from conventional rotation-based precoding is approximately 3 dB, while the original SSM-MSM configuration from Equation (5.12) achieves a 2.5 dB gain. However, with rotation-based precoding, SSM-MSM realizes an impressive 9 dB improvement over conventional MSM. Although amplitude scaling-based precoding gains less, it still achieves a notable 6 dB enhancement over conventional MSM. Furthermore, joint phase rotation and amplitude scaling-based precoding offer slightly higher gains than phase rotation-based precoding alone.

With a lower transmit correlation parameter of $\alpha_t = 0.9$, the results for Rayleigh fading channels are shown in Figure 5.4, and for Rician fading channels in Figure 5.5. As depicted in Figure 5.4, the differences among the various precoding schemes are minimal, all providing less than a 1 dB improvement, except for joint phase rotation and amplitude scaling-based precoding, which achieves an additional 0.5 dB gain compared to the others. However, even with this correlation parameter, Figure 5.5 demonstrates that the proposed precoding technique yields a significant enhancement on Rician channels. Compared to conventional MSM, SSM-MSM gains only 1 dB, but with rotation-based precoding or amplitude scaling-based precoding, the gain increases by 2.5 dB, resulting in a total gain of 3.5 dB over conventional MSM. Using joint phase rotation and amplitude scaling-based precoding, SSM-MSM achieves a gain of over 4 dB over conventional MSM.

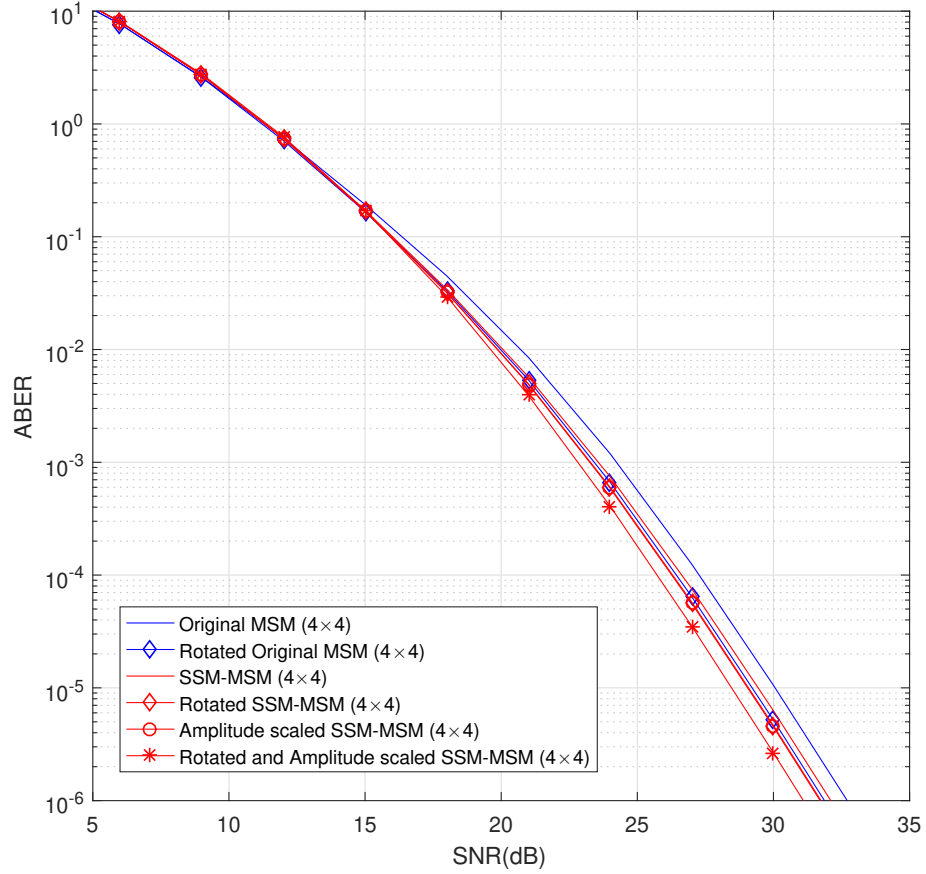


Figure 5.4. Performance of Precoded SSM-MSM over correlated Rayleigh channels with $\alpha_t = 0.9$, $N_t = 4$, $N_a = 2$, $N_r = 4$, and $M = 16$.

At this juncture, it is essential to provide an intuitive explanation for why precoding yields a greater performance enhancement on Rician channels compared to Rayleigh channels. Referring to (5.5), with a Rician factor $K = 3$, the fixed component of the received signal comprises 75% of the energy. This means that the random component, which differentiates signals received from various antennas, accounts for only 25% of the total energy. Consequently, SM performance is not only reduced on Rician channels, but sensitivity to antenna and channel correlations is also heightened. Thus, precoding, which mitigates these correlations, results in more substantial gains on Rician channels than on Rayleigh fading channels.

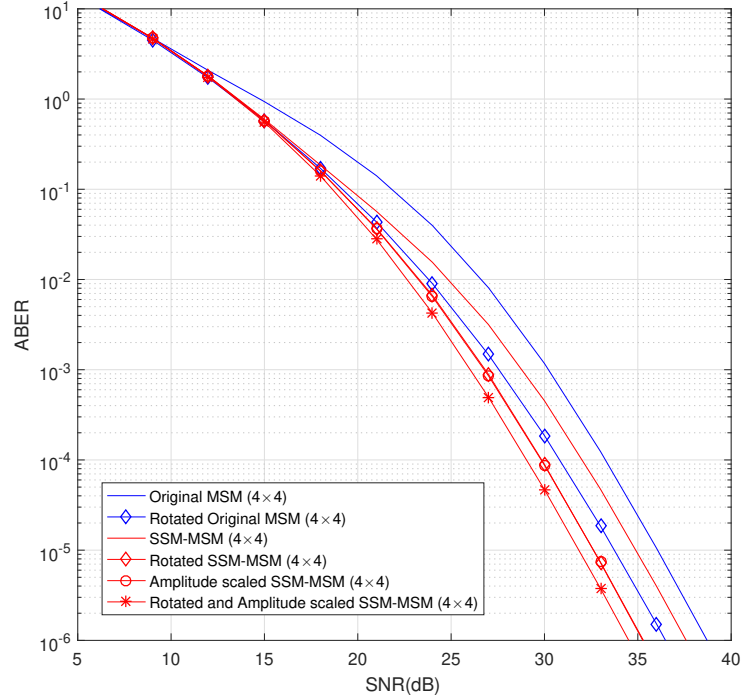


Figure 5.5. Performance of Precoded SSM-MSM over correlated Rician channels with $\alpha_t = 0.9$, $N_t = 4$, $N_a = 2$, $N_r = 4$, and $M = 16$.

5.5. Conclusions

We have developed an innovative precoding method to mitigate the effects of transmit antenna and channel correlations in multi-stream SM. This approach leverages the concept of signature constellations, but rather than assigning a unique constellation to each antenna, a set of signature constellations is allocated to each active antenna

combination. Additionally, phase rotation and/or amplitude scaling are employed to address any singularities, even under conditions of full correlation. Using 16-point signal constellations, 4 transmit antennas, and two-stream SM, we examined performance on highly correlated Rayleigh and Rician fading channels. The results demonstrate that the proposed technique significantly outperforms conventional rotation-based precoding, particularly in Rician channels. Future research will aim to extend this technique to MSM with a greater number of transmit antennas.



6. PRE-PROCESSED SPATIAL MODULATION FOR BACKSCATTER COMMUNICATION

6.1. Introduction

Backscatter communications (BC), a novel ultra-low-power communication paradigm, was introduced in [44–47]. This technology fundamentally utilizes tiny tags that harvest energy from transmitted signals to receive, modulate, and reflect radio frequency (RF) signals. BC is particularly appealing for its potential to operate without batteries and its efficient use of the spectrum, making it a promising solution for powering internet of things (IoT) applications. This has sparked substantial and continuous research interest. BC systems are broadly divided into three main categories based on their configuration: monostatic, as seen in radio frequency identification (RFID) systems [98–101], bistatic [102], [103], and ambient backscatter communication (AmBC) systems [45]. In a monostatic architecture, both the RF source and the receiver are combined into a single unit, unlike the bistatic architecture where these components are physically separated. This separation in bistatic systems, which is explored in depth, leads to enhanced performance in a cost-effective way, largely due to its simpler design compared to the monostatic approach. In AmBC systems, similar to the bistatic setup, the transmitter is detached from the backscatter receiver. However, AmBC systems do not have a dedicated RF source; they utilize existing ambient RF sources like cellular base stations and TV towers. This ambient setup effectively uses the spectrum by leveraging already present RF signals. Yet, this configuration faces a challenge with direct-link interference from these ambient sources, which adversely affects the detection performance and limits the transmission range. These systems operate by switching the backscatter tag’s impedance between reflective and absorptive states to transmit information, all without generating RF signals independently, thereby conserving power. Such efficiency is particularly advantageous for battery operated IoT devices, as it extends their life and reduces the need for frequent battery replacements. This chapter primarily examines a bistatic BC system, chosen for its straightforward design which offers improved performance at a lower cost compared to

its monostatic counterpart. Additionally, the bistatic setup reduces interference from ambient signals, a notable challenge in AmBC systems.

Backscatter tags are divided into three categories: passive, semi-active, and active, based on their energy requirements [104–108]. This study employs semi-active tags [109], which represent an intermediary technology between passive and active tags, as utilized in [110]. Similar to passive tags, semi-active tags rely on RF signal excitation, but they also contain an internal power source or battery. This additional power supports sensors or memory storage, enhancing reliability and accessibility similar to active transmitter tags. This capability allows for the development of more complex applications, albeit at the cost of higher expenses and a shorter lifespan compared to passive tags. In this context, the backscatter device discussed here includes a semi-active transmitter tag, a switched load impedance, a receiver, a micro-controller, a signal processor, and an energy harvester.

Like other wireless technologies, the push for better reliability, spectral efficiency, and broader transmission range in IoT networks motivates the evolution of sophisticated BC strategies. The objective is to enhance these attributes while maintaining the system's inherent cost-effectiveness and low-power consumption. In addressing this challenge, the incorporation of MIMO technology into BC systems is anticipated to be key in Beyond-5G networks. This integration is expected to offer significant improvements, including increased spectral efficiency, higher data transmission rates, and lower error probabilities, as highlighted in studies such as [101], [111–113].

A promising approach to further boost these capabilities involves the use of index modulation (IM), a specific technique within MIMO technology, which has been widely studied in recent research as demonstrated by references [114–118]. SM, initially introduced in [1] as an early form of IM, uniquely transmits information by utilizing both the signal dimension and the antenna indices. In [114] and [116], proposals for SM and generalized SM-based (GSM) AmBC systems have been made, where the index of the transmitter tag is used as an additional dimension for data transmission. [115] discusses SM and GSM-based bistatic BC strategies, where activating one or more tag antennas helps improve spectral efficiency and reliability. This work also

introduces a GSM scheme incorporating space-time diversity using Alamouti coding, referred to as GSMA. Additionally, advanced BC transmission methods are introduced in [119] and [120]. [119] addresses a multi-access issue in the uplink of a symbiotic AmBC non-orthogonal multiple access (NOMA) system with a multi-user blind detection algorithm that utilizes interference cancellation, showing enhanced performance even when the number of active backscatter devices is unknown. Furthermore, [120] explores a novel method using high-order modulation with M-PSK for backscattering to increase data rates and details the design of an optimal multilevel energy detector, providing a closed-form formula for calculating the symbol error rate.

In the realm of SM-MIMO communications, one sophisticated method involves implementing precoding at the transmitter to shape the transmitted information bits as desired. This concept has been extensively studied in works such as [11], [31], [33] and [121–125]. The studies [11] and [33] introduced adaptive constellation shaping precoding techniques aimed at increasing robustness against detrimental channel effects like correlation and line-of-sight components. Other researches as [31], [121] and [123–125] involve the use of known channel state information at the transmitter (CSIT) to design linear precoders that mitigate channel impacts on the transmitted symbols. Another innovative approach called pre-processing aided receive spatial modulation (PSM) has been introduced in communication systems, as described in [31], [121] and [123–125]. PSM, a variant of SM, transmits additional information bits by strategically selecting the indices of the receive antennas rather than the transmit antennas. Specifically, with CSIT, the transmitter can direct transmit power precisely to a single receive antenna, using precoding to convey information bits by the index of the received antenna alongside the symbol from a traditional modulation constellation. This strategy mitigates channel effects through pre-processing and achieves better detection performance compared to traditional SM techniques used at the transmitter. Given the advantages of the PSM scheme, this thesis chapter aims to expand this transmission technique to the bistatic BC system.

While advanced SM-based BC schemes utilizing tag-side pre-processing have been discussed in [115] and [118], focusing on space-time codebook design for diversity gains, the use of tag-side pre-processing in SM-BC systems for channel mitigation like pre-

equalization remains relatively underexplored. This work seeks to harness the benefits of a bistatic backscatter architecture with tag-side pre-processing-assisted SM transmission, establishing a comprehensive mathematical framework and introducing a system model suitable for a wide range of next-generation IoT applications. In the proposed scheme, termed tag-side pre-processing aided SM-BC (PSM-BC), a subset of incoming bits destined for the tags addresses a specific receiver at the RN, while the remaining bits select the modulated signal. Subsequently, the modulated signals are transmitted after being scaled by a pre-processing matrix, designed based on the targeted reader and following either the zero-forcing (ZF) or minimum mean-squared error (MMSE) criteria.

The contributions of this chapter are as follows:

- The main innovation of this research is the integration of PSM within BC. Implementing PSM is key to achieving diversity via precoding techniques such as zero-forcing (ZF) or minimum mean-squared error (MMSE) in BC systems. This approach not only simplifies the design of transmitters and enhances energy efficiency but also surpasses traditional SM architectures in BC by effectively countering channel impairments and improving detection capabilities. Moreover, PSM-BC shows greater resilience to channel estimation errors compared to other PSM applications in different communication environments. Additionally, the use of PSM, a sophisticated MIMO technique, significantly boosts the overall efficiency of BC systems, positioning them as a potent solution for 5G IoT applications.
- Furthermore, this work provides a theoretical performance analysis of the proposed system, detailing upper bounds on PEP and BER. Simulation results corroborate the improved performance of the techniques introduced, highlighting the advantages under various scenarios. In comparative analyses, the PSM-BC system is evaluated against standard SM-BC schemes concerning complexity, energy efficiency, and BER performance. Notably, in scenarios of erroneous channel estimation, the PSM-BC system demonstrates increased robustness to imperfect channel estimations compared to conventional SM-based BC systems. The proposed method not only shows reduced computational complexity but also de-

livers enhanced performance over benchmark systems, further establishing an energy-efficient communication framework as evidenced by both theoretical and simulation results.

6.2. System Model

We consider a bistatic BC system as shown in Fig. 6.1 that is comprised of an RF source equipped with a single antenna, a backscatter transmitter (BT) equipped with N_t semi-active transmitter tag antennas and a RN with N_r antennas, where $N_t \geq N_r$ is assumed. Additionally, the number of receive antennas is assumed to be a power of 2, and the system employs an M -ary digital modulation scheme.

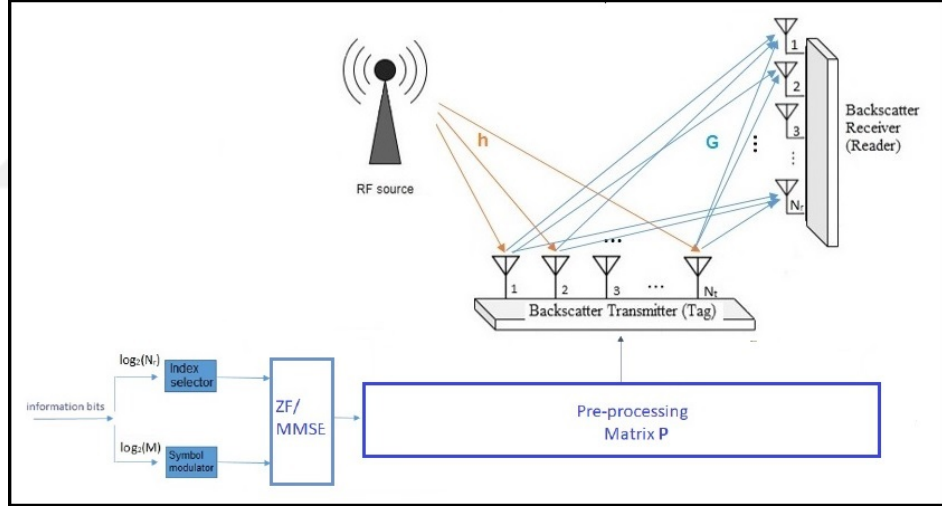


Figure 6.1. System model for PSM based BC system.

As shown in the figure, an un-modulated continuous wave (CW) signal is transmitted from the RF source, which is being used to wake up the semi-active backscatter tag antennas and turn the backscattering mode on for the transmission of the information by using the battery power. A pre-processing block is implemented according to either ZF or MMSE principle within the backscatter tag unit to shape the transmitted symbols, aiming to target a single antenna at the RN. In simpler terms, in this model, all BT antennas are turned on at the same time. However, the proposed PSM-based system directs the transmission to a specific receive antenna at the RN which is ex-

pected to capture most of the signal power. This targeted approach of choosing the receive antenna is employed for transmitting data in the spatial domain. During each time slot, a single receive antenna out of the N_r antennas at the RN is selected and activated. This activation conveys an additional set of $n_r = \log_2(N_r)$ information bits. Simultaneously, in the modulation domain, $m = \log_2(M)$ bits are transmitted using an M -ary constellation. This implies that PSM transmits a total of $k_{tot} = m + n_r$ bits through backscatter transmitters within a single symbol period. For instance, PSM mapping for a 4×2 BC system employing quadrature phase-shift keying (QPSK) constellation is shown in Table 6.1.

Table 6.1. PSM mapping for 4×2 BC system employing QPSK.

Bits	Active reader index	Transmitted symbol
000	1	+1
001	1	+j
010	1	-1
011	1	-j
100	2	+1
101	2	+j
110	2	-1
111	2	-j

In the proposed precoded backscatter modulation technique, the resulting tag gains, which correspond to the pre-processed modulation symbols, are assigned to the tags via variable load impedance matching. For symbols coming from finite alphabets (e.g. quantized gain values) the gain can be adjusted by switching among multiple load impedances. Alternatively for real values a variable impedance can be employed. If complex valued tag coefficients are to be sent two impedance units are employed per tag, etc. For more detail on tag gain control through load impedance matching operations we refer the interested reader to [44], [126] and [127].

The entire communication chain from the RF source to the RN encompasses two distinct paths. The unmodulated signal travels from the RF source to the BT, form-

ing the forward channel denoted as $\mathbf{h}$ and defined as $\mathbf{h} = [h_1, h_2, \dots, h_{N_t}]^T \in \mathbb{C}^{N_t \times 1}$. The modulated signal is backscattered from the BT to the RN, through the backward channel denoted as $\mathbf{G}$ matrix and defined as $\mathbf{G} = [\mathbf{g}_1, \mathbf{g}_2, \dots, \mathbf{g}_r] \in \mathbb{C}^{N_t \times N_r}$, where $\mathbf{g}_r = [g_{r1}, g_{r2}, \dots, g_{rN_t}]^T \in \mathbb{C}^{N_t \times 1}$ for $r = 1, 2, \dots, N_r$. It is important to highlight that, this work disregards the direct link between the RF source and the reader, mainly due to the presence of obstacles and deep scattering as also considered in [115] and each antenna at the RF source, BT, and RN are statistically independent, due to the adequate spacing between the antennas. We consider block flat Rayleigh fading channel where the channel coefficients are statistically independent of each other. In other words, the coefficients for both the forward and backscattered channels are assumed to be independent and identically distributed (i.i.d.) complex Gaussian variables with zero mean and unit variance, conventionally represented as $\sim \mathcal{CN}(0, 1)$. The equivalent complete channel matrix of the system from RF source to the reader passing through the tags, by the above considerations, becomes $\mathbf{F} \in \mathbb{C}^{N_t \times N_r}$ matrix, whose r -th column is the elementwise multiplication of $\mathbf{g}_r$ with the forward channel vector of $\mathbf{h}$ and represented as $\mathbf{f}_r = \mathbf{g}_r \odot \mathbf{h}$, for $r = 1, 2, \dots, N_r$. A pre-processing matrix is essential at the semi-active transmitter tags to reconfigure the transmitted signal vector. This ensures the reception of nearly all power by the intended receiver at the RN. Therefore, a precoding matrix, denoted as $\mathbf{P} \in \mathbb{C}^{N_t \times N_r}$ is designed based on specific optimization criteria outlined in this work. The received signal at the RN can be expressed as

$$\mathbf{y} = \sqrt{P_s} \mathbf{F}^T \mathbf{P} \mathbf{x}_m^i + \nu, \quad (6.1)$$

where P_s is the average transmitted power from the RF source which is unknown to the reader. ν is an $N_r \times 1$ additive white Gaussian noise vector, whose entries are distributed according to $\mathcal{CN}(0, N_0)$, where $N_0 = 1$ is assumed in this work for simplicity. The input vector $\mathbf{x}_m^i \in \mathbb{C}^{N_r \times 1}$ contains a modulation symbol b_m on the i -th position and all other entries are zero. The modulation symbol b_m takes one out of M values from the modulation alphabet of $\mathcal{X} = \{X_1, \dots, X_k, \dots, X_M\}$ and the index of the intended receive antenna determines the value of i , ($i \in \{1, \dots, N_r\}$). The input vector should have unit power, so $\mathbb{E}\{|\mathbf{x}_m^i|^2\} = 1$. Correspondingly, the precoding matrix $\mathbf{P}$ should satisfy the condition of $\text{Tr}(\mathbf{P}\mathbf{P}^H) = N_r$, where $\text{Tr}(\cdot)$ denotes the trace of a square matrix, to ensure the unit power for the signals sent from the BT antennas.

6.3. Precoder Design

Notice that the computation of the pre-processing matrix $\mathbf{P}$ plays a crucial role, as it should be designed to maximize the output from the desired receive antenna while minimizing the interference (leakage) on other receive antennas. If equal probability for all symbols is assumed, the receiver employs a maximum likelihood (ML) detection rule to detect the intended receive antenna index and the modulation symbol. This contribution explores the application of two common transmitter pre-processing schemes: ZF and MMSE. Notice that the channel is assumed to be perfectly estimated and known at the transmitter for the design of the precoder. Channel estimation in BC systems can be carried out through various methods as shown in works such as [128–133]. Among these in this work we assume the adoption of the two-phase least squares (LS) channel estimation approach proposed in [132] and [133] for its efficiency and simplicity. In this method, the forward and backscatter channel coefficients are estimated in two subsequent and disjoint LS estimation phases. Please note that, while majority of the derivations and results are presented under the perfect channel assumption, we also consider and provide results on how channel estimation errors effect the precoder design and the overall performance of the system as well in this work. Assuming the knowledge of the channel at the transmitter and with the aid of the similarities between PSM scheme and conventional multi-user transmitter pre-processing shown in [134], [135] the pre-processing matrix $\mathbf{P}$ employing ZF criteria can be derived as

$$\mathbf{P} = \beta \mathbf{F}^* (\mathbf{F}^T \mathbf{F}^*)^{-1}, \quad (6.2)$$

where β is a scaling coefficient needed to maintain the power requirements which is computed as

$$\beta = \sqrt{N_r / \text{Tr}((\mathbf{F}^T \mathbf{F}^*)^{-1})}, \quad (6.3)$$

here $\mathbf{F}^*$ denotes the conjugation of the $\mathbf{F}$ matrix and $\text{Tr}(\cdot)$ stands for the trace operation. Similarly, employing the MMSE criteria for pre-processing, $\mathbf{P}$ is

$$\mathbf{P} = \beta \mathbf{F}^* (\mathbf{F}^T \mathbf{F}^* + \sigma^2 \mathbf{I}_{N_r})^{-1}, \quad (6.4)$$

where β is

$$\beta = \sqrt{N_r / \text{Tr}((\mathbf{F}^T \mathbf{F}^* + \sigma^2 \mathbf{I}_{N_r})^{-2} \mathbf{F}^T \mathbf{F}^*)}, \quad (6.5)$$

here $\mathbf{I}_{N_r}$ is the $N_r \times N_r$ dimensional identity matrix and $\sigma^2 = \frac{1}{P_s}$. Please notice that the same results for the ZF and MMSE can also be obtained directly from other works as well such as [136] by appropriate choice of the system parameters.

6.4. Detection and Performance Analysis

6.4.1. Detection

Given the received signal model in equation (6.1), the signals at the intended receive antenna i for ZF based pre-processing are

$$y_i = \sqrt{P_s}\beta b_m + \nu_i, \quad (6.6)$$

and at the other receive antennas j , ($j = 1, \dots, N_r; j \neq i$) for ZF based pre-processing are

$$y_j = \nu_j, \quad j = 1, \dots, N_r; j \neq i. \quad (6.7)$$

In other words, for ZF based precoding, only the signal corresponding to the intended antenna contains both signal and noise, while all others contain only noise. Considering the received signals in equations (6.6) and (6.7), the resulting decision vector, which is of length N_r , can be rewritten as

$$\mathbf{y} = \sqrt{P_s}\beta \mathbf{x}_m^i + \boldsymbol{\nu}. \quad (6.8)$$

The receiver employs a ML detection rule for identifying the targeted receive antenna index and the selected symbol from the constellation. Consequently, for ZF pre-processing, the ML detection rule by using the expression in equation (7.8) can be formulated as

$$(\hat{m}, \hat{i}) = \arg \min_{m,i} \|\mathbf{y} - \beta \sqrt{P_s} \mathbf{x}_m^i\|^2, \quad (6.9)$$

with $\hat{m}$ and $\hat{i}$ denoting the detected modulation symbol and receive antenna index respectively.

In the case of MMSE precoding, the received signal corresponding to intended receive antenna contains both the desired signal and noise. In contrast, the signal corresponding to other receive antennas contain solely noise, coupled with the residual interference introduced by the MMSE pre-processing. The signal at the intended receive antenna i for MMSE based pre-processing is given as

$$y_i = \sqrt{P_s} \mathbf{f}_i^T \mathbf{p}_i b_m + \nu_i = \sqrt{P_s} \alpha_{ii} b_m + \nu_i, \quad (6.10)$$

and other receive antennas j , for MMSE based pre-processing is given as

$$y_j = \sqrt{P_s} \mathbf{f}_j^T \mathbf{p}_i b_m + \nu_j = \sqrt{P_s} \alpha_{ji} b_m + \nu_j \quad (6.11)$$

for $j = 1, \dots, N_r; j \neq i$. Here $\mathbf{f}_i$ and $\mathbf{f}_j$ are the i -th and j -th columns of the equivalent channel matrix of $\mathbf{F}$ and $\mathbf{p}_i$ is the i -th column of the pre-processing matrix of $\mathbf{P}$. In equations (6.10) and (6.11), the coefficient $\alpha_{ii} = \mathbf{f}_i^T \mathbf{p}_i$ and $\alpha_{ji} = \mathbf{f}_j^T \mathbf{p}_i$ are introduced to simplify the notation.

Correspondingly, ML detection rule by using the equation (6.10) is expressed as

$$(\hat{m}, \hat{i}) = \arg \min_{m,i} \|\mathbf{y} - \alpha_{ii} \sqrt{P_s} \mathbf{x}_m^i\|^2, \quad (6.12)$$

where $\hat{m}$ and $\hat{i}$ are expressions for the detected modulation symbol and receive antenna index respectively.

6.4.2. Performance Analysis

The analytical BEP of the PSM-BC scheme with ML detection is computed using the well-known union-bound approach given in [86], such as

$$\text{BEP} \leq \frac{1}{(k_{tot})2^{k_{tot}}} \sum_{i=1}^{N_r} \sum_{m=1}^M \sum_{j=1}^{N_r} \sum_{n=1}^M d(x_m^i \rightarrow x_n^j) \text{PEP}(x_m^i \rightarrow x_n^j), \quad (6.13)$$

where $d(x_m^i \rightarrow x_n^j)$ and $\text{PEP}(x_m^i \rightarrow x_n^j)$ represent the number of differing bits and the PEP between the input vectors $\mathbf{x}_m^i$ and $\mathbf{x}_n^j$, respectively. where (i, m) and (j, n) denote the desired and detected receive antenna and modulation symbol index pairs respectively. The error events can be categorized into three types: i) those where only antenna index is detected with an error, ii) those where only the symbol index is detected erroneously, iii) those where both indices are detected erroneously. These

three error types correspond to distinct categories in the PEP expressions, which can be derived based on the pre-processing scheme, as outlined below.

6.4.2.1. Performance Analysis of ZF Based PSM-BC. If the detected receive antenna j is different from the intended receive antenna i , the PEP representation for the antenna error is expressed as

$$\text{PEP}(x_m^i \rightarrow x_m^j) = Q\left(\beta|b_m|/\sqrt{\sigma^2}\right), \quad (6.14)$$

where the term $Q(\cdot)$ represents the Q-function.

In case the detected modulation symbol b_n differs from the sent symbol b_m , the PEP expression for the symbol error is expressed as

$$\text{PEP}(x_m^i \rightarrow x_n^i) = Q\left(\beta \frac{(|b_m|^2 + |b_n|^2)/2 - \text{Re}\{b_m^* b_n\}}{|b_n - b_m|\sqrt{\sigma^2/2}}\right). \quad (6.15)$$

Finally if both the detected receive antenna and modulation symbol differ from the intended ones, the PEP expression becomes

$$\text{PEP}(x_m^i \rightarrow x_n^j) = Q\left(\beta \sqrt{\frac{|b_m|^2 + |b_n|^2}{2\sigma^2}}\right). \quad (6.16)$$

Plugging the PEP expressions given in equations (6.14)-(6.16) we can evaluate the BEP from equation (6.13).

6.4.2.2. Performance Analysis of MMSE based PSM-BC. In the case of MMSE-based pre-processing, the analytical expressions for the PEP associated with the three error events can be derived as follows.

If the detected receive antenna j is different from the intended receive antenna i , then the PEP expression for antenna error can be written as

$$\text{PEP}(x_m^i \rightarrow x_m^j) = Q\left(\frac{\alpha_{ii}^2 + \alpha_{jj}^2 - 2\alpha_{jj}\text{Re}\{\alpha_{ji}\}}{\sqrt{2\sigma^2(\alpha_{ii}^2 + \alpha_{jj}^2)}}|b_m|\right). \quad (6.17)$$

If the detected modulation symbol b_n is different from the sent symbol b_m , then the PEP for the symbol error is obtained as

$$\text{PEP}(x_m^i \rightarrow x_n^i) = Q\left(\alpha_{ii} \frac{|b_n - b_m|}{\sqrt{2\sigma^2}}\right). \quad (6.18)$$

Finally if both the detected receive antenna and modulation symbol are different from the intended ones, the corresponding PEP can be expressed as

$$\text{PEP}(x_m^i \rightarrow x_n^j) = Q\left(\frac{\alpha_{ii}^2 |b_m|^2 + \alpha_{jj}^2 |b_n|^2 - 2\alpha_{jj} \text{Re}\{\alpha_{ji} b_m b_n^*\}}{\sqrt{2\sigma^2(\alpha_{ii}^2 |b_m|^2 + \alpha_{jj}^2 |b_n|^2)}}\right), \quad (6.19)$$

where these expressions are incorporated into equation (6.13) to compute the BEP. α_{ii} , α_{jj} and α_{ji} , are determined based on error considerations and defined as $\alpha_{ii} = \mathbf{f}_i^T \mathbf{p}_i$, $\alpha_{jj} = \mathbf{f}_j^T \mathbf{p}_j$ and $\alpha_{ji} = \mathbf{f}_j^T \mathbf{p}_i$.

6.5. Complexity Analysis

In this section, we present a comprehensive analysis of the computational complexity associated with the PSM-based BC systems. This analysis involves a comparative assessment against conventional SM, GSM, and GSMA-based BC systems, as proposed in [115]. The quantification of computational complexity is based on the number of computations required for the optimization problem, following the methodology outlined in [137]. In the SM-BC system, the complexity of the ML search grows linearly with both the number of transmitter tag antennas and the size of the signal set. This implies that the computational load increases proportionally with these factors. This results in a complexity order of $\mathcal{O}(N_t M)$ for the proposed system model. However, in the case of PSM-based BC, the complexity order is $\mathcal{O}(N_r M)$ as indicated in [121], which implies that the ML search complexity increases linearly with the number of antennas at RN and with the size of the signal set. As mentioned in prior works on PSM systems, it is assumed that the number of transmit antennas is greater than or equal to the number of receive antennas. This assumption means that the complexity order of the PSM-based system is no higher than that of the conventional SM-BC scheme, where the number of receive antennas equals the number of transmit antennas

($N_r = N_t$). Other detection methodologies can be employed to achieve a more favorable trade-off between computational complexity and system performance. This is particularly relevant in high spectral efficiency scenarios where ML detection becomes impractical due to the challenges associated with exhaustive search. To address this issue, several low-complexity detection algorithms have been introduced in the literature for various PSM systems. These algorithms, as demonstrated in [123], [137–142], have the capability to achieve performances that are close to the ML approach while significantly reducing computational complexity. Furthermore, It is noteworthy that GSM and GSMA entail ML decoding complexities with orders of $\mathcal{O}(N_t^2 M)$ and $\mathcal{O}(2cM)$ respectively, where c represents the number of combinations at the tag-side with two active transmit antennas, as elaborated in [122]. It can be readily confirmed that the PSM-BC exhibits a lower complexity order compared to GSM-BC. Additionally, given that $c \geq N_t$ for $N_t \geq 4$, the complexity order of the proposed PSM-BC system is smaller than that of GSMA-based BC. More specifically, for a BC system with 4 antennas at the tag-side and 2 antennas at the RN, employing QPSK for signal transmission, when conventional SM is utilized, the complexity order will be $\mathcal{O}(16)$. In the scenario of GSM-based BC, the complexity order can be demonstrated as $\mathcal{O}(64)$. On the other hand, for GSMA-based BC, where $c = 6$, the complexity order is $\mathcal{O}(48)$. For the proposed system, which utilizes the indices of the antennas at the RN to transmit information, the complexity order is $\mathcal{O}(8)$. A comprehensive analysis of complexity reveals that the proposed pre-processing-based system achieves a lower complexity order compared to previously designed systems. Furthermore, the optimization of the pre-processing matrix introduces only a marginal increase in transmitter memory. The complexity orders for all aforementioned systems are given completely in Table 6.2.

Table 6.2. Complexity order for ML detection in different BC systems.

ML detection	Order of Complexity
SM-BC	$\mathcal{O}(N_t M)$
GSM-BC	$\mathcal{O}(N_t^2 M)$
GSMA-BC	$\mathcal{O}(2cM)$
PSM-BC	$\mathcal{O}(N_r M)$

6.6. Simulation Results

In this section, simulation results are presented to demonstrate the advantages achieved through the implementation of the proposed PSM approach within the established BC system model in terms of improved BER performance, energy efficiency and robustness to channel estimation errors. This validation serves to corroborate the accuracy of our analytical expressions. Our results are categorized into three parts. Firstly, in the initial subsection, the BER outcomes of the proposed system are represented, comparing with conventional benchmark approaches in a scenario with perfect channel estimation. Additionally, the analytical expressions are validated through Monte-Carlo simulations. The second subsection focuses on BER results under imperfect channel estimation, presenting a performance comparison between conventional and pre-processed SM-BC systems. Lastly, the third subsection articulates the energy efficiency of the designed system, accompanied by a comparison with the conventional system.

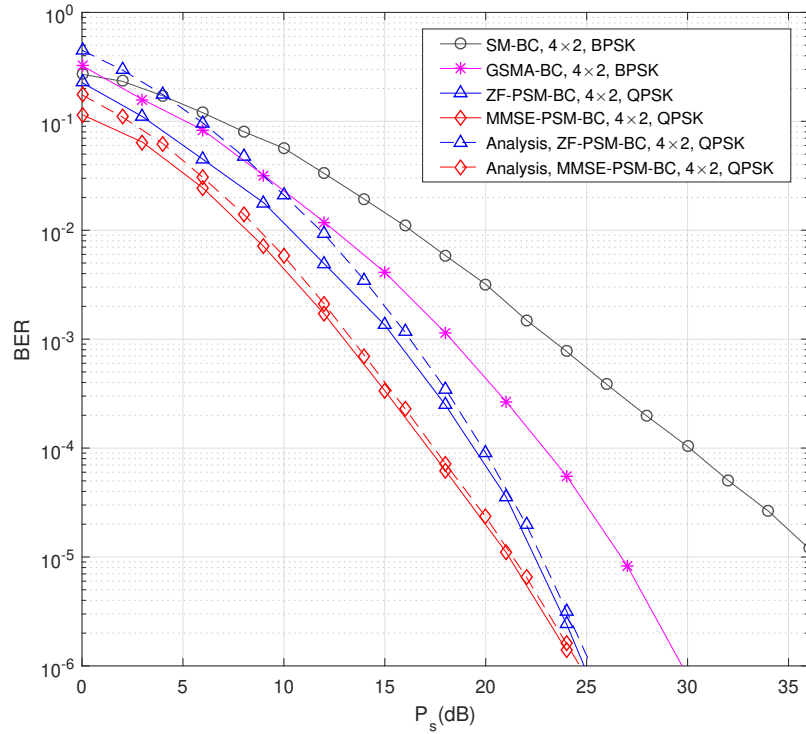


Figure 6.2. BER of PSM-BC at $R = 3$ bits/s/Hz.

6.6.1. BER Performance in Perfect Channel Estimation

The performance in terms of BER is evaluated and these results are compared with SM-based BC system and comparisons are extended to include GSM-BC and GSMA-BC in specific scenarios. This analysis emphasizes the performance enhancements offered by our proposed system compared to those detailed in [115]. All results consider the Rayleigh fading channel, with channel information at the transmitter. In each figure, same system configuration is assumed with the same spectral efficiency. In the following figures, the dashed lines are used to represent the analytical results whereas the solid lines are utilized to show the Monte-Carlo simulation results to confirm the accuracy of the theoretical analysis. In Figure 6.2, the performance of ZF and MMSE-based PSM-BC systems are compared with that of SM-BC and GSMA-BC designed in [115], all operating at the spectral efficiency of 3 bits/s/Hz. In GSMA system two antennas are activated in two time slots and two symbols are transmitted by these antennas which are chosen from the M-ary constellation diagram. More specifically GSMA can be defined as the expansion of GSM in the time domain, by applying the Alamouti coding scheme in two out of the multiple available antennas. This comparison serves to highlight the enhancements introduced by the proposed precoding scheme.

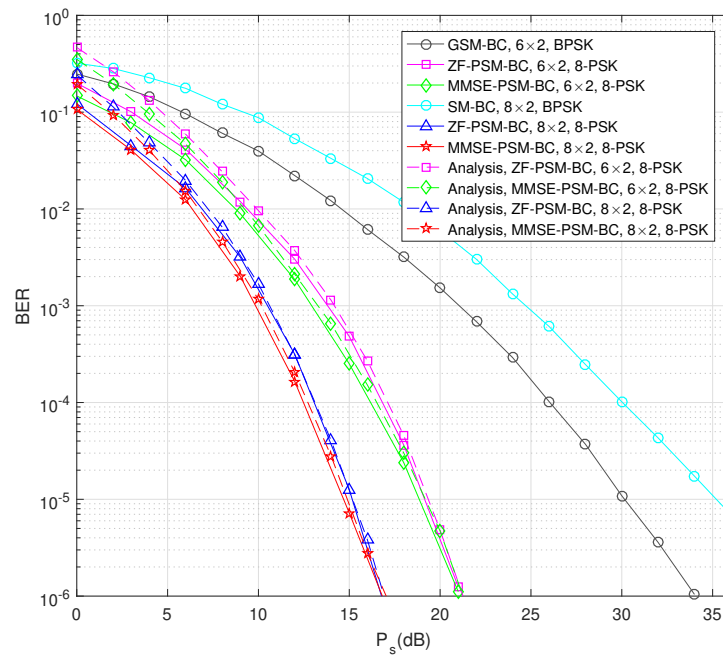


Figure 6.3. BER of PSM-BC system at $R = 4$ bits/s/Hz.

For the BC systems considered, comprising 4 tag antennas and 2 antennas at the RN, the figure illustrates a noteworthy enhancement in terms of BER performance. This enhancement is especially notable when MMSE based precoding is being utilized. Specifically, at a BER level of 1×10^{-5} , the ZF-based PSM-BC system exhibits a performance improvement of 14 dB and 4 dB in comparison to the SM-BC and GSMA-BC systems, respectively. Moreover, at this BER level, the MMSE-based PSM-BC system achieves an additional gain of 2 dB over the ZF-PSM-BC system. This implies that ZF scheme suffers from the problem of noise amplification which significantly degrades the performance. However, as the transmit power increases, there is a convergence in the performance of MMSE and ZF as it is expected from the equations. Comparisons can be done for higher order systems also, however as the complexity of the GSMA system increases dramatically with the number of transmit antennas, the results for high spectral efficiencies is not included for the GSMA scheme in this chapter.

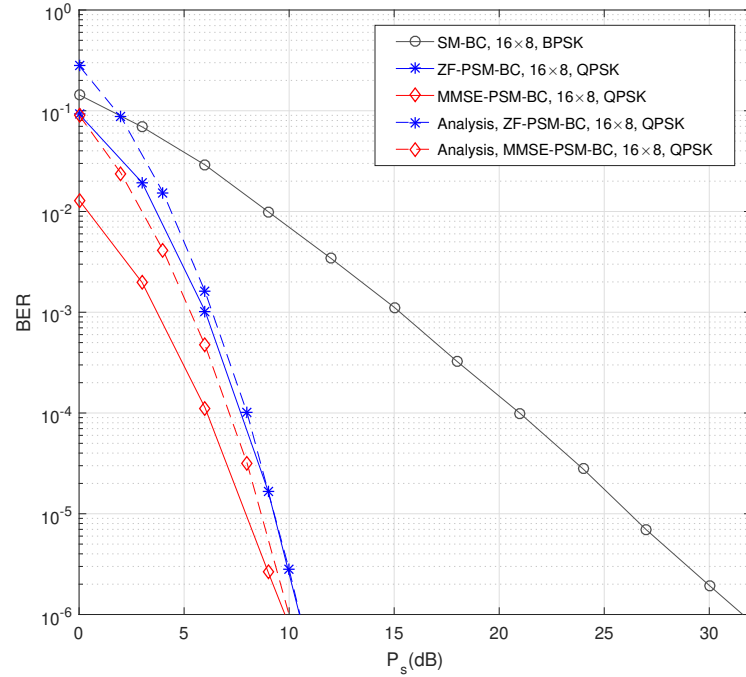


Figure 6.4. BER PSM-BC system at $R = 5$ bits/s/Hz.

In Figure 6.3, a performance comparison for the spectral efficiency of 4 bits/s/Hz is provided. Notably, to achieve this spectral efficiency, we focus on 6×2 and 8×2 systems respectively to present the behavior of the system, while varying the number of

antennas in the BT. In the case of 6×2 system, the comparison involves the proposed scheme and GSM-BC introduced in [115], utilizing the same number of tag and RN antennas. This substitution is made as conventional SM-BC operates only when the number of tags or antennas is a power of 2. In GSM-BC, two tags are activated simultaneously to transmit information by backscattering the RF source signal. The figure illustrates that the performance enhancement provided by the proposed PSM-BC system is more prominent when a larger number of tags are used. Specifically, the proposed system outperforms GSM-BC by 11 dB at a BER level of 1×10^{-5} . Furthermore, in the same figure, a comparison between the 8-tag SM-BC and PSM-BC is presented. It is clear that PSM-BC surpasses the conventional SM scheme by approximately 20 dB at a BER level of 1×10^{-5} .

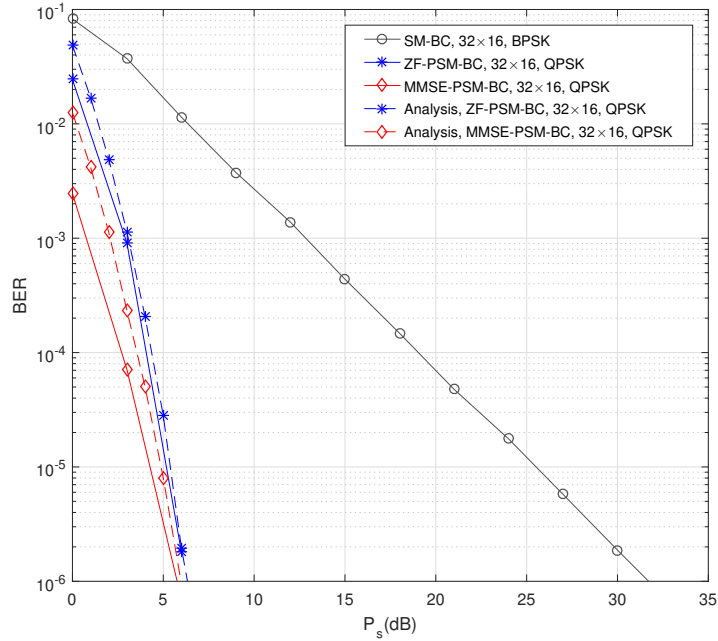


Figure 6.5. BER of PSM-BC system at $R = 6$ bits/s/Hz.

In both figures, the upper-bound curves, which are based on theoretical calculations, closely align with the simulation results in the moderate power and BER regions. These regions are typically the key operational points of interest for the analyzed systems. In Figure 6.4, the performance is depicted for the spectral efficiency of 5 bits/s/Hz. This figure illustrates a notable performance improvement of 19 dB and 17 dB for MMSE and ZF pre-processing, respectively, at a BER level of 1×10^{-5} .

Figure 6.5, illustrates the enhanced performance of the proposed approach at higher spectral efficiency levels. Specifically, considering a spectral efficiency of 6 bits/s/Hz, the analysis focuses on a 32×16 PSM-based BC configuration, employing QPSK modulation. For both MMSE and ZF-based PSM systems, it is observed that the mentioned pre-processing schemes achieve a BER level of 1×10^{-5} at approximately 21 dB and 20 dB lower transmitted powers, respectively.

6.6.2. BER Performance for Imperfect Channel Estimation

In the preceding part, our assumption centered around the transmitter tag possessing perfect knowledge of the CSI, and the design of the precoder was relied on the idealized scenario of perfect channel estimation. However, in practical implementations, obtaining a perfect channel matrix at the transmitter is often unattainable due to inherent errors in the channel estimation process. To explore the dynamics of the PSM-BC system under the influence of imperfect channel estimation, we consider an estimation error matrix denoted as Δ to the authentic channel matrix $\mathbf{F}$. Consequently, this results in an erroneous channel matrix, denoted as $\mathbf{F}_e$, expressed as

$$\mathbf{F}_e = \mathbf{F} + \Delta, \quad (6.20)$$

where Δ is an $N_t \times N_r$ error matrix, where its elements are modeled as i.i.d. circular complex Gaussian random variables with zero mean and variance of σ_e^2 . Notice that this channel error model has been frequently associated with LS channel estimators such as the one assumed to be used in this work and also in works such as [143], [144] and [125].

In the case of erroneous channel estimation, the optimization of the $\mathbf{P}$ matrix and the determination of the power normalization factor of β are conducted using the erroneous channel matrix $\mathbf{F}_e$ instead of the actual channel matrix of $\mathbf{F}$. Two different scenarios are viable for addressing erroneous channel estimation: the first involves a fixed σ_e^2 , where the variance of the error matrix remains constant while signal-to-noise ratio (SNR) values change, to show the pure effect of the imperfect channel estimation on the error performance. The second scenario introduces a variable σ_e^2 , where the variance of the estimation error is dependent on SNR variations.

For imperfect channel estimation case, ZF based pre-processing matrix derivations can be written as

$$\mathbf{P}_{ZF,e} = \beta_{ZF,e} \mathbf{F}_e^* (\mathbf{F}_e^T \mathbf{F}_e^*)^{-1}, \quad (6.21)$$

and MMSE based pre-processing matrix as

$$\mathbf{P}_{MMSE,e} = \beta_{MMSE,e} \mathbf{F}_e^* (\mathbf{F}_e^T \mathbf{F}_e^* + \sigma^2 \mathbf{I}_{N_r})^{-1}, \quad (6.22)$$

while the β coefficient for ZF based pre-processing is given as

$$\beta_{ZF,e} = \sqrt{N_r / \text{Tr}((\mathbf{F}_e^T \mathbf{F}_e^*)^{-1})}, \quad (6.23)$$

and for MMSE based pre-processing

$$\beta_{MMSE,e} = \sqrt{N_r / \text{Tr}((\mathbf{F}_e^T \mathbf{F}_e^* + \sigma^2 \mathbf{I}_{N_r})^{-2} \mathbf{F}_e^T \mathbf{F}_e^*)}. \quad (6.24)$$

Subsequently the vector representing the received signal at the RN in the context of imperfect channel estimation is formulated as

$$\mathbf{y} = \sqrt{P_s} \mathbf{F}^T \mathbf{P}_e \mathbf{x}_m^i + \nu. \quad (6.25)$$

In Figure 6.6, the BER results are shown for erroneous channel estimation case with constant variance of $\sigma_e^2 = 0.01$, together with the comparison between the proposed PSM-BC architecture and conventional SM and GSM based BC systems. As expected, the imperfect estimation of the channel degrades the performance of the conventional and pre-processed BC systems considerably. Additionally, an error floor begins at high SNR values owing to the channel estimation error. Moreover it is evident that the PSM-BC system outperforms the conventional SM and GSM based BC counterpart systems in the proposed spectral efficiencies.

Based on the observations from Figure 6.6, PSM-BC systems perform better than benchmark systems both in the moderate BER levels and also in the error floor region. One conclusion to draw from these observations is that the application of pre-processing at the tag-side in order to achieve receive SM in BC systems not only provides improved performance, energy efficiency and lower complexity but also results in improved robustness against channel estimation errors. In other words, the advantages of the

application of PSM in to the BC becomes more evident if channel estimation errors are present as shown in Figure 6.6. A final observation is that since at high SNR values the constant estimation error dominates the performance and cause the error floor, the performances of ZF and MMSE precoded SM-BC systems approximate to each other as expected.

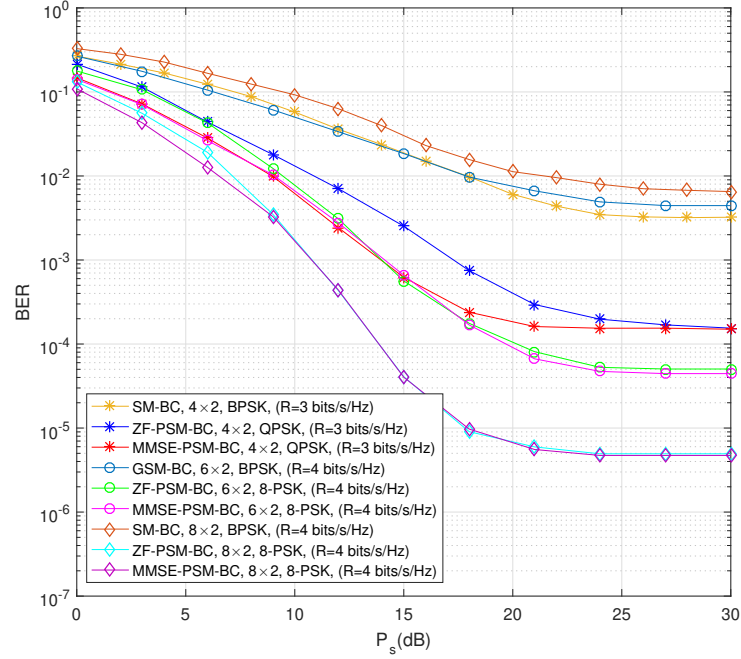


Figure 6.6. BER of PSM-BC system under imperfect channel estimation with channel estimation error variance of $\sigma_e^2 = 0.01$.

6.6.3. Energy Efficiency Results

Finally, in Figure 6.7, we also depict the energy efficiency versus the spectral efficiency performance of the proposed system at BER level of 1×10^{-4} . As seen in the figure, the proposed pre-processing aided SM technique consumes less power or operates at lower energy levels compared to the benchmark SM based approaches. Especially as the spectral efficiency increases, the difference between conventional and proposed precoded systems becomes more pronounced. For instance, at a spectral efficiency of 4 bits/s/Hz, a ZF-based BC system requires 11.5 dB less power than a conventional system to achieve the same BER level. For MMSE-based system, this difference is even more, that is there is a 14 dB difference in the required SNR. Similar conclusions

can be also be drawn at other configurations corroborating the fact that the proposed system not only offers better BER results but also provides higher energy efficiency (considering constant noise levels) than conventional approaches.

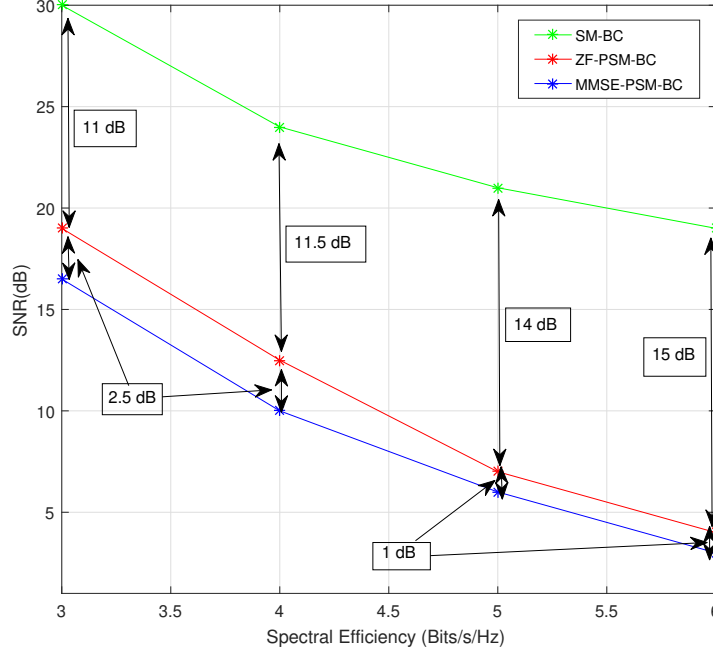


Figure 6.7. Energy efficiency of SM and PSM systems in BC to achieve BER level of 1×10^{-4} at different spectral efficiencies.

6.7. Conclusion

In this thesis, we introduced a tag-side pre-processing aided SM scheme, designed for bistatic BC systems operating under the CSIT mode utilizing semi-active transmitter tags. Optimization of pre-processing and signal detection within this scheme are analysed in this work. Both MMSE and ZF architectures are proposed, accompanied by theoretical error performance and complexity analyses. The theoretical derivations are validated by Monte-Carlo simulation results performed under different configurations. Our analysis and performance results demonstrate that the PSM offers an alternative way for transmitting signal in MIMO-BC systems as a novel and high efficient solution for Beyond-5G IoT technology. Results reveal that the proposed PSM-BC system offers both lower complexity, improved BER performances and energy efficiency in comparison to the state-of-the art benchmark approaches. It is also demonstrated the

performance gains of the proposed system increases by the number of antennas used on the tag-side. Please notice that utilizing semi-active tags in this design results in increased hardware complexity and cost at the transmitter side. In addition for a given channel matrix the precoding matrix is computed once and stored in the memory throughout the entire transmission. This results in the usage of memory elements and gain control circuitry in the backscatter tags. As seen in the complexity analysis and simulation results, this approach offers lower complexity at the receiver and improved BER and power gains, albeit at the expense of slightly increased complexity on the tag-side.

We also extend our tag-side precoding design to the imperfect channel estimation case and present how the BER performance is affected by erroneous channel estimation. It can be concluded that the PSM technology is more robust to the channel estimation errors than the conventional SM based BC system. An interesting direction for future work is to apply PSM technique to AmBC system, as well as to extend our work to RIS-assisted PSM based BC systems as in the next chapter.

7. RIS-AIDED PRECODED SPATIAL MODULATED BACKSCATTER COMMUNICATION

7.1. Introduction

The concern of managing wireless channels is experiencing a notable upswing in attention, with a particular focus on facilitating the exchange of data between smart devices without utilizing network resources. This stands as a crucial prerequisite for the efficient functioning of smart systems in advanced 5G and potential 6G applications [35], driven by the creative deployment of reconfigurable intelligent surfaces (RIS) that empower the adjustment of signals in an adaptive manner [36–40]. The fundamental principle of RIS, also known as software-defined meta-surfaces, involves the intelligent manipulation of electromagnetic waves using passive elements, each with the capability to adjust the signal’s amplitude and/or phase independently. This enables precise three-dimensional reflection beamforming, without requiring active radio frequency (RF) signal generation. Despite the implementation of meta-surfaces, what makes the RIS technology in terms of energy efficient, stems from its ability to amplify and forward incoming signals without the need for a power amplifier, which is accomplished through the careful design of phase shifts applied by each reflecting element to efficiently merge reflected signals. Additionally, the compact hardware design of RIS structures facilitates effortless integration into the communication environment, allowing for straightforward deployment on ceilings of factories and indoor spaces, laptop cases, and even integration into clothing [41]. RIS consists of a large number of low-cost nearly passive elements, which create additional indirect links between transmitter and receiver in addition to the capability to substantially decrease energy consumption by increasing signal strength, thereby empowering devices to function at reduced power levels. Furthermore, RIS exhibits significant promise in addressing the growing needs for increased data rates, extended coverage, and enhanced energy efficiency in wireless networks. By dynamically adapting to shifting channel dynamics and user needs, RIS strengthens system robustness, simultaneously countering multi-path fading, augmenting received signal quality, and elevating overall reliability [40], [42], [43]. Beyond

the dynamic control of RF waves, the integration of energy harvesting capabilities within devices is foreseen as a critical necessity in a wide range of wireless applications. Within this framework, backscatter communication (BC) [44–47] emerges as a pioneering technology for energy harvesting, which utilizes small tags capable of harnessing energy emitted by the transmitter. These tags then perform tasks such as receiving, modulating, and backscattering the RF signals. The potential for battery-free and spectrum-efficient communication in ultra-low power consumption mode, positions BC as a feasible technology to energize applications within the internet-of-things (IoT). This capability has garnered substantial and continually growing research interest. This marks a significant shift in the domain of sourcing sustainable power, indicating a fundamental change in how energy is obtained and employed. The distinction between BC and RIS lies in their respective functionalities. Unlike BC, which involves low-cost passive tags reflecting and modulating RF signals while harvesting energy [46], the primary function of RIS is to enhance signal quality, between the source and destination and it does not transmit any information of its own [48].

In the context of BC, the backscatter tag undergoes impedance alteration, toggling between reflective and absorptive states to convey information. Notably, this occurs without the need for independent RF signal generation, resulting in minimal power consumption. As a consequence, such systems prove highly compatible with battery-powered IoT devices, extending operational lifetimes and minimizing the necessity for frequent battery changes. The architecture of BC systems can be categorized in three distinct configurations, namely monostatic, bistatic, and ambient BC (AmBC) [145]. In monostatic system, the carrier emitter and reader are integrated into a single device, as in the RF identification (RFID) example. This co-location of the carrier emitter and reader, causes round-trip path loss which limits monostatic BC systems to short-range applications. Unlike monostatic BC systems, carrier emitter and reader are separated in bistatic configuration, improving the flexibility of the system and the efficiency in cost due to simpler design than the monostatic counterpart. In the bistatic configuration, a dedicated RF source is essential, especially for long-range applications. In an AmBC system, similar to bistatic configurations, the carrier emitter is separated from the backscatter receiver. However, instead of relying on a dedicated RF source, the ambient configuration utilizes existing RF sources such as TV towers and WiFi

access points, reducing the cost and power consumption of a dedicated carrier emitter. It is worth noting that in an ambient configuration, direct-link interference caused by ambient signals adversely affects detection performance and consequently imposes limitations on the transmission range. The mentioned properties for RIS and BC systems leads the industry and academia to integrate both configurations to improve network coverage, reliability, lifetime and capacity [49], [51], [56–58], [145–147].

Beyond utilizing BC and RIS, the incorporation of MIMO technology is anticipated to play a crucial role in beyond-5G systems. Specifically within the context of BC, the utilization of multiple antennas at the tag side offers the potential to attain the necessary high spectral efficiency and high data rates. The core principle of index modulation in MIMO technology, relies on using the indices of accessible resource dimensions to convey information bits. In SM [1], a specific instance of IM, the spatial domain, which refers to the locations of multiple antennas, is employed to transmit supplementary data alongside the conventional signal modulation. This approach brings advantages such as improved spectral efficiency, reduced complexity compared to conventional MIMO systems, and resilience against certain types of signal fading. An alternative form of SM known as space shift keying (SSK) [148] has emerged as a practical transceiver architecture due to its simplified complexity and commendable performance. A preprocessing aided spatial modulation (PSM) scheme is introduced in [124], which uses the indices of receiving antennas as a constellation for conveying information, in addition to incorporating a conventional amplitude phase modulation (APM) scheme. To elaborate further, the first component is conveyed through the utilization of receive antenna indices, while the second component is transmitted using the conventional APM technique. In the PSM scheme, transmitter preprocessing is employed to ascertain the targeted receive antenna. This enables the detector to extract the first information component from the spatial domain, while the second information component is retrieved through the traditional APM approach. The analysis and performance outcomes indicate that PSM presents a viable alternative for signal transmission within MIMO systems, and enhances overall BER performance, surpassing the BER performance of the SM technique. In the existing literature, numerous MIMO communication models have employed SM/SSK. For more detailed information, we direct the reader to [6], [13], [62], [76], [77], [79], [85], [149]. Specifically, in [114–118],

the utilization of SM is explored in the context of BC applications. In reference [150], both SM and SSK techniques were employed at the receiving end, while integrating RIS as reflectors. In [151], SM is applied in meta-surface based modulation, where it was utilized to choose the elements of RIS. The capability of an RIS-assisted IM system to achieve high spectral efficiency at low SNR values was validated in [150]. Furthermore, studies in [37] and [41] demonstrated that RIS offers substantial improvements in energy efficiency compared to traditional relay-assisted communication systems. Moreover, the utilization of RIS-assisted SM in conjunction with BC for selecting specific tag antennas has been documented in [53]. In [152], an RIS-assisted multi-tag BC system over Nakagami-m fading channel is considered and outage probability and ABER is analyzed and it is shown that an increase in the number of tags and reflecting elements, results in significant improvement of the performance.

Recently a line of work is introduced which uses RIS to enhance BC performance [49–54]. For instance, the performance of AmBC system using RIS is analyzed in [51], and the BER is computed while RIS is used in the source-to-reader link, source-to-tag link and in both links. The results have shown that by increasing the number of RIS elements in the source-to-reader link, the performance degrades while increasing the RIS elements in the source-to-tag link improves the performance. In [55], a novel integration of RIS into a bistatic BC system is introduced which demonstrates significant transmit power savings in both single-tag and multi-tag systems, even with moderate RIS sizes. In [56], the signaling design and phase shift optimization of an IRS-assisted BC system in indoor settings is done to improve the detection performance. In [54], RIS-assisted backscatter communication is modeled in such a case that there is no direct link between tag and the reader and the link is just provided by the use of RIS elements, where the system performance is analysed under hardware imperfections. An experimental assessment is done for the first time in [57], to demonstrate that RIS can improve the performance of ambient backscatter systems. RIS is used in cooperative BC in [52], in which the RIS elements are either on reflecting or energy harvesting modes, where in this work a transmission design is proposed to minimize the RIS power consumption. The integration of RIS into backscatter communication systems for enabling 6G IoT networks is presented in [58].

In this chapter, we introduce the comprehensive signal model for an RIS-aided bistatic BC system, which uses PSM for transmitting information, marking the first instance where RIS is combined with pre-processing to enhance the overall performance of the BC system. In RIS-assisted bistatic configuration, the RIS's ability to reconfigure the wireless medium can enhance the performance of both the forward and backscattered links. The coherent addition of the signals results in reduced transmit power requirements, extended transmission ranges, and enhanced device lifetimes. Moreover, applying RIS to a bistatic BC system allows for flexible placement and deployment of tags, providing an additional degree of freedom. Additionally, RIS performance is not hindered by noise amplification or self-interference, since reflectors remain unaffected by these issues [42].

It is worth noting that the application of RIS to the tag-side pre-processing within SM-BC systems, aimed at mitigating channel effects (such as for pre-equalization purposes), has received limited attention in existing literature. This is despite being a well-established technique with an extensive body of work in conventional SM-based communication. In light of this, our work introduces a fresh perspective by presenting an innovative RIS assisted BC system that incorporates pre-processing techniques, involving multiple tags and multiple receivers at the reader node (RN).

7.2. System Model

A RIS-assisted bistatic BC system is proposed in this work as shown in Figure 7.1, which transmits the information bits through pre-processed spatial modulation technique. The proposed architecture is comprised of an RF source with single antenna, a BT equipped with N_t semi-active transmitter tag antennas, a backscatter receiver, also known as RN, with N_r antennas and a RIS with N reflecting elements. The RIS can be either portable or permanently fixed to a building, wall, or another structure, improving the backscatter communication of the tags. In the proposed architecture $N_t \geq N_r$ is assumed which is a necessary condition for PSM transmission scheme. In this configuration it is assumed that there is no direct link between the RF source and RIS elements and also between the RF source and the RN due to presence of obstacles and deep scattering. In addition it is assumed that the strength of the signal after two

or more reflections is negligible, which eliminates the problem of multiple reflections between tags and RIS elements, what makes the RIS elements acting like backscatter devices. In this setup, sufficient spacing between the antennas at the RF source, BT and RN is assumed, ensuring the statistical independence among antennas. As illustrated in Figure 7.1, an unmodulated continuous wave (CW) signal is emitted from the RF source activating the semi-active backscatter tag antennas and switches them to backscatter mode, which is crucial for initiating the information transmission process. This activation and transmission are powered by battery energy within the tags. To enhance the signal quality and direct the transmission accurately toward a specific antenna at the RN, a pre-processing block is integrated into the backscatter tag unit. This block is configured based on either ZF or MMSE principle. The primary objective of this pre-processing is to refine the transmitted symbols, ensuring that they are optimally shaped to communicate with a single designated antenna at the RN. More specifically, the RF source emits a CW un-modulated signal, which is being utilized by the BT antennas for information transmission. The BT uses a part of the CW signal for energy harvesting, powering its integrated circuit as a result. Then BT transmits the information to the RIS elements and RN through appropriate load modulation. Each tag is equipped with variable load impedances connected to its antenna, representing states for different symbol transmissions. It's also noteworthy that thermal noise at the tag is not considered, as the tag depends on RF reflection instead of having an active transmitter circuitry.

The entire communication pathway from the RF source to the RN consists of two distinct paths. One path involves modulation by the tags and reflection by the RIS elements, while the second path is the one that modulated by the tags and transmitted to the reader directly. Rayleigh fading is employed in the channel model, which means that the forward, backscattered and all RIS channel coefficients are independent and identically distributed (i.i.d.) complex Gaussian variables, typically denoted as $\sim \mathcal{CN}(0, 1)$. It's important to highlight that the selection of the fading model does not affect the overall analysis of this work.

In the PSM scheme employed for BC in this work, all transmit antenna tags are activated simultaneously. The information is then transmitted using the index of a

receiver antenna at the RN, along with the M -ary modulation symbol, as illustrated in Figure 7.1. In this scheme, $n_r = \log_2(N_r)$ bits are transmitted via the indices of the receive antennas. Simultaneously, $m = \log_2(M)$ bits are transmitted in the modulation domain, considering an M -ary constellation. This implies that PSM transmits a total of $k_{tot} = m + n_r$ bits via backscatter transmitters during one symbol period. The wireless fading channel between t -th BT antenna and the n -th reflecting element of the RIS is defined by $h_{t,n} = \alpha_{t,n}e^{-j\theta_{t,n}}$ and denoted by matrix $\mathbf{H} \in \mathbb{C}^{N_t \times N}$. The channel between n -th RIS reflector and the r -th reader antenna is denoted as $\tilde{h}_{n,r} = \beta_{n,r}e^{-j\phi_{n,r}}$ shown by matrix $\tilde{\mathbf{H}} \in \mathbb{C}^{N \times N_r}$, for $t = 1, 2, \dots, N_t$, $r = 1, 2, \dots, N_r$ and $n = 1, 2, \dots, N$. In addition $\mathbf{h}_d \in \mathbb{C}^{N_t \times 1}$ denotes the channel vector from the RF source to tag.

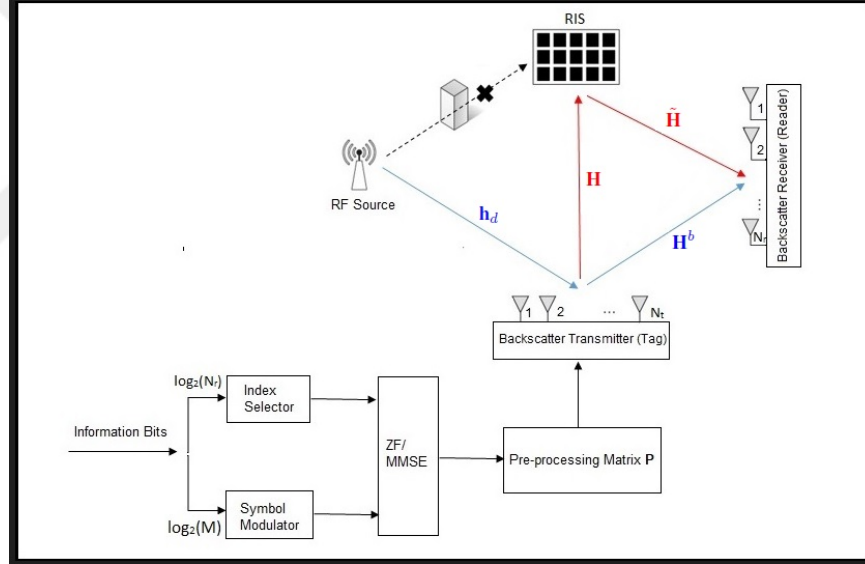


Figure 7.1. RIS assisted PSM based bistatic BC system model.

The direct channel from tags to the RN can be expressed by $\mathbf{H}^b \in \mathbb{C}^{N_t \times N_r}$. We consider block flat fading channels which are statistically independent of each other. Each RIS element, denoted as $n \in \{1, \dots, N\}$, reflects the sum of all incident signals with a phase shift, represented by ψ_n . Amplitude gain of each RIS element is assumed as unity in this work. The vector containing the phase shifts of all elements is denoted as $\boldsymbol{\Psi} = [\psi_1, \psi_2, \dots, \psi_N]^T$, with $\psi_n \in [0; 2\pi)$. The RIS phase shift matrix can be represented by the diagonal matrix of $\boldsymbol{\Theta} = \text{diag}[e^{j\psi_1} e^{j\psi_2} \dots e^{j\psi_N}]$. The equivalent channel matrix between backscatter tag and the RN is denoted by $\mathbf{f} = [\mathbf{f}_1, \mathbf{f}_2, \dots, \mathbf{f}_r] \in \mathbb{C}^{N_t \times N_r}$, where

$\mathbf{f}_r = [f_{r1}, f_{r2}, \dots, f_{rN_t}]^T \in \mathbb{C}^{N_t \times 1}$ for $r = 1, 2, \dots, N_r$, which equals $\mathbf{F} = \mathbf{H}\mathbf{\Theta}\tilde{\mathbf{H}} + \mathbf{H}^b$. The equivalent complete channel matrix of the system from RF source to the reader passing through the tags, by the above considerations, becomes $\mathbf{G} \in \mathbb{C}^{N_t \times N_r}$ matrix, whose r -th column is the elementwise multiplication of $\mathbf{f}_r$ with the forward channel vector of $\mathbf{h}_d$ and represented as $\mathbf{g}_r = \mathbf{f}_r \odot \mathbf{h}_d$, for $r = 1, 2, \dots, N_r$. In this configuration, in order to eliminate the phases of the tag-RIS-reader channel between the tag-reader pair of (t, r) , phase elements of RIS are adjusted as $\psi_n = \theta_{t,n} + \phi_{n,r}$.

In the proposed PSM method, every BT antenna is activated at the same time. However, the process is organized in such a way that only one designated receiver antenna at the RN is involved in the communication at any given time. This ensures that this chosen antenna receives the majority of the signal's power. The main goal of selecting this specific receive antenna is to facilitate data transmission through the spatial dimension. To implement this approach, two distinct strategies for transmit precoding, based on ZF and MMSE criteria, are proposed and will be thoroughly examined in the next section.

In RIS-aided backscatter communication systems, accurate channel estimation is crucial for optimizing system performance. In addition, channel estimation in RIS-assisted wireless systems is significantly more challenging than conventional systems due to the inability of passive RIS elements to sense and estimate channel information. As a result, it becomes necessary to rely on the receiver to estimate the channels backscattered by the tags towards the reader from the RF source through RIS elements, as well as directly from the RF source without reflection by RIS elements, by observing a noisy composite of the two channels. Since multiple reflections are neglected in this work, the cascaded channel estimation is performed as described in [153–157]. These methods, originally developed for RIS-assisted communications, can be directly applied to RIS-assisted backscatter communications due to their similar channel models. In the PSM scheme proposed in this work, the channel components is assumed to be perfectly known at the transmitter side, which is a feedback result from the reader in the previously mentioned channel estimation methods.

A pre-processing matrix is essential at the transmitter tags to reconfigure the transmitted signal vector what ensures the reception of nearly all power by the intended receiver at the RN. Therefore, a precoding matrix, denoted as $\mathbf{P} \in \mathbb{C}^{N_t \times N_r}$ is designed based on specific optimization criteria outlined in this work. The received signal at the RN can be expressed as

$$\mathbf{y} = \sqrt{P_s} \mathbf{G}^T \mathbf{P} \mathbf{x}_m^i + \nu, \quad (7.1)$$

where in this model, P_s is the average transmitted power from the RF source which is unknown to the reader. ν is an $N_r \times 1$ additive white Gaussian noise vector, with each component following a complex normal distribution, $\mathcal{CN}(0, N_0)$, where N_0 is set to 1 for simplicity. The input vector $\mathbf{x}_m^i \in \mathbb{C}^{N_r \times 1}$ includes a modulation symbol b_m on the i -th position and all other positions filled with zeros. This modulation symbol b_m is selected from one of M possible values in the modulation alphabet, $\mathcal{X} = \{X_1, \dots, X_k, \dots, X_M\}$ and the index of the intended receive antenna determines the value of i , ($i \in \{1, \dots, N_r\}$). The input vector should have unit power, so $\mathbb{E}\{\|\mathbf{x}_m^i\|^2\} = 1$. Correspondingly, the precoding matrix $\mathbf{P}$ should satisfy the condition of $\text{Tr}(\mathbf{P}\mathbf{P}^H) = N_r$, where $\text{Tr}(\cdot)$ denotes the trace of a square matrix, to maintain the unit power for the signals sent from the BT antennas.

7.3. Precoder Design

The computation of the pre-processing matrix $\mathbf{P}$ plays an important role, as it should be designed in a way to maximize the output power from the desired receive antenna while minimizing the interference (leakage) on other receive antennas. By the assumption of equal probability for all transmitted symbols, maximum likelihood (ML) detection rule is employed at the receiver side to detect the intended receive antenna index and the modulation symbol. The application of two commonly used ZF and MMSE transmitter pre-processing schemes are proposed in this work. Notice that the channel is assumed to be perfectly estimated and known at the transmitter for the design of the precoder. Please note that, the derivations and results are presented under the perfect channel estimation consideration, and the precoder design and the phase optimization is done based on this assumption.

Assuming the knowledge of the channel at the transmitter and with the aid of the similarities between PSM scheme and conventional multi-user transmitter pre-processing shown in [134, 135], the pre-processing matrix $\mathbf{P}$ employing ZF criteria can be derived as

$$\mathbf{P} = \beta \mathbf{G}^* (\mathbf{G}^T \mathbf{G}^*)^{-1}, \quad (7.2)$$

where β is a scaling coefficient needed to maintain the power requirements which is computed as

$$\beta = \sqrt{N_r / \text{Tr}((\mathbf{G}^T \mathbf{G}^*)^{-1})}, \quad (7.3)$$

here $\mathbf{G}^*$ denotes the conjugation of the $\mathbf{G}$ matrix and $\text{Tr}(\cdot)$ stands for the trace operation. Similarly, employing the MMSE criteria for pre-processing, $\mathbf{P}$ is obtained as

$$\mathbf{P} = \beta \mathbf{G}^* (\mathbf{G}^T \mathbf{G}^* + \sigma^2 \mathbf{I}_{N_r})^{-1}, \quad (7.4)$$

where β can be derived as

$$\beta = \sqrt{N_r / \text{Tr}((\mathbf{G}^T \mathbf{G}^* + \sigma^2 \mathbf{I}_{N_r})^{-2} \mathbf{G}^T \mathbf{G}^*)}, \quad (7.5)$$

here $\mathbf{I}_{N_r}$ is the $N_r \times N_r$ dimensional identity matrix and $\sigma^2 = \frac{1}{P_s}$. Please notice that the same results for the ZF and MMSE can also be obtained directly from other works as well such as [136] by appropriate choice of the system parameters.

7.4. Detection and Performance Analysis

Based on the received signal model specified in Equation (7.1), the outputs at the targeted receive antenna i , by using ZF-based preprocessing is as follows

$$y_i = \sqrt{P_s} \beta b_m + \nu_i, \quad (7.6)$$

and other antennas j , with $j = 1, \dots, N_r$ and $j \neq i$,

$$y_j = \nu_j. \quad (7.7)$$

This indicates that in ZF-based precoding, the signal at the designated antenna includes both the signal and noise, whereas the other antennas receive only noise.

From the signals described in equations (7.6) and (7.7), the decision vector, with a length of N_r , is

$$\mathbf{y} = \sqrt{P_s}\beta\mathbf{x}_m^i + \nu. \quad (7.8)$$

The receiver utilizes an ML detection strategy to pinpoint the intended receive antenna index and the symbol from the constellation. Thus, the ML detection rule for ZF preprocessing, based on Equation (7.6), is defined

$$(\hat{m}, \hat{i}) = \arg \min_{m,i} \|\mathbf{y} - \beta\sqrt{P_s}\mathbf{x}_m^i\|^2, \quad (7.9)$$

where $\hat{m}$ and $\hat{i}$ represent the identified modulation symbol and receive antenna index, respectively.

For MMSE preprocessing, the signal received at the intended antenna encompasses both the desired signal and noise, whereas the other antennas primarily receive noise alongside residual interference from MMSE preprocessing. The relevant signals for the designated antenna i in MMSE-based preprocessing is given

$$y_i = \sqrt{P_s}\mathbf{g}_i^T \mathbf{p}_i b_m + \nu_i = \sqrt{P_s}\alpha_{ii}b_m + \nu_i, \quad (7.10)$$

and other antennas j

$$y_j = \sqrt{P_s}\mathbf{g}_j^T \mathbf{p}_i b_m + \nu_j = \sqrt{P_s}\alpha_{ji}b_m + \nu_j, \quad (7.11)$$

for $j = 1, \dots, N_r; j \neq i$. The coefficients $\alpha_{ii} = \mathbf{g}_i^T \mathbf{p}_i$ and $\alpha_{ji} = \mathbf{g}_j^T \mathbf{p}_i$ simplify the notations. The ML detection rule using Equation (7.10) is then formulated

$$(\hat{m}, \hat{i}) = \arg \min_{m,i} \|\mathbf{y} - \alpha_{ii}\sqrt{P_s}\mathbf{x}_m^i\|^2. \quad (7.12)$$

with $\hat{m}$ and $\hat{i}$ representing the detected modulation symbol and receive antenna index, respectively. The analytical BEP of the PSM-BC scheme with ML detection is computed using the well-known union-bound approach given in [86], such as

$$\text{BEP} \leq \frac{1}{(k_{tot})2^{k_{tot}}} \sum_{i=1}^{N_r} \sum_{m=1}^M \sum_{j=1}^{N_r} \sum_{n=1}^M d(x_m^i \rightarrow x_n^j) \text{PEP}(x_m^i \rightarrow x_n^j), \quad (7.13)$$

where $d(x_m^i \rightarrow x_n^j)$ and $\text{PEP}(x_m^i \rightarrow x_n^j)$ represent the number of differing bits and the PEP between the input vectors $\mathbf{x}_m^i$ and $\mathbf{x}_n^j$, respectively. where (i, m) and (j, n) denote the desired and detected receive antenna and modulation symbol index pairs.

The decision vector formulation takes into account different error events, categorized into three types: errors in detecting only the antenna index, only the symbol index, and both indices, each corresponding to specific categories in the probability of error (PEP) calculations based on the pre-processing strategy employed, which can be found, as outlined in Tables 7.1 and 7.2 for ZF and MMSE based pre-processing respectively. α_{ii} , α_{jj} and α_{ji} , expressed in the tables are determined based on error considerations and defined as $\alpha_{ii} = \mathbf{g}_i^T \mathbf{p}_i$, $\alpha_{jj} = \mathbf{g}_j^T \mathbf{p}_j$ and $\alpha_{ji} = \mathbf{g}_j^T \mathbf{p}_i$. Plugging the PEP expressions given in Table 7.1 and Table 7.2, the BEP can be evaluated from equation (7.13).

Table 7.1. PEP calculations for pairwise error events in RIS assisted ZF-based PSM-BC.

$\text{PEP}(x_m^i \rightarrow x_m^j)$	$Q\left(\beta b_m /\sqrt{\sigma^2}\right)$
$\text{PEP}(x_m^i \rightarrow x_n^i)$	$Q\left(\frac{\beta(b_m ^2+ b_n ^2)/2-\text{Re}\{b_m^* b_n\}}{ b_n-b_m \sqrt{\sigma^2/2}}\right)$
$\text{PEP}(x_m^i \rightarrow x_n^j)$	$Q\left(\beta\sqrt{\frac{ b_m ^2+ b_n ^2}{2\sigma^2}}\right)$

Table 7.2. PEP calculations for pairwise error events in RIS assisted MMSE-based PSM-BC.

$\text{PEP}(x_m^i \rightarrow x_m^j)$	$Q\left(\frac{\alpha_{ii}^2+\alpha_{jj}^2-2\alpha_{jj}\text{Re}\{\alpha_{ji}\}}{\sqrt{2\sigma^2(\alpha_{ii}^2+\alpha_{jj}^2)}} b_m \right)$
$\text{PEP}(x_m^i \rightarrow x_n^i)$	$Q\left(\alpha_{ii}\frac{ b_n-b_m }{\sqrt{2\sigma^2}}\right)$
$\text{PEP}(x_m^i \rightarrow x_n^j)$	$Q\left(\frac{\alpha_{ii}^2 b_m ^2+\alpha_{jj}^2 b_n ^2-2\alpha_{jj}\text{Re}\{\alpha_{ji}b_m b_n^*\}}{\sqrt{2\sigma^2(\alpha_{ii}^2 b_m ^2+\alpha_{jj}^2 b_n ^2)}}\right)$

7.5. Simulation Results

This section provides the simulation results to illustrate the benefits of the newly introduced RIS assisted PSM technique within the BC system framework, highlighting enhancements in BER performance and energy conservation. This confirmation supports the precision of the theoretical formulations.

In Figure 7.2-7.4, BER performance of the proposed system is assessed and these outcomes are compared with the SM-based BC system, extending these comparisons

to PSM bases BC following ZF and MMSE precoding criteria. The analysis highlights the improved performance of our proposed system over those outlined in [115], with all tests conducted on a Rayleigh fading channel where the transmitter has channel information. In these figures, each graph maintains a consistent system setup with identical spectral efficiency. Analytical results are depicted with dashed lines, while solid lines are used for the Monte-Carlo simulation results, validating the accuracy of the theoretical analysis.

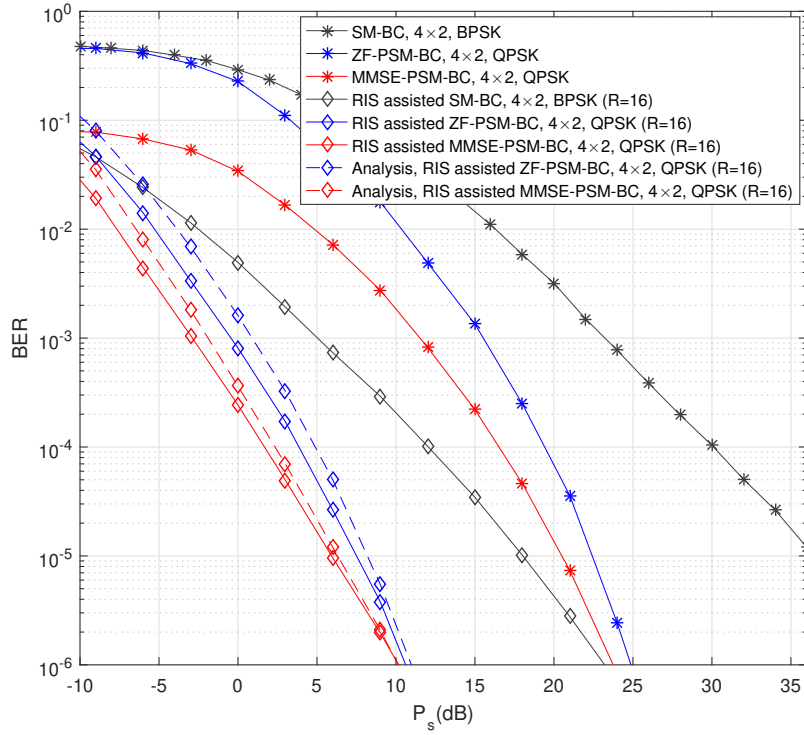


Figure 7.2. BER of RIS assisted PSM-BC system at $R = 3$ bits/s/Hz.

In Figure 7.2, the BER comparison is done for RIS assisted ZF and MMSE-based PSM-BC systems against RIS assisted SM-BC, conventional SM-BC [115], and PSM-BC systems referenced in [158], all operating at a spectral efficiency of 3 bits/s/Hz. The BC configurations analyzed, featuring 4 tag antennas and 2 RN antennas, assisted with $N = 16$ RIS elements, demonstrate significant BER improvements. These improvements are particularly pronounced with MMSE-based precoding. Specifically, at a BER of 1×10^{-4} , the ZF-based RIS aided PSM-BC system shows a 26 dB and 16 dB improvement over the SM-BC and ZF based PSM-BC systems, respectively.

In addition this system performs 8 dB better than RIS assisted SM-BC system at a BER level of 1×10^{-4} . Additionally, at the same BER threshold, the RIS aided MMSE-based PSM-BC system offers an extra 2 dB gain compared to the RIS assisted ZF-PSM-BC system. This indicates that the ZF scheme is prone to noise amplification, which adversely affects its performance. Nevertheless, as the transmit power increases, the performance of the MMSE and ZF schemes tends to converge, as predicted by the theoretical formulations. The comparisons for higher RIS elements are not included in this work due to the space limitations.

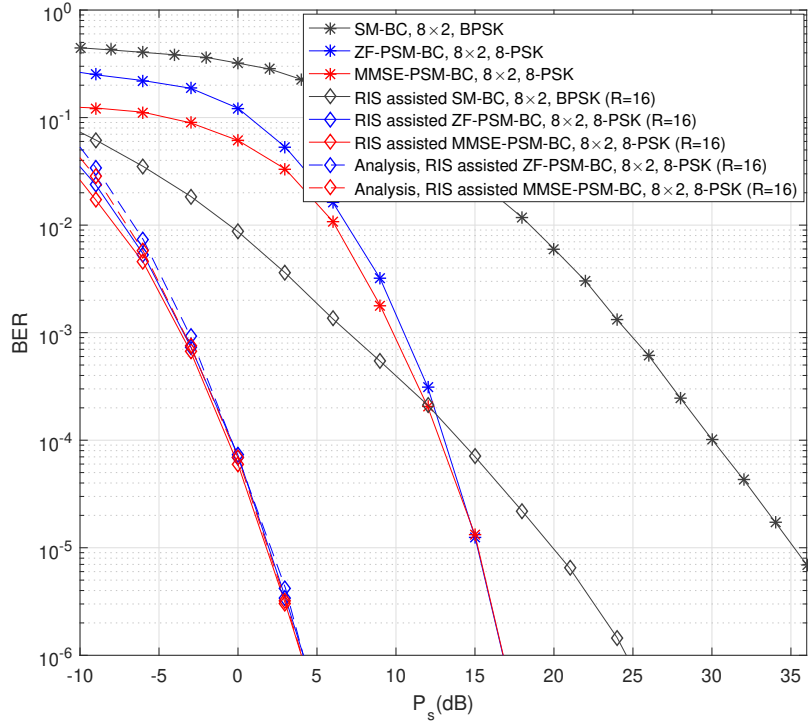


Figure 7.3. BER of RIS assisted PSM-BC system at $R = 4$ bits/s/Hz.

Figure 7.3 provides a performance comparison at a spectral efficiency of 4 bits/s/Hz, focusing on system configuration of 8×2 . The analysis includes the proposed scheme with $N = 16$ RIS elements, PSM-BC, RIS assisted SM-BC and the SM-BC model detailed in [115], all using an equivalent number of tag and RN antennas. The results indicates that the proposed system's performance gains are more substantial with an increased tag count, outperforming SM-BC by 30 dB at a BER of 1×10^{-4} . Additionally, a comparison between the proposed architecture with RIS assisted SM-BC reveals

that the RIS assisted PSM-BC utilizes 15 dB less transmit power to achieve the same BER level. The theoretical upper-bound curves in both scenarios align well with the simulation outcomes, particularly in the moderate power and BER ranges, which are critical for operational assessment.

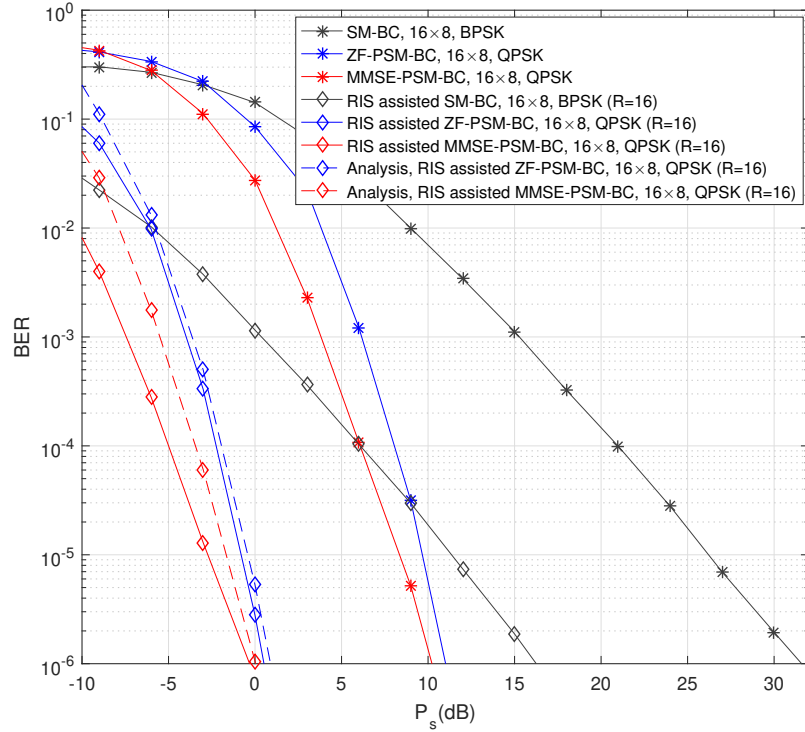


Figure 7.4. BER of RIS assisted PSM-BC system at $R = 5$ bits/s/Hz.

In Figure 7.4, performance is evaluated at a spectral efficiency of 5 bits/s/Hz, revealing notable improvements of approximately 15 dB and 12 dB for RIS assisted MMSE and ZF-based preprocessing with RIS assisted SM-BC system, respectively, at a BER of 1×10^{-5} .

Figure 7.5 illustrates how the number of RIS elements affect the BER performance. To show this effect we assume three different RIS assisted PSM systems with 16, 32 and 64 elements. It can be shown that as the number of RIS elements increases, the performance in terms of BER improves significantly. For example in the case of MMSE preprocessing scheme, when the number of RIS elements is increased from 16 to 64, a 12 dB improvement is obvious at a BER level of 1×10^{-5} . This is the re-

sult of more precise control and manipulation of the communication channel by using more RIS elements, which contribute to improved BER performance in communication systems.

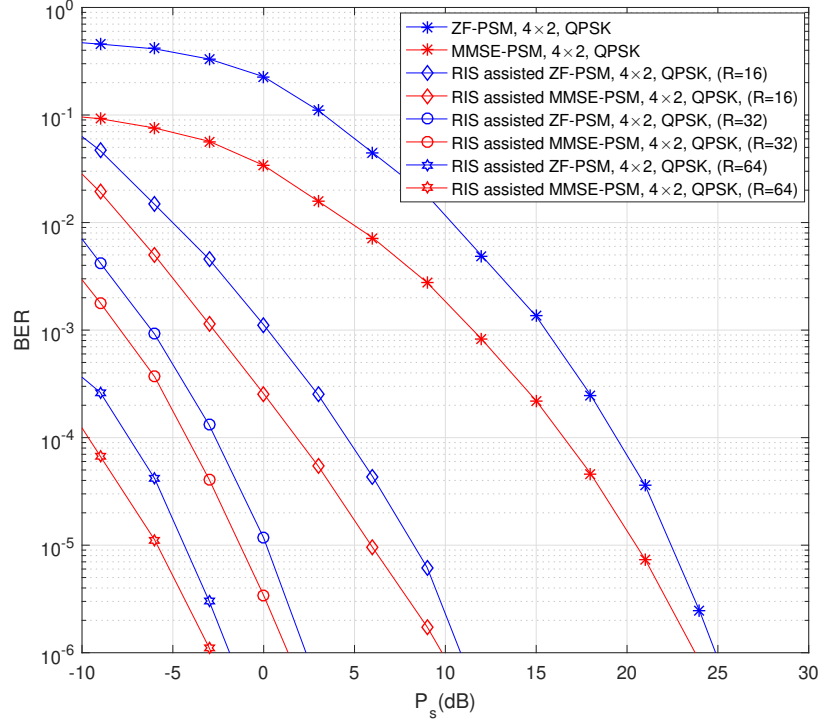


Figure 7.5. BER of RIS assisted PSM-BC systems employing different RIS element numbers at $R = 4$ bits/s/Hz.

Figure 7.6 presents the energy efficiency relative to spectral efficiency for the proposed system at a BER of 1×10^{-4} . The illustration indicates that the RIS assisted preprocessed SM technique used in this work, requires significantly less energy compared to PSM and SM based BC methods. This disparity in energy consumption becomes increasingly apparent with higher spectral efficiencies. For example, to achieve a BER of 1×10^{-4} at a spectral efficiency of 4 bits/s/Hz, the RIS assisted ZF-based BC system utilizes 13.5 dB less power than its conventional counterpart.

The RIS assisted MMSE-based system demonstrates a high reduction in required energy at the spectral efficiency of 3 bits/s/Hz, approximately 14.5 dB compared to MMSE based PSM. These observations are consistent across various configurations,

confirming that the proposed system not only achieves better BER outcomes but also operates more efficiently in terms of energy consumption under consistent noise conditions.

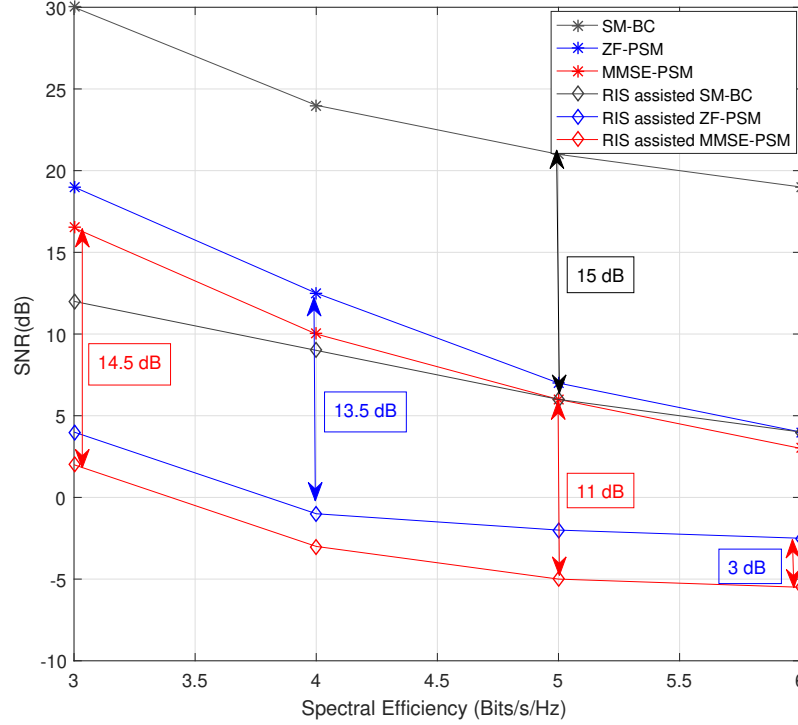


Figure 7.6. Energy efficiency comparison of RIS assisted bistatic BC-SM and BC-PSM systems with conventional BC-SM and PSM-BC to achieve BER level of 1×10^{-4} at different spectral efficiencies.

7.6. Conclusion and Future Work

This work presented a novel RIS aided structure for pre-processed SM based bistatic BC, which proposes the strategic manipulation of RIS reflector phases as a highly reliable method for future wireless communication networks, presenting it as a promising and potential solution. This is done in order to have control over the transmission channel and improve the performance in terms of BER and energy efficiency. Considering the simplicity of implementation and deployment, this technology does not increase the receiver's detection complexity. Extensive Monte-Carlo simulations, confirmed the analytical derivations, which reveals the BER improvement of the

RIS assisted system in comparison to the conventional ones. This design requires the knowledge of the channel phases at the RIS, which is assumed to be fully known in this work. However in the next work we can extend our design to the imperfect channel estimation effects on the performance of RIS aided system. It can be verified by the results that, this architecture significantly offers better BER results in addition to providing higher energy efficiency (considering constant noise levels) than conventional approaches.



8. CONCLUSION

This thesis is composed of 6 main chapters each introducing a novel SM technology for communication systems.

In Chapter 2, the results clearly demonstrate that QSM surpasses SM in performance under uncorrelated Rayleigh channels when maintaining equal spectral efficiency. However, as the channel conditions shift towards a Rician model (with an increasing K-factor), the advantages of QSM diminish, particularly at lower SNR ratios. This suggests that QSM may not be particularly effective in low spectral efficiency contexts, though it proves highly efficient in scenarios requiring higher throughput, especially at high SNRs. Furthermore, it is observed that with increasing channel correlation or deterministic characteristics, it becomes more advantageous to allocate resources to modulation bits rather than spatial bits. Precoding emerges as a significant strategy to improve QSM performance in the presence of channel correlation and Rician characteristics. The primary benefit of the proposed approach is its ability to enhance performance with only a marginal increase in complexity at the transmitter and receiver, without necessitating detailed channel knowledge at the transmitter. The essential requirements include the fixed channel component, the Rician factor, and the correlation among the transmitting antennas.

In Chapter 3, the impact of a precoding strategy on DP-QSM systems across slow-fading, correlated channels is investigated. Optimal rotation values are determined through minimizing the ABEP upper limit at high SNR levels. Both theoretical analysis and comprehensive simulations indicate that this method can significantly enhance performance relative to traditional DP-SM, DP-QSM, and UP-SM systems. Overall, a correlated Rician channel model is introduced, treating uncorrelated and Rayleigh channels as particular instances within this framework using the established Kronecker model. Assuming accurate channel estimation, expressions for PEP, APEP, and ABEP are formulated, and the precoding coefficients that minimize ABEP are identified. Subsequently, using these optimal coefficients, a comparative analysis of error performance is conducted among conventional DP-SM, UP-SM, and DP-QSM

systems versus the newly proposed precoded DP-QSM system under various channel conditions. The benefits of the precoding approach become particularly pronounced in scenarios characterized by high correlation and Rician fading. Hence, integrating the benefits of dual-polarization with precoding methods could facilitate the development of QSM-MIMO systems that demonstrate enhanced resilience to Rician fading and spatial correlations on the transmitter end.

In Chapter 4, the benefits of SSM and dual-polarized transmission are merged to create a resilient SM-MIMO transceiver architecture that withstands the impact of transmit antenna and channel correlations. Theoretical analysis and detailed simulations reveal that this innovative technique may yield substantial performance improvements over traditional DP-SM and comparable UP-SM systems.

In Chapter 5, introduces an innovative precoding strategy designed to mitigate the effects of transmit antenna and channel correlations in multi-stream SM. This method utilizes signature constellations, but diverges from traditional approaches by allocating a distinct group of signature constellations to each combination of active antennas. Additionally, phase rotation and/or amplitude scaling are employed to address potential singularities, even under conditions of complete correlation. Results indicate that this novel technique substantially enhances performance compared to traditional rotation-based precoding, especially in Rician channel environments. Future investigations will seek to expand this technique for MSM configurations with a greater number of transmit antennas.

In Chapter 6, we developed a pre-processing aided SM strategy for bistatic BC systems, specifically tailored for CSIT mode operations with semi-active transmitter tags. We explore the optimization of pre-processing and signal detection within this framework. Proposals for both MMSE and ZF architectures are detailed, along with their theoretical error performance and computational complexity assessments. These theoretical predictions are corroborated by Monte-Carlo simulations under various system configurations. Our findings indicate that the PSM approach provides a promising alternative for signal transmission in MIMO-BC systems, representing an innovative and efficient solution for Beyond-5G IoT technologies. The PSM-BC system is shown

to offer reduced complexity, enhanced BER performance, and greater energy efficiency compared to contemporary benchmark systems. Additionally, the benefits of the proposed system scale with the number of antennas on the tag-side, although incorporating semi-active tags increases hardware complexity and costs at the transmitter. Furthermore, once a precoding matrix is calculated for a given channel matrix, it is stored and utilized throughout the transmission, which necessitates memory elements and gain control circuitry in the backscatter tags. Despite the increased complexity at the tag-side, this approach reduces receiver complexity and improves BER and power efficiency.

We further adapt our tag-side precoding model to account for imperfect channel estimations, evaluating how BER performance is impacted by inaccuracies in channel estimation. Our analysis suggests that PSM technology exhibits greater robustness to channel estimation errors compared to traditional SM-based BC systems. An exciting avenue for future exploration involves applying the PSM technique to Ambient Backscatter Communication (AmBC) systems and exploring extensions to RIS-assisted PSM-based BC systems.

In Chapter 7, an innovative RIS-supported structure for pre-processed SM in bistatic BC systems is proposed, which utilizes the strategic adjustment of RIS reflector phases to reliably enhance future wireless communication networks. This approach aims to control the transmission channel, thereby improving BER and energy efficiency metrics. Due to its straightforward implementation and deployment, this technology does not complicate the receiver's detection processes. Comprehensive Monte Carlo simulations substantiate the analytical findings, demonstrating the superior BER performance of the RIS-assisted system relative to traditional models. This design presumes complete knowledge of the channel phases at the RIS for the current study. Future research could explore the impact of imperfect channel estimation on the efficacy of the RIS-aided system. The results confirm that this architecture not only achieves notably better BER outcomes but also enhances energy efficiency compared to conventional methods, assuming constant noise levels.

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