ISTANBUL TECHNICAL UNIVERSITY ★ GRADUATE SCHOOL OF SCIENCE
ENGINEERING AND TECHNOLOGY

LOW-COMPLEXITY DETECTION AND COOPERATIVE COMMUNICATION FOR SPATIAL MODULATION SYSTEMS

Ph.D. THESIS

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Department of Electronics and Communication Engineering

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İSTANBUL TEKNİK ÜNİVERSİTESİ ★ FEN BİLİMLERİ ENSTİTÜSÜ

UZAYSAL MODÜLASYON SİSTEMLERİ İÇİN DÜŞÜK KARMAŞIKLIKLI SEZİM VE İŞBİRLİKLİ HABERLEŞME

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Gökhan ALTIN, a Ph.D. student of ITU Graduate School of Science Engineering and Technology 504132301 successfully defended the thesis entitled "LOW-COMPLEXITY DETECTION AND COOPERATIVE COMMUNICATION FOR SPATIAL MODULATION SYSTEMS", which he prepared after fulfilling the requirements specified in the associated legislations, before the jury whose signatures are below.

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To my family and my wife

FOREWORD

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ABBREVIATIONS

ABEP	: Average Bit Error Probability
AF	: Amplify-and-Forward
APM	: Amplitude/Phase Modulation
AWGN	: Additive White Gaussian Noise
BEP	: Bit Error Probability
BER	: Bit Error Rate
BPSK	: Binary Phase Shift Keying
CPEP	: Conditional Pairwise Error Probability
CSI	: Channel State Information
D	: Destination
DF	: Decode-and-Forward
EBCS	: Enhanced Bayesian Compressive Sensing
GSM	: Generalized Spatial Modulation
GSIM	: Generalized Spatial Index Modulation
ICI	: Inter-Channel Interference
iid	: Independent, Identically Distributed
IR	: Incremental Relaying
LS	: Least Squares
MGF	: Moment Generation Function
ML	: Maximum Likelihood
MIMO	: Multiple-Input Multiple-Output
MISO	: Multiple-Input Single-Output
MMSE	: Minimum Mean Squared Error
MRC	: Maximal Ratio Combining/Combiner
OB-MMSE	: Ordered Base Minimum Mean Squared Error
pdf	: Probability Distribution Function
PEP	: Pairwise Error Probability
PLNC	: Physical Layer Network Coding
PSK	: Phase Shift Keying
R	: Relay
RF	: Radio Frequency
rv	: Random Variable
QAM	: Quadrature Amplitude Modulation
QPSK	: Quadrature Phase Shift Keying
S	: Source
SD	: Sphere Decoding
SR	: Selective Relaying
SIMO	: Single-Input Multiple-Output
SM	: Spatial Modulation
SNR	: Signal to Noise Ratio
STBC	: Space-Time Block Code

STBC-SM : Space-Time Block Coded Spatial Modulation
SSK : Space Shift Keying
V-BLAST : Vertical Bell Lab Layered Space-Time
ZF : Zero Forcing

SYMBOLS

$\mathbf{x}$	: Transmitted Symbol Vector
$\mathbf{H}$	: Channel Matrix
$\mathbf{y}$	: Received Symbol Vector
$\mathbf{n}$	: Noise Vector
$\mathbf{M}$	: Symbol Constellation Size
$\mathbf{N_t}$, $\mathbf{N_r}$	: Number of Transmit and Receive Antennas
$\mathbf{N_{act}}$	: Number of Active Transmit Antennas
$\mathbf{N_c}$	: Number of Antenna Combination
$\gamma_{\mathbf{SD,SR,RD}}$	: Effective SNR Between Source-Destination, Source-Relay and Relay-Destination
$\mathbf{N_0}$	: Noise Spectral Density
σ^2	: Variance of a Random Variable
$\mathbf{P_{out}}$	: Outage Probability
ρ	: Received SNR
$\mathbf{d_{SD,SR,RD}}$	: Distance Between Source-Destination, Source-Relay and Relay-Destination
α	: Path Loss Exponent
$\mathbf{C}$	: Covariance Matrix
$\mathbf{P_{D,R}(x \rightarrow \hat{x})}$	: Pairwise Error Probability at Destination/Relay
$\lambda_{\mathbf{x}}$	: Eigenvalue of x
$\mathbf{f_X(x)}$	: Probability Density Function of x
$\mathbf{F_X(x)}$	: Cumulative Distribution Function of x
$\mathbf{I(X;Y)}$	: Mutual Information Between X and Y
$\mathscr{CN}(0,\sigma^2)$	: Circularly Symmetric Zero-Mean Complex Gaussian Distribution With Variance σ^2
$\mathbf{I_M}$	: Identity Matrix With Dimensions $M \times M$
$\mathbb{C}^{m\times n}$	: Dimensions Of A Complex-Valued Matrix
$Gamma(\alpha,\beta)$	: Gamma Distribution With Shape And Scale Parameters α and β
$Q(\cdot)$	: Tail Probability of Standard Gaussian Distribution
$\Gamma(\cdot)$	: Gamma Function
$M_\gamma(s)$	: Moment Generating Function
$\Re(c)$	: Real Part of a Complex Variable c

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LOW-COMPLEXITY DETECTION AND COOPERATIVE COMMUNICATION FOR SPATIAL MODULATION SYSTEMS

SUMMARY

Spatial modulation (SM) is an alternative method to classical multiple-input multiple-output (MIMO) spatial multiplexing techniques with vertical-Bell Labs layered space-time (V-BLAST) decoding. In SM, the information is transmitted via the conventional amplitude-phase modulation (APM) symbols along with the active antenna indices. Space shift keying (SSK) is a special case of SM where the information is transmitted through only transmit antenna indices. SSK systems are relatively simple; however, their data rate is lower compared to SM for the same number of transmit antennas. Since the SM symbols are conveyed by only one transmit antenna, one radio frequency chain is used at the transmitter and inter-channel interference (ICI) is removed. On the other hand, transmission rate is limited due to the use of single antenna. Generalized spatial modulation (GSM) mitigates this problem by activating more than one transmit antenna. In GSM, information is both transmitted through the combination of transmit antennas and APM symbols. Hence the transmission rate is increased with the use of multiple active transmit antennas. SM techniques have attracted considerable attention from researchers in the past few years and have been considered as potential candidates for next generation wireless networks.

Since SM scheme exploits the spatial domain for data transmission, its capacity calculation is different than the classical systems. In SM, the information is conveyed not only through the M-ary constellation domain but also through the antenna domain, the capacity of SM is expressed as the sum of the capacities of these two domains. A detailed analysis of the outage probability of SM has not been given in the literature yet. In this thesis, first, we derive the outage probability performance of the classical SM system. It is shown that SM systems provide better performance compared to conventional modulation systems in terms of outage probability.

The optimum performance of SM can be achieved with the maximum likelihood (ML) signal detection but computational complexity increases with exhaustive search. Since the inter-antenna interference arises during the transmission, GSM receiver is much more complicated than the SM scheme. In this work, we investigated a novel low complexity detection algorithm for SM systems (especially for GSM systems due to their receiver complexity). The algorithm can be split into two stages: in the first stage, least squares (LS) estimate of the transmitted signal is used in the ML detection to find the sub-optimal solution of the antenna combinations. The N best estimate of antenna combinations are selected in order to reduce the search space. In the second stage, this reduced set and the LS estimates of transmitted symbols based on this set are then sent to the optimal ML detector to find the correct antenna combinations and APM

symbols. It is shown that the proposed algorithm obtains near ML performance with a lower computational complexity.

Cooperative communication has attracted numerous researchers over the past decade. In a cooperative scenario, a source (S) transmits its data to the relay (R) and the destination (D) in the first phase and the relay forwards the source's information either decoding the received signal (decode-and-forward, DF) or amplifying it (amplify-and-forward, AF) in the second phase. This forwarding concept forms a virtual MIMO system to combat fading and it is very effective to gain a larger coverage.

In this thesis, the advantages of SM and cooperative communication systems have been combined and the outage probability analysis of SM system is extended to the cooperative scenarios under some relaying techniques. Cooperative SM systems are also compared to conventional modulation systems and provide better performance in terms of outage probability.

Finally, we propose novel cooperative SM systems with AF and DF techniques where all nodes have multiple transmit and/or receive antennas, an issue which has not been studied before. The previous studies in the literature for cooperative SM systems generally consider the SSK technique instead of SM. Additionally, most of these studies assume single transmit and/or receive antenna at the relay(s) and destination. As known, the SM/SSK schemes need at least two transmit antennas to map information bits to the antenna indices. Furthermore, a cooperative SM system with DF relaying where the relay(s) has only one transmit antenna is not a complete SM system since the relay(s) can not re-encode the decoded data into the SM symbols. Moreover, to improve the error performance compared to APM, an SM system requires at least two receive antennas. To the best of our knowledge, a comprehensive work on cooperative SM systems that have multiple transmit and/or receive antennas at R and D has not been performed in the past yet. In this work, we derive the average bit error probabilities (ABEP) for both the MIMO-AF and the MIMO-DF systems and validate them with the computer simulations results. Furthermore, these two cooperative SM systems are compared with the classical cooperative systems using M-ary modulations in terms of BER performance. Computer simulations and analytical expressions show that the proposed MIMO-DF and MIMO-AF systems provide considerable error performance improvements over conventional cooperative APM systems. Lastly, the BER comparison of MIMO-AF and MIMO-DF cooperative SM systems are presented.

UZAYSAL MODÜLASYON SİSTEMLERİ İÇİN DÜŞÜK KARMAŞIKLIKLI SEZİM VE İŞBİRLİKLİ HABERLEŞME

ÖZET

Kanal sığasını ve güvenilirliğini artırmada etkili bir rol oynayan, verici ve alıcıda çok sayıda antenin bulunduğu çok-girişli çok-çıkışlı (Multiple-Input Multiple-Output, MIMO) sistemler yeni nesil iletişim sistemleri için en önemli yöntemlerden biridir. Bu nedenle, MIMO sistemler araştırmacılar tarafından üzerinde çokça çalışılmış konulardan olmuştur. Bu çalışmalar sonucunda iki önemli MIMO iletim tekniği geliştirilmiştir. Bunlar, uzay-zaman blok kodlama (space-time block coding, STBC) ve uzaysal çoğullamadır (spatial multiplexing). Uzaysal çoğullamanın en önemli kod çozme uygulaması olan V-BLAST (Vertical-Bell Lab Layered Space-Time) yönteminde tüm antenlerin iletimde olması nedeniyle yüksek derecede kanallar arası girişim (Inter-Channel Interference, ICI) oluşmakta, bu da hata performansı açısından optimum olan en büyük olabilirlikli (Maximum Likelihood, ML) sezim yönteminin performansından uzaklaşılmasına neden olmaktadır. Yine aynı şekilde STBC'nin alıkarmaşıklığıişaret kümesinin boyutu ile üstel şekilde artmaktadır. Son yıllarda bahsedilen bu çoğullama yöntemlere alternatif olarak uzaysal modülasyon (Spatial Modulation, SM) tekniği geliştirilmiştir. SM'in temel prensibi iki boyutlu sinyal kümesine üçüncü bir boyut olarak uzayın (anten indisi) da eklenmesidir. Sinyal sadece klasik genlik/faz modülasyonu (amplitude/phase modulation, APM) ile değil aynı zamanda anten indisi ile de taşınmaktadır. Böylece belirli bir anda vericide sadece tek bir anten aktif olduğundan ICI tamamen ortadan kaldırılmış olmakta, aynı zamanda da yine vericide tek bir radyo frekans katı yeterli olmaktadır. Jeganathan ve arkadaşları tarafından ise SM'in özel bir hali olan uzay kaydırmalı anahtarlama (space shift keying, SSK) yöntemi ortaya atılmıştır. Burada ise klasik APM işaretleri gönderilmemekte, sadece anten indisi ile bilgi taşınmaktadır. Böylelikle sistem karmaşıklığı azaltılmakta ancak aynı anten sayısı için SM'e göre veri hızı düşük kalmaktadır. SM ve SSK'da tek bir antenin aktif olması aynı anten sayısı için diğer çoğullama tekniklerine göre veri hızında düşüşe neden olmaktadır. Veri hızını artırmak amacıyla genelleştirilmiş uzaysal modülasyon (generalized SM, GSM) yöntemi geliştirilmişitir. GSM'de birden fazla anten aktif olmakta ve bilgi hem APM işaretleri ile hem de anten kombinasyonları ile taşınmaktadır. Ancak burada da birden fazla anten aktif olduğu için ICI söz konusu olacaktır.

SM sisteminde bilgi hem anten boyutunda hem de APM boyutunda taşınması nedeniyle kanal sığasının hesaplanması da geleneksel yöntemlerden biraz daha farklı olacaktır. SM'de hem APM boyutu hem uzay boyutu olması nedeniyle, SM sığası bu iki boyutun sığalarının toplamı kadardır. Kesinti olasılığı da o sistemde kanalın veri hızını destekleyip desteklemeyeceği yani kanal sığasıyla ilintili olduğundan, SM'in kesinti olasılığı analizi de yine klasik sistemlerden farklı olacaktır. Kesinti olasılığı hesaplanırken kanalın anlık sığasının bilinmesi gerekmektedir. Bu tezde ilk olarak, daha önce literatürde çalışılmamış olan kanalın anlık sığası verilmiştir.

Bu sığaya bağlı olarak SM'in kesinti olasılığı analizi yapılmış ve kesinti olasılığı açısından SM sisteminin klasik sistemlere göre daha düşük hata başarımına sahip olduğu gösterilmiştir.

SM sistemlerinde en uygun hata performansı ML kod çözücüler ile sağlanmaktadır. ML yönteminde olası tüm APM sembolleri ve anten indisleri taranacağından hesaplama karmaşıklığı özellikle yüksek anten sayıları ve işaret kümeleri için çok fazla olacaktır. Aynı zamanda, GSM'de birden fazla antenin aktif olması nedeniyle antenler arası girişim oluşacağından, alıcı karmaşıklığı daha da artacaktır. Bu çalışmada, ikinci olarak, SM sistemleri için (özellikle GSM sistemi için) düşük karmaşıklığa sahip yeni bir kod çözme algoritması geliştirilmiştir. Algoritma iki evreye ayrılabilir: birinci evrede, anten kombinasyonlarının alt en uygun çözümünü bulmak için ML kod çözmede gönderilen işaretin en küçük kareler (least squares, LS) kestirimi kullanılır. Burada arama uzayını düşürmek için anten kombinasyonlarının hepsi değil sadece N adet en iyi kestirimi seçilir. İkinci evrede ise bu küçültülmüş küme ile bu kümeye bağlı gönderilen işaretin LS kestirimi, doğru anten kombinasyonunu ve APM işaretini bulmak için ML kod çözücüye gönderilir. Benzetim sonuçları göstermiştir ki, geliştirilen bu yeni algoritma hem çok düşük bir karmaşıklığa sahiptir hem de hata olasılığı açısından en uygun çözüm olan ML sonucuna yakın sonuç vermektedir.

Son yıllarda özellikle küçük boyutlu kablosuz cihazlara yönelim giderek artmaktadır. Aynı zamanda yine bir çok cihaz donanım karmaşıklığına sahiptir. Hem boyut, hem donanım karmaşıklığı, söz konusu cihazlara ancak tek bir antenin yerleştirilmesini zorunlu kılmaktadır. Yukarıda bahsedildiği gibi MIMO sistemlerin avantajlarını elde etmek için alıcı ve/veya vericide birden çok antene ihtiyaç duyulmaktadır. İşte bu dezavantajı avantaja çevirmek için çok kullanıcılı (veya röleli) ortamda anten paylaşımı esasına dayanan ve bir nevi sanal MIMO sistem oluşturmayı hedefleyen işbirlikli haberleşme (cooperative communication) ortaya atılmıştır. İşbirlikli haberleşme ile dağıtık anten dizisi oluşturularak bir esneklik kazanılmış ve aynı zamanda uzaysal çeşitlilik elde edilmiştir. Bir işbirlikli senaryoda, birinci aşamada, kaynak (source, S) kendi bilgisini röle (relay, R) ve hedefe (destination, D) göndermektedir. İkinci aşamada ise R aldığı bu bilgiyi çeşitli yöntemlerle D'ye aktarır. R'nin kullandığı yöntem işbirlikli haberleşme sisteminin yapısını ve protokollerini oluşturmaktadır. Genel olarak iki yöntem ön plana çıkmaktadır. Birinci yöntemde, R aldığı bilgiyi sadece kuvvetlendirerek aktarır, kuvvetlendir-ve-aktar (amplify-and-forward, AF). Burada sinyalin gücü artırıldığı gibi gürültüde kuvvetlendirilecektir. İkinci yöntemde ise R aldığı bilgiyi çözüp tekrar kodlayarak D'ye aktarır, çöz-ve-aktar (decode-and-forward, DF). Burada ise R hata yaptığı durumda D'ye bu hatalı işaret gönderilecek ve bir hata yayılımı söz konusu olacaktır. Bu yöntemlere ilave olarak dinamik yöntemlerde kullanılmaktadır. DF yönteminde hata yayılımı nedeniyle D'de bir çeşitleme gelmeyecektir. O halde, DF yönteminde çeşitlemenin artması R'nin bilgiyi doğru çözmesine bağlıdır. Eğer S ve R arası kanal güvenirliği (anlık sinyal gürültü oranı, SNR) yeterli ise R iletime katılmakta aksi halde S kendi bilgisini tekrar göndermektedir. Seçmeli röle (selection relaying, SR) olarak adlandırılan bu yöntem ile röle her zaman iletimde olmamakta, böylelikle röleden kaynaklanacak hata yayılımı azalmaktadır. AF ve DF yöntemlerinde alınan işaret D tarafından doğru çözülecek olsa bile R aldığı bilgiyi iletmektedir. Bu durum veri hızında düşüşe neden olmaktadır. Eğer S ve D arası kanalın güvenirliği yüksek ise R'nin yayın yapmama esasına dayalı olan ve artımlı röle (incremantal relaying, IR) olarak adlandırılan yöntemde ise, S ve D arası SNR'ın belirli bir eşik değerinden yüksek olması durumunda R iletime katılmamakta, eşik seviyesinden düşük ise R iletime katılmaktadır. Böylelikle, ikinci evre har zaman olmadığından veri hızı artırılmış olmaktadır.

Bu tezde, hem SM'in hem de işbirlikli haberleşmenin avantajları birleştirilerek yeni bir sistem yapısı ortaya konulmuş ve önceki bölümlerde verilmiş olan SM'in kesinti olasılığı analizi bir aşama daha ileriye taşınarak işbirlikli SM sisteminin kesinti olasılığı analizi yapılmıştır. Sabit AF ve DF yöntemlerine ek olarak literatürdeki diğer yöntemler olan DF-SR, DF-IR ile AF-IR yöntemlerinin de kesinti olasılığı analizleri yapılmıştır. Yine burada da kesinti olasılığı açısından işbirlikli SM sistemi klasik işbirlikli sistemlerle karşılaştırılmış ve daha düşük kesinti olasılıkları elde edilmiştir.

Ayrıca, SM'in ve işbirlikli haberleşmenin avantajlarının kullanıldığı, daha önce literatürde olmayan tüm noktalarda (S, R ve D) çoklu antenlerin kullanıldığı yeni bir işbirlikli SM sistemi hem AF hem de DF tekniği açısından incelenmiştir. Literatürde bulunan önceki çalışmalar genel olarak SM'i değil SSK'yi kullanmış aynı zamanda bu çalışmaların çoğu R ve D'de tek alıcı ve/veya verici anten olduğu varsayımını yapmıştır. Bilindiği gibi, SM/SSK bilgi bitlerini anten indisine eşlemek için en az iki antene ihtiyaç duyar. Ayrıca, R'de tek bir anteni bulunan DF işbirlikli SM sistemi, S'den aldığı bilgiyi R'nin çözüp tekrar SM sembolüne kodlayamayacak olmasından dolayı eksik bir işbirlikli SM sistemi olacaktır. Bunun yanında, SM'in klasik yöntemlere göre hata başarımının daha iyi olması için alıcıda en az iki antenin olması gerekmektedir. Bilindiği kadarıyla, literatürde hem R hem de D'de çok antenin bulunduğu kapsamlı bir işbirlikli SM çalışması yoktur. Bu çalışmada, MIMO-DF SM sistemi ile gönderilen işaret R tarafından çok sayıdaki alıcı anten ile alınarak çözülür. İkinci evrede R çözülen işareti tekrar SM işaretine eşler ve D'ye gönderir. D ise çok sayıdaki alıcı anteni ile hem S'den gelen hem de R'den gelen işaretleri alarak ML kod çözme uygular. MIMO-AF SM sistemi ile de birinci evrede S'den gönderilen SM işareti R'de tüm antenlerden alınır, kuvvetlendirilir ve yine tüm antenlerden D'ye akatarılır. D'de yine S ve R'den gelen işaretler için ML kod çözme uygulanarak karar verilir. Çalışmada hem MIMO-AF hem de MIMO-DF sistemleri için ortalama bit hata olasılığı çıkarılmış ve bilgisayar benzetimleri ile doğruluğu sınanmıştır. Ayrıca, bu iki yöntem M'li modülasyon kullanan klasik işbirlikli haberleşme sistemleri ile de bit hata olasılığı açısından karşılaştırılmıştır. Bilgisayar benzetimleri göstermiştir ki, çıkarılan analitik bit hata olasılığı ifadeleri ile bilgisayar benzetim sonuçları örtüşmektedir. Aynı zamanda, işbirlikli SM sistemleri klasik işbirlikli sistemlere göre dikkate değer ölçüde hata performansında iyileştirme sağlamıştır. Son olarak, MIMO-AF ile MIMO-DF SM sistemlerinin karşılaştırmaları verilmiştir.

1. INTRODUCTION

The demand for higher data rates and improved error performance lead the researchers to seek new and efficient transmission techniques. Multiple-input multiple-output (MIMO) transmission technologies cover this demand in the sense of increased channel capacity and/or improved error performance [1]. Two main MIMO strategies are Space Time Block Coding (STBC) and spatial multiplexing. The former improves the reliability by exploiting transmit diversity while distributing conventional two dimensional signal constellation into space and time dimensions [2]. However, for more than two transmit antennas, the symbol rate of STBC is upper bounded by $3/4$ symbols per channel use when it has full diversity. The latter provides high spectral efficiency with powerful decoding methods such as Vertical-Bell Lab Layered Space Time (V-BLAST) [3] which satisfy the high data rate demand. In V-BLAST, the multiple data streams are transmitted over multiple antennas to increase the capacity. As a result of simultaneous data transmission over all antennas, a high level Inter Channel Interference (ICI) occurs.

Spatial modulation (SM) [4] is an alternative method to spatial multiplexing methods mentioned above. In SM, the information is transmitted via the conventional amplitude-phase modulation (APM) symbols along with the active antenna indices. $\log_2(N_t)$ bits are mapped to the index of activated transmit antenna and $\log_2(M)$ bits are allocated for the APM, where N_t is the number of transmit antennas and M is the constellation size. Hence, the total number of transmitted bits becomes $\log_2(N_tM)$ for the SM scheme. Since only one transmit antenna is active during the transmission, one radio frequency (RF) chain is sufficient at the transmitter of SM and additionally, ICI is eliminated.

After SM was introduced, special cases of SM have been investigated by researchers. Space shift keying (SSK) is one of the special cases of SM where the information is transmitted through only transmit antenna indices [5]. SSK systems are relatively simple; however, their data rate is lower compared to SM for the same number of transmit antennas.

Since transmission rate is limited due to the use of single antenna for both SM and SSK, generalized SM (GSM) [6], [7] is introduced as another special case. GSM mitigates this problem (low data rate) by activating more than one transmit antenna. In GSM, information is both transmitted with the combination of transmit antennas and APM symbols. Hence the transmission rate is increased with the use of multiple active transmit antennas. This advantage comes at a price. The GSM receiver is much more complicated than the SM scheme since the inter-antenna interference arises during the transmission. The optimum performance can be achieved with the maximum likelihood (ML) signal detection but computational complexity increases with exhaustive search.

SM techniques have attracted considerable attention from researchers in the past few years and have been considered as potential candidates for next generation wireless networks [8].

Cooperative communications which forms a virtual MIMO system to combat fading and which is very effective to gain a larger coverage, has been investigated extensively during the recent times. In a cooperative scenario, a source (S) transmits its data to the relay (R) and to the destination (D) in the first phase, and the relay forwards the source's information either by first decoding the received signal (decode-and-forward, DF) or amplifying it (amplify-and-forward, AF) in the second phase. Besides the relay processing techniques, different kind of diversity protocols can be employed for a cooperative scenario [9]. In selection relaying (SR), S sends its information to R and D in the first phase. In the second phase, if the instantaneous SNR between S and R falls below a certain threshold, S continues transmitting to D and R does not transmit (R remains silent); if it is above the threshold, R forwards what it received using DF technique. The spectral efficiency for fixed AF/DF relaying and SR is halved since the transmission takes two time slots. This is improved by incremental relaying (IR). In IR, if the S-D link SNR is not high enough for appropriate communication, the relay participates the communication and transmits the received signal using DF or AF techniques. Otherwise, R remains silent and there is no second phase.

Bit error probability (BEP) is a key performance measure for a communication system which transmits digital information from one point to another. Additionally, when the signal power drops below a certain threshold (i.e., an outage event occurs), it is highly likely to have decoding failure. In that sense, the outage probability is another important performance measure for communication systems operating over fading channels.

In this thesis, we, first, derive the outage probability performance of a classical SM system. Secondly, we investigate a novel low complexity detection algorithm for SM systems (especially for GSM systems due to their receiver complexity). Moreover, we combine the advantages of SM and cooperative communication systems and propose novel cooperative SM systems with AF and DF techniques where all nodes have multiple transmit and/or receive antennas, an issue which has not been studied before. Therefore, we extend the outage probability analysis of SM system to the cooperative scenario and finally, we derive the average BEP (ABEP) for both the MIMO-AF and the MIMO-DF systems.

1.1 Purpose of Thesis

In this thesis, first, the outage probability analysis of an SM system is investigated. As known, a detailed analysis of the outage probability of SM has not been given in the literature yet. Since the information is conveyed by both antenna and APM domains, the capacity of an SM system will be the sum of the capacities of these two domains. Hence, the outage probability analysis of the SM scheme will not be the same as the conventional systems. Moreover, SM capacity averages the mutual information of antenna and APM domain from all transmit antennas. However, since only one antenna is active for a given time interval, the outage event occurs over the activated link. Hence, the instantaneous capacity is, first, investigated. After defining the instantaneous capacity, the outage probability of SM is given. It is shown that SM systems provide better performance compared to conventional modulation systems in terms of outage probability.

GSM appears as two different architecture in the literature. In one of them, different APM symbols are send from the activated transmit antenna combinations which is also called Generalized Spatial Index Modulation (GSIM) [10]. In the other one, which is the special case of the first, same symbol is transmitted from the activated antennas[1]. In this study, secondly, we propose new and simple detection algorithms which reduce the computational complexity of GSM. The proposed algorithms can be split into two stages: in the first stage, least squares (LS) estimate of the transmitted signal is used in the ML detection to find the sub-optimal solution of the antenna combinations. The N best estimate of antenna combinations are selected in order to reduce the search space. Corresponding APM symbols are then solved with the quantization (slicing) operation after the detection of the active antennas. In the second stage, this reduced set and the LS estimates of transmitted symbols based on this set are then sent to the optimal ML detector to find the correct antenna combinations and APM symbols. We introduce three different algorithm since the GSM and GSIM are slightly different. In the first algorithm, APM symbols are solved with the quantization (slicing) operation after the detection of the active antennas. For GSM, quantized APM symbols are the optimum solution for the optimum antenna combination. On the other hand, for GSIM, quantization operation is not sufficient to reach closer to the ML solution especially for high activated transmit antenna numbers. A simple way to approach the ML performance is to use all possible symbol combinations after defining the N best estimate of antenna combinations. This is the second algorithm. However, this approach brings extra complexity. For further complexity reduction, APM symbols set is reduced using sphere decoding (SD) in Algorithm 3. It is shown that the proposed algorithms obtain near ML performance with a lower computational complexity.

Afterwards, the advantages of SM and cooperative communication systems have been combined. This subject that constitutes the main work of the thesis examined in two topics. In one of them, we extend the outage probability analysis of SM system to the cooperative scenarios. In the other one, ABEP is analyzed for MIMO-DF and MIMO-AF SM systems where all nodes have multiple transmit and/or receive antennas. As is known, these two topics also has not been studied before.

The instantaneous capacity and outage probability of SM are extended to a cooperative system under fixed, selective and incremental relaying techniques. The outage probability analysis for classical modulation techniques and different relay and diversity protocols can be found in [9]. In this study, we have used the results for these

[1]During the thesis, we call the first case as GSIM and the second case as GSM

protocols and applied them to the cooperative SM systems. It is shown that, the outage probability of cooperative SM scheme provides better performance than conventional modulation techniques.

The previous studies in the literature for cooperative SM systems generally consider the SSK technique instead of SM. Additionally, most of these studies assume single transmit and/or receive antenna at the relay(s) and destination. As known, the SM/SSK schemes need at least two transmit antennas to map information bits to the antenna indices. Furthermore, a cooperative SM system with DF relaying where the relay(s) has only one transmit antenna is not a complete SM system since the relay(s) can not re-encode the decoded data into the SM symbols. Moreover, to improve the error performance compared to APM, an SM system requires at least two receive antennas. To the best of our knowledge, a comprehensive work on cooperative SM systems that have multiple transmit and/or receive antennas at R and D has not been performed in the literature yet.

In this thesis, we, lastly, consider a cooperative SM scheme where S, R and D have multiple transmit/receive antennas. In this scheme, S sends the SM symbols to R and D in the first phase. In the second phase, R processes and forwards the received signal either by amplifying or decoding. When it uses the AF strategy, it only amplifies the received signal at each antenna and sends it from all antennas (MIMO-AF). When DF strategy is utilized, R decodes the received signal using maximum likelihood (ML) detection and maps the estimated signal to a new SM symbol and forwards it to D in the second phase (MIMO-DF). At D, ML detection is employed to determine the transmitted SM symbol. In this study, we derive the average bit error probabilities (ABEP) for both the MIMO-AF and the MIMO-DF systems and validate them with the computer simulations results. Furthermore, these two cooperative SM systems are compared with the classical cooperative systems [9] using M-ary modulations in terms of BER performance. Computer simulations and analytical expressions show that the proposed MIMO-DF and MIMO-AF systems provide considerable error performance improvements over conventional cooperative APM systems. Finally, the BER comparison of MIMO-AF and MIMO-DF cooperative SM systems are presented.

1.2 Thesis Organization

The rest of the thesis is organized as follows. Chapter 2 provides the background for SM and cooperative systems. Also in this chapter, a comprehensive literature survey is given. It describes the prior works for SM capacity, low complexity detection of GSM and both ABEP and outage probability analysis of cooperative SM systems. In Chapter 3, the outage probability analysis of SM is presented. Chapter 4 investigates low-complexity detection algorithms for GSM systems. The outage probability of cooperative SM systems is discussed in Chapter 5. In Chapter 6, the ABEP analysis of cooperative SM system with DF and AF relaying are given, respectively. Finally, Chapter 7 concludes this research.

Notation: A scalar, a vector and a matrix will respectively be denoted by a lower/upper-case italic, a lower-case boldface and an upper-case boldface letter. $(\cdot)^T$, $(\cdot)^H$ and $\|\cdot\|$ represent transpose, Hermitian transpose and Euclidean/Frobenius norm of a vector/matrix, respectively. $\mathbb{C}^{m\times n}$ represents the dimensions of a complex-valued matrix. $\Pr\{\cdot\}$ denotes the probability of an event and $E\{\cdot\}$ is the expectation operation. $\Re(c)$ represents the real part of a complex variable c. The probability density function (pdf) and the cumulative distribution function (cdf) of a random variable (rv) X are given as $f_X(x)$ and $F_X(x)$, respectively. $I(X;Y)$ represents the mutual information between X and Y and $H(\cdot)$ is used for the entropy function. $\mathcal{CN}(0,\sigma^2)$ denotes the circularly symmetric zero-mean complex Gaussian distribution with variance σ^2. $\mathbf{I}_M$ is the identity matrix with dimensions $M\times M$. $Gamma(\alpha,\beta)$ denotes the Gamma distribution with shape and scale parameters α and β, respectively. $Q(\cdot)$ is the tail probability of standard Gaussian distribution, $\binom{\cdot}{\cdot}$ denotes the binomial coefficient and $\lfloor . \rfloor$ is the floor operation. $\Gamma(\cdot)$ is the gamma function. $M_\gamma(s) = E\{e^{s\gamma}\}$ is the moment generating function (MGF) of a rv γ and $\mathrm{tr}(\cdot)$ is the trace operator.

2. THEORETICAL BACKGROUND

In this chapter, a brief explanation for a MIMO channel model which is assumed throughout the thesis is given. An alternative method to classical MIMO transmission techniques, SM, is also introduced. System model for SM as well as special cases of SM, SSK and GSM systems are presented. Furthermore, general concept and relaying techniques for cooperative communications are also given in this chapter.

2.1 Multiple-Input Multiple-Output (MIMO) Channels

Since a wireless channel is relatively dynamic and unpredictable, the analysis of a wireless communication system is much more complicated compared to wire-line channels. In the wireless channel, radio transmission experiences three physical effects: reflection, diffraction and scattering [11], [12]. The attenuated, delayed and shifted in phase and/or frequency versions of transmitted signal, which are also called multipath components, may occur in the receiver due to these three phenomena. When the multipath components and the transmitted signal are summed at the receiver, they produce a distortion in the received signal.

Transmitted signal power can be affected in two different ways. First one is the large-scale effect where the signal power changes over long distances. This is also called attenuation, path loss or large-scale fading [12], [13]. Second one is the small-scale fading (or just fading) where the signal power changes over short distances and/or time interval [12], [13]. An example of the relationship between large-scale and small-scale fading can be seen in Figure 2.1.

A fading channel can be characterized by the multipath delay causing frequency selective or frequency flat channels and by time variation due to mobile speed (Doppler spread) causing fast fading or slow fading. The classification of fading channels can be seen in Figure 2.2.

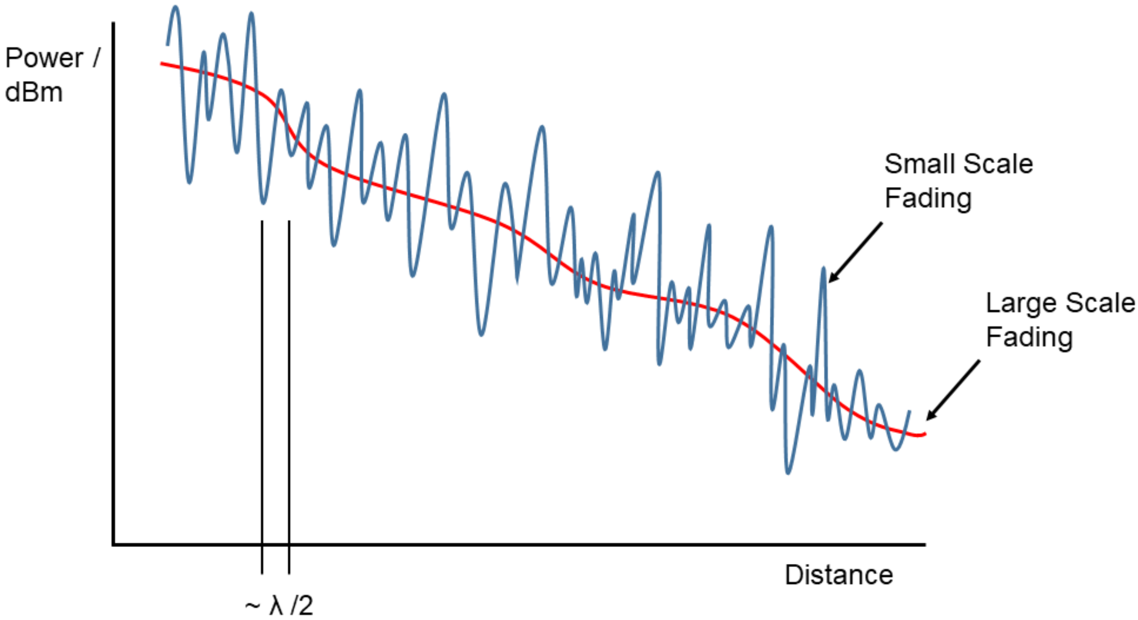

Figure 2.1 : Large-scale vs. small-scale fading.

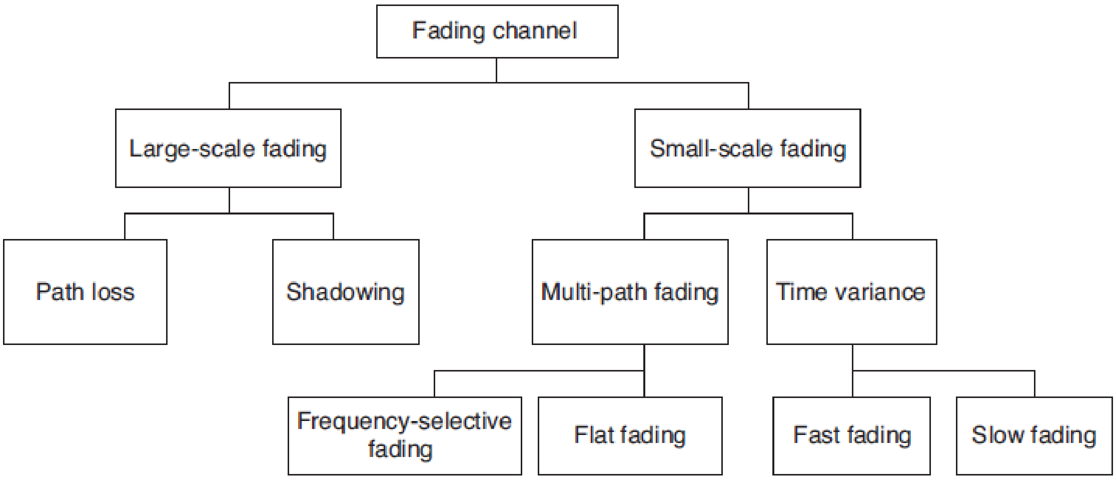

Figure 2.2 : Classification of a fading channel [1].

Let $s(t)$ be a baseband signal at the transmitter. The corresponding passband signal will be

$$x(t) = \Re\left[s(t)e^{j2\pi f_c t}\right]. \tag{2.1}$$

If the passband signal in (2.1) passes through a wireless channel which has K different multiple paths with different Doppler shifts, the received signal will be

$$y(t) = \Re\left[\sum_{k=1}^{K} a_k e^{j2\pi(f_c+f_k)(t-\tau_k)} x(t-\tau_k)\right] \tag{2.2}$$

where a_k, f_k and τ_k represent the amplitude, Doppler shift and delay of the kth path, respectively. (2.2) can be written as

$$y(t) = \Re\left[r(t)e^{j2\pi f_c t}\right] \tag{2.3}$$

where $r(t) = \sum_{k=1}^{K} a_k e^{-j2\pi(f_c+f_k)\tau_k + f_k t} x(t-\tau_k)$ is the received baseband signal. (2.3) denotes the linear-time varying nature of the communication channel. If we assume that the channel bandwidth is wider than the signal bandwidth, τ_k can be approximated as a constant and the received passband signal can be re-written as

$$\begin{aligned} y(t) &= \Re\left[\sum_{k=1}^{K} a_k e^{-j\phi_k} x(t-\tau)\right] \\ &= r_I(t)\cos 2\pi f_c t - r_Q(t)\sin 2\pi f_c t \end{aligned} \tag{2.4}$$

where $r_I(t)$ and $r_Q(t)$ are the in-phase and quadrature components of $r(t)$, $\phi_k(t) = 2\pi(f_c + f_k)\tau - f_k t$ and we assume $x(t-\tau) = 1$. When the multiple paths are large enough, i.e., $K \gg 1$, in-phase, $r_I(t) = \sum_{k=1}^{K} a_k \cos\phi_k(t)$, and quadrature, $r_Q(t) = \sum_{k=1}^{K} a_k \sin\phi_k(t)$, components approximated as independent, identically distributed (iid) Gaussian random variables (rv) by the central limit theorem. Since $r_I(t)$ and $r_Q(t)$ are iid Gaussian rv.s, the amplitude of the received signal, $|y(t)| = \sqrt{r_I^2(t) + r_Q^2(t)}$ follows the Rayleigh distribution. The probability density function (pdf) of the Rayleigh distribution is

$$f_X(x) = \frac{x^2 e^{\left(\frac{-x^2}{2\sigma^2}\right)}}{\sigma^2}, \qquad x \geq 0 \tag{2.5}$$

where σ^2 is the variance of $r_I(t)$ and $r_Q(t)$.

If some of the multipath components are much stronger than others, i.e., if there is a line of sight (LOS) component, the amplitude of the received signal is no longer Rayleigh and it follows Rician distribution. The pdf of Rician distribution can be given as

$$f_X(x) = \frac{z e^{\left(-\frac{z^2+s^2}{2\sigma^2}\right)}}{\sigma^2} I_0\left(\frac{xs}{\sigma^2}\right), \qquad x \geq 0 \tag{2.6}$$

where $s \geq 0$ is the noncentrality parameter and $I_0(.)$ is the zeroth order modified Bessel function of the first kind.

At the receiver, the received signal in (2.4) is demodulated and sent to the matched filter to obtain the baseband signal. The matched filter output can be given as

$$r_t = h s_t + n_t \tag{2.7}$$

where h is the zero mean, complex Gaussian rv, s_t and n_t are the discrete time representation of transmitted signal and the additive noise, respectively. (2.7) is called fading channel model, where h represents channel gain and its amplitude follows Rayleigh distribution and n_t is the additive Gaussian noise.

The fading channel in (2.7) causes very significant decreases in the received signal power. Since the noise power is usually constant, the instantaneous signal to noise ratio (SNR) can be exposed to deep fading. It is essential that SNR have to be above a threshold in order to determine the received signal. If we have more copy of the transmitted signal passed through the fading channel, they are affected from fading separately and the probability of excessive fading of each will decrease. This phenomenon has brought about the concept of diversity which increases the reliability of the communication channel.

Diversity can be reached in different ways such as frequency, time and space (antenna) diversities [11], [12]. Space diversity can be implemented with two or more antennas at the transmitter and/or receiver. This multiple antenna concept is called single-input multiple-output (SIMO) for the systems that has single antenna at the transmitter and multiple antennas at the receiver, multiple-input single-output (MISO) for multi-antenna transmitter and single antenna receiver systems and more generally multiple-input multiple-output (MIMO) for both multiple antennas at the transmitter and the receiver. In this thesis, one of the transmission techniques which is called spatial modulation (SM) based on the space diversity (MIMO systems) are studied.

2.2 Spatial Modulation (SM)

SM is a new approach for MIMO systems and an interesting alternative to classical multiplexing techniques. In the last decade, it has been seen from the literature that the concept of SM has been handled by some researchers with different names and implementations. First work in this area is [14]. In here, it is thought that the differences in the signals from the different transmit antennas can be used to transmit information and a novel system is proposed which is called space shift keying (SSK). In this system, as opposed to classical SM, the signal is transmitted from all antennas. In [15], the idea of carrying the information over the antenna indices is proposed. However, in this data multiplexing scheme, information bits are multiplexed in an orthogonal fashion at frequency domain rather than the space domain. Classical SM can be seen first in [16] with a name, *channel hopping*. In the literature, Mesleh, *et.al.* is standardized this idea and used the term SM [4].

2.2.1 Working principles of SM

Let us consider a MIMO system where the number of transmit antennas is N_t and the number of receive antennas is N_r. MIMO channel matrix composed of channel fading coefficients can be given as $\mathbf{H} \in \mathbb{C}^{N_r \times N_t}$. Each element of the above matrix is modeled as an iid complex Gaussian rv with $\mathscr{CN}(0, \sigma_h^2)$ and the channel obeys the Rayleigh flat fading model. An SM symbol carries $\log_2(N_t M)$ information bits where N_t is the number of transmit antennas and M is the constellation size for the conventional APM. The first $\log_2(N_t)$ bits are mapped into the transmit antenna index while the remaining bits are mapped into APM. A unit energy, $E[\mathbf{x}^H \mathbf{x}] = 1$, SM symbol can be regarded as $\mathbf{x} = [\,\underbrace{0, \ 0, \ \ldots, \ 0,}_{l-1} \ x_q, \ \underbrace{0, \ \ldots, \ 0}_{N_t - l}\,]^T$, where l is the active antenna index and x_q is the APM constellation symbol. The received signal is given as

$$\mathbf{y} = \mathbf{h}_l x_q + \mathbf{n} \tag{2.8}$$

where $\mathbf{h}_l \in \mathbb{C}^{N_r} \times 1$ vector is the l^{th} column of the MIMO channel matrix, $\mathbf{H} = \begin{bmatrix} \mathbf{h}_1 & \mathbf{h}_2 & \ldots & \mathbf{h}_{N_t} \end{bmatrix}$ and $\mathbf{n}$ is the $N_r \times 1$ additive white Gaussian noise (AWGN) vector whose entries are modeled as $\mathscr{CN}(0, N_0)$ with noise spectral density $N_0/2$ per dimension.

In [4], the detection rule for SM based on the maximum ratio combining (MRC) is suggested as

$$z_l = \left|\mathbf{h}_l^H \mathbf{y}\right| \qquad l = 1, \ldots, N_t$$

$$\hat{l} = \arg\max_l z_l \tag{2.9}$$

$$\hat{x}_q = Q\left\{z_{l=\hat{l}}\right\} \tag{2.10}$$

which separately decides the antenna index and transmitted signal. In 2.10, $Q\{.\}$ is the quantization (slicing) operator. This rule is modified in [17] as $z_l = \frac{|\mathbf{h}_l \mathbf{y}|}{\|\mathbf{h}_l\|_F^2}$ and it is a sub-optimal detection. In this work the joint detection, i.e., ML detection rule, is also introduced. The detector having ideal channel state information (CSI) estimates antenna index, l, and the APM signal, x_q, with the ML decision rule as follows [17],

[18]:

$$[\hat{l}, \hat{x}_q] = \arg\min_{l,q} \left\| \mathbf{y} - \mathbf{h}_l x_q \right\|^2 . \tag{2.11}$$

2.2.2 Advantages and disadvantages of SM

The advantages of SM technique over conventional MIMO transmission schemes can be summarized as follows [19]:

1. *Higher throughput.* Since the additional information bits are transmitted by antenna indices in the SM system, the bandwidth efficiency of the SM increases logarithmically with the increasing number of transmit antennas.

2. *Simpler receiver design.* ICI in SM has been completely eliminated. Therefore, the receiver of this system is simpler than that of MIMO systems such as V-BLAST, since it transmits single symbol, thus it does not need complex interference cancellation algorithms.

3. *Simpler transmitter design.* Since only one transmit antenna is used at a symbol period, in the transmitter, a single active RF chain is sufficient to implement SM.

4. *Lower transmit power.* Since the single RF chain is used to achieve the multiplexing gain, overall consumed power is reduced.

5. *No antenna synchronization.* There is no need for antenna synchronization in SM.

6. *No limit for number of receive antennas.* Unlike V-BLAST, there is no lower limit for the number of receiving antennas for the SM system. In other words, SM can work smoothly for $N_r < N_t$.

7. *More complexity reduction.* It is possible to further reduce the complexity of the SM structure with the SSK which will be described in the next section.

Some disadvantages of SM technique compared to conventional MIMO transmission systems are as follows [19]:

1. *Lower throughput.* Compared with V-BLAST, the bandwidth efficiency with increasing number of antennas does not increase linearly. Therefore, the number of antennas for SM system that will be needed to achieve high bandwidth efficiencies

achieved with the V-BLAST system can go well beyond the acceptable limits. If the number of antennas is kept at the acceptable limits, the APM constellation size must be increased which degrades the symbol error performance.

2. *Fast antenna switching.* Since a single antenna is active and a single RF chain is sufficient for the SM system, fast RF switching is needed.

3. *Independent channels.* If the channels between the transmitter and the receiver are not independent, i.e., correlated, the error performance of the SM will degrade.

4. *Perfect channel knowledge.* Receiver needs perfect channel knowledge for detection.[1]

2.3 Special Cases of SM

As SM gained considerable popularity, researchers have sought to explore alternative ways to improve the SM. Two main special cases of SM are Space Shif Keying and Generalized SM.

2.3.1 Space shift keying (SSK)

SSK, which is completely different from [14], activates only one transmit antenna and uses this activated antenna index to convey information by using simple carrier signal rather than the classical PSK/QAM modulated signals [5]. Thus, the bandwidth efficiency of the SSK system is $\log_2(N_t)$ bits/sec/Hz, which is lower than $\log_2(N_t M)$ bits/sec/Hz which is the bandwidth efficiency of the SM system. This elimination of APM provides some advantages over SM. This advantages can be written as,

- Decoding complexity of SSK is lower than SM and its performance is very close to SM's performance.
- Since the amplitudes and phases of the pulses do not convey information, the transmitter-receiver structure of the SSK is even simpler and therefore, non-coherent receivers can be used.

[1] In [20], it is specified that SM requires perfect channel information, otherwise its performance could be seriously worsened. However, some recent studies in the literature, as well as analyzes made in [21], proved that this is not true.

- Because the structure of SSK is simple, it can be simply integrated into technologies such as ultra wide band (UWB).

The SSK received symbol can be written as

$$\mathbf{y} = \mathbf{h}_l + \mathbf{n} \tag{2.12}$$

where l is same as SM and $x_q = 1$ for all q. The ML detection rule for SSK is

$$\hat{l} = \arg\min_{l} \|\mathbf{y} - \mathbf{h}_l\|^2 . \tag{2.13}$$

In [5], the performance of the SSK system has been examined in detail and it has been shown that the performance of the SSK worsens with increasing number of transmit antennas. On the other hand, SSK's error performance improves with increasing number of receive antennas. In [22], the SSK system has been enhanced to a structure called generalized SSK (GSSK) which activates multiple transmit antennas. In this study, an optimization was performed in the space domain to minimize the error probability. However, the number of RF chains used in this method is higher than that of SM and SSK systems.

2.3.2 Generalized spatial modulation (GSM)

Since the SM symbols are conveyed by only one transmit antenna, transmission rate is limited due to the use of single antenna.

GSM mitigates this problem by activating more than one transmit antenna [6], [7]. In GSM, information is both transmitted with the combination of transmit antennas and APM symbols. Hence the transmission rate is increased with the use of multiple active transmit antennas.

This advantage comes at a price. The GSM receiver is much more complicated than the SM scheme since the inter-antenna interference arises during the transmission.

GSM appears as two different architecture in the literature. In one of them, different APM symbols are send from the activated transmit antenna combinations which is also called Generalized Spatial Index Modulation (GSIM) [10]. In the other one, which is the special case of the first, the same symbol is transmitted from all activated antennas [6], [7].

In GSM/GSIM, $m_a = \lfloor \log_2 \binom{N_t}{N_{act}} \rfloor$ bits are mapped to the index of antenna combinations where N_{act} is the number of activated transmit antennas. Additionally, $m_s = N_{act} \log_2(M)$ ($m_s = \log_2(M)$ for GSM) bits are mapped to the APM symbols, s_l, $l = 1, \ldots, M$ where M is the constellation size, and hence totally $m = m_a + m_s$ bits are allocated for a GSM/GSIM symbol. Since the antenna combinations must be an order of two, any $N_c = 2^{m_a}$ combinations can be used for the transmission. For instance, there are $\binom{5}{2} = 10$ antenna combinations for $N_t = 5$ and $N_{act} = 2$, but only $N_c = 8$ antenna combinations can be used (any 8 out of 10 can be selected). Let the total information bits be $m = 7$. First three bits are mapped to antenna combinations, i.e., first and third antennas, and remaining four bits are mapped to two 4-PSK/QAM symbols to construct the GSIM transmitted symbol as

$$\mathbf{x}_{GSIM} = \begin{bmatrix} s_1 & 0 & s_2 & 0 & 0 \end{bmatrix}^T. \tag{2.14}$$

When the GSM symbol is considered, total number of bits will be $m = 5$ and only one APM symbol takes place in the transmitted symbol as

$$\mathbf{x}_{GSM} = \begin{bmatrix} s & 0 & s & 0 & 0 \end{bmatrix}^T. \tag{2.15}$$

The received signal for both systems can be written as

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{n} \tag{2.16}$$

where $\mathbf{n} \in \mathbb{C}^{N_r \times 1}$ is the AWGN vector whose entries are modeled as $\mathscr{CN}(0, N_0)$ with noise spectral density $N_0/2$ per dimension.

The detector at the receiver, which has the ideal channel state information, detects the GSM/GSIM symbols applying the ML decision rule as

$$\hat{\mathbf{x}} = \arg\min_{\mathbf{x} \in \mathbb{S}} \|\mathbf{y} - \mathbf{H}\mathbf{x}\|^2 \tag{2.17}$$

where $\mathbb{S}$ is the set of all possible antenna combinations and APM symbols.

2.4 Cooperative Communications

As stated in Section 2.1, diversity increases the reliability of the communication channel. Multiple antennas on novel communication transceivers achieve spatial diversity with the advances in theory on MIMO systems. On the other hand, the

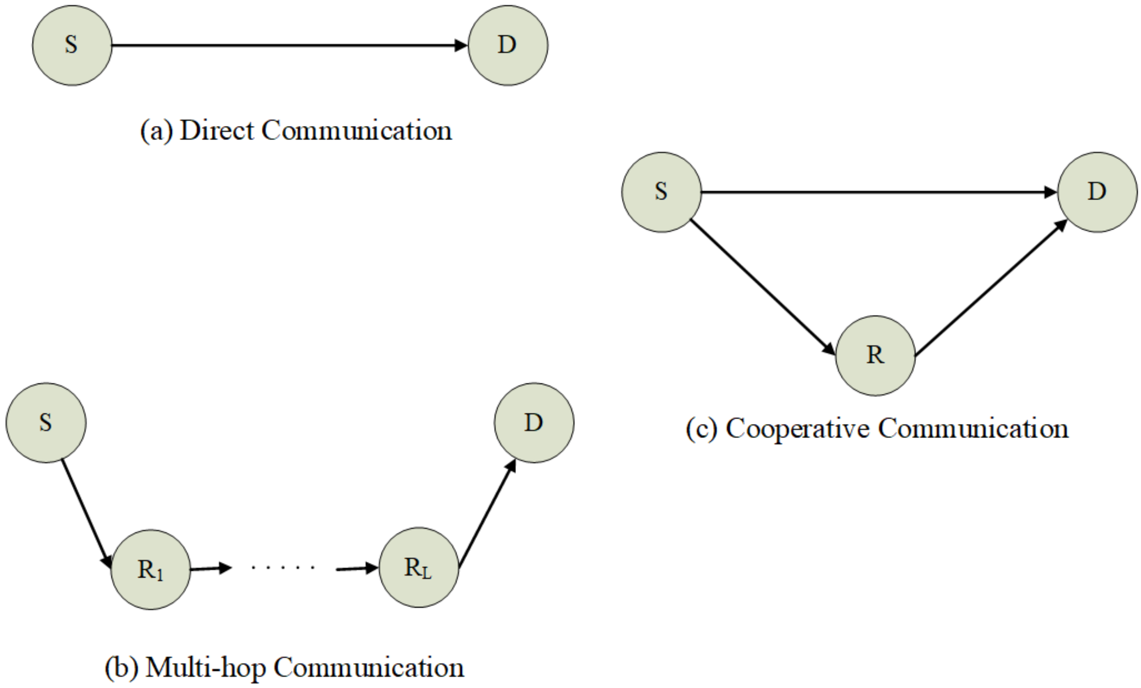

Figure 2.3 : Direct, Multi-hop and Cooperative Communications.

demand for small size and cost effective devices increases. Wireless devices such as cellular phones and sensor networks can not be practical to implant more than one antenna. In this case, the idea of sharing antennas to form multiple antenna system has emerged. Cooperative communication is an effective way to form a distributed antenna system where the nodes cooperate with each other.

Communication from a single source to a single destination without any help is called *direct* or *point-to-point* communication as in Figure 2.3(a). As the range increases, direct communication may be impractical. Hence, locating a relay in a long link to enhance signal quality overcomes this problem. The chain of the point-to-point link as in Figure 2.3(b) is called *multi-hop* communication [23]. Figure 2.3(c) illustrates the *relay* or *cooperative* communication which can be thought as the combination of the direct and multi-hop channel. The works on relay channels date back to [24], [25] and [26] but, the works that helped to raise interest to cooperative communication in recent years is [9], [27] and [28].

A relay channel consists of (at least) three nodes: *Source* (S), *Relay* (R) and *Destination* (D). S sends its information to D with the aid of R where it is not actually an information source. Generally, cooperative communication has two phases: first, when S transmits and, R and D receive, it is called *Broadcasting* phase, and second, when

S and R transmits (or just R forwards) and D receive, it is called *Multiple-access* (MAC) (or *Relaying*) phase [29]. Cooperative communication can be classified into four different modes based on the above two phases [29]:

First Phase	Second Phase
S $\longrightarrow$ R, D	S, R $\longrightarrow$ D
S $\longrightarrow$ R	S, R $\longrightarrow$ D
S $\longrightarrow$ R, D	R $\longrightarrow$ D
S $\longrightarrow$ R	R $\longrightarrow$ D (Multi-hop)

Table 2.1 : Cooperative Communication Modes

2.4.1 Half-duplex and full-duplex relaying

A cooperative communication system can operate in two modes: Half-duplex mode and Full-duplex mode. An R is said in half-duplex mode when the broadcasting and relaying phases do not operate in the same band, i.e., the transmission and reception channels are orthogonal. This orthogonality can be in time-domain, in frequency-domain or using orthogonal signals in time-frequency plane. If the transmission and reception of R are in the same band, it is called full-duplex mode. Undoubtedly, when R operates in full-duplex mode, transmitted and received signals interfere. Conceptually, interference cancellation can be applied when R knows the transmitted signal. However, full cancellation can not be made when the transmitted signal is not powerful than the received signal [29]. On the other hand, optimal relay selection with dynamic switching between half-duplex and full-duplex modes are proposed to overcome the effect of interference. Furthermore, some works, such as self interference mitigation, antenna selection and transceiver beamforming have been introduced to improve the performance of full-duplex relay system [30]. Since there are some difficulties in accurate interference cancellation, full-duplex mode is not commonly used. In this thesis, we consider the half-duplex relaying.

2.4.2 Relaying protocols

Depending on the nature of the relay, two relay strategies can be mentioned in general. In the first one, R just amplifies the received signal and transmits to D. In the second one, R decodes the received signal from S, re-encodes it and forwards to D.

2.4.2.1 Amplify-and-forward relaying

In amplify-and-forward (AF) relaying, R amplifies the received signal and forwards to destination without using any processing techniques. This protocol is also called *non-regenerative relaying* protocol [23]. R does not need to be smart and does not have knowledge of modulation or encoding schemes that is applied at S. AF protocol can be very practical especially when the S-R link does not allow the signal to be decoded correctly.

Let, S transmits the symbol x_n, $n = 0, \ldots, N-1$ to both R and D in the first phase, where N is the length of the codeword. The received signal at R and D is

$$y_n^{SR} = \sqrt{P_s} h^{SR} x_n^s + w_n^{SR} \tag{2.18}$$

$$y_n^{SD} = \sqrt{P_s} h^{SD} x_n^s + w_n^{SD} \tag{2.19}$$

respectively, where P_s is transmission power of S, $w_n^{SR(SD)}$ is the AWGN and modeled as $\mathscr{CN}(0, N_0)$ with noise spectral density $N_0/2$ per dimension. In the second phase, R amplifies the received signal and sends to D. The received signal at D becomes

$$\begin{aligned} y_n^{RD} &= G\sqrt{P_r} h^{RD} y_n^{SR} + w_n^{RD} \\ &= G\sqrt{P_s P_r} h^{RD} h^{SR} x_m^s + \tilde{w}_n \end{aligned} \tag{2.20}$$

where $G = \sqrt{\frac{1}{P_s \left|h^{SR}\right|^2 + N_0}}$ is the normalization factor in order to yield the unit energy transmitted signal when R knows the $\left|h^{SR}\right|^2$ and $\tilde{w}_n = G\sqrt{P_r} h^{RD} w_n^{SR} + w_n^{RD}$. If R has no knowledge about the S-R link, the expected value of it, i.e., $E\left\{\left|h^{SR}\right|^2\right\}$, can be used. This scheme is called *fixed-gain AF relaying*.

The effective SNR at the maximum ratio combiner (MRC) is

$$\gamma_{AF} = \gamma_{SD} + \frac{\gamma_{SR}\gamma_{RD}}{\gamma_{SR} + \gamma_{RD} + 1} \tag{2.21}$$

where $\gamma_{SD/SR/RD} = P_s \left|h^{SD/SR/RD}\right|^2 / N_0$.

2.4.2.2 Decode-and-forward relaying

In decode-and-forward (DF) protocol, R decodes the received signal, re-encodes into a new symbol and forwards to D. This relaying scheme is also called *regenerative relaying* [23].

In the literature, some studies such as [9], [23] have defined the DF system that R forwards only when it decodes correctly. This is not always implemented due to limited transmitter and receiver capabilities and/or lack of knowledge of channel information. We, in here, define the DF protocol that R always transmits the decoded signal. This scheme is also known as *demodulate-and-forward (DeF)* [23]. Of course, if R makes an error, it will also be transmitted to D which produces an error propagation.

The first phase is same as AF relaying, hence, the received signal at R and D will be as in (2.18) and (2.19). In the second phase, R decodes the received signal based on a decision rule and re-encodes to a new symbol such as x_n^r, $n = 0, \ldots, N-1$. The received signal at D is

$$y_n^{RD} = \sqrt{P_r} h^{RD} x_n^r + w_n^{RD}. \tag{2.22}$$

Note that, if R decodes correctly then $x_n^r = x_n^s$ and the SNR at the output of MRC is

$$\gamma_{DF} = \gamma_{SD} + \gamma_{RD}. \tag{2.23}$$

2.4.3 Dynamic relaying techniques

In [9], it is stated that diversity order of the DF scheme is 1 unless R successfully decodes the transmitted signal. Since the successful decoding capability of R depends on the S-R link, the *selection DF relaying (SR)* is introduced.

On the other hand, AF and DF protocols may not use the bandwith efficiently, compared to direct transmission. In order to use bandwith effectively, one can use *incremental relaying (IR)*.

2.4.3.1 Selection DF relaying

In the SR, S retransmits the message itself and R remains silent in the second phase when the instantaneous SNR of the S-R link below a pre-determined threshold. As in the classical DF scheme, S transmits its information to R and D in the first phase. Let we assume that S and R have knowledge of the CSI on the S-R link. If the instantaneous SNR of the S-R link is below a threshold, S infers that R makes error with high probability and re-transmits its message in the second phase. During this second phase, R will not forward the message. Most of the time, the pre-determined threshold is chosen as $2R \leq \log_2(1+\gamma_{SR}) \Rightarrow \gamma_{SR} \gtreqless 2^{2R}-1$ where R is the transmission rate. Than, the received signal at D is

$$y_n^{SR} = \begin{cases} \sqrt{P_r} h^{RD} x_n^s + w_n^{RD} & \text{if } \gamma_{SR} \geq 2^{2R}-1 \\ \sqrt{P_s} h^{SD} x_n^s + w_n^{RD} & \text{if } \gamma_{SR} < 2^{2R}-1 \end{cases} \tag{2.24}$$

where we use the fact that $x_n^r = x_n^s$ when $\gamma_{SR} \geq 2^{2R}-1$. The effective SNR will then be

$$\gamma_{SDF} = \begin{cases} \gamma_{SD} + \gamma_{RD} & \text{if } \gamma_{SR} \geq 2^{2R}-1 \\ 2\gamma_{SD} & \text{if } \gamma_{SR} < 2^{2R}-1 \end{cases} \tag{2.25}$$

.

2.4.3.2 Incremantal relaying

The information of S is transmitted over two time period in AF and DF relaying even though D is likely to successfully decode the received signal for the first phase. This can be improved by the possibility that R does not transmit, if the SNR from S-R link is higher than a certain threshold. Thus, the second phase is not always needed and the bandwidth efficiency is improved. The IR scheme can be implemented with a simple feedback from D.

S transmits its message to R and D in the first phase. If the instantaneous SNR of S-D link is above a threshold, i.e., $\gamma_{SD} \geq 2^{2R}-1$, D broadcasts an acknowledgment (ACK) to S and R to indicate that decoding is successful and S transmits a new message while R remains silent. If it is below the threshold, D sends negative ACK (NACK) and R forwards the received signal as regular AF or DF relaying. Hence, the effective SNR will be same as direct transmission for the first case and regular AF or DF relaying for the second case.

Since the transmission is utilized over one or two period, transmission rate varies and average rate have to be computed. Therefore average transmission rate can be evaluated as

$$\bar{R} = 2R\,\mathrm{Pr}\left(\gamma_{SD} \geq 2^{2R}-1\right) + R\,\mathrm{Pr}\left(\gamma_{SD} < 2^{2R}-1\right) \tag{2.26}$$

.

2.5 Literature Survey

In this section, capacity and outage probability analysis of SM, low-complexity detection algorithms for GSM and cooperative SM systems recently proposed in the literature have been examined.

2.5.1 Capacity and outage probability analysis of SM

Since SM scheme exploits the spatial domain for data transmission, its capacity calculation is different than the classical systems. The first study for the analysis of the SM capacity is given in [31]. In this study, the mutual information of the space and the APM domains are calculated separately to obtain the total capacity for the SM scheme. In [32], this work is extended to SM with finite alphabet inputs for MISO channels. A comprehensive analysis is given in [33] where the mutual information of the ML transmit antenna detector for the space domain is analyzed and SM-MISO capacities are given for complex Gaussian, real Gaussian and constant modulus APM distributions. Additionally, a lower bound is computed for SM-MIMO capacity for the complex Gaussian APM distribution.

An outage probability analysis for the SM system based on transmit antenna selection can be found in [34]. However, the capacity of the SM system is assumed the same as conventional SIMO capacity. In [35], the outage probability of SM is given for the transmit antenna selection only and considering high SNR values.

As can be seen from the previous works, the outage probability analysis of an SM system is not given in the literature. In this thesis, a comprehensive study on SM's outage probability is investigated. Furthermore, this analysis is extended to a cooperative system under different relaying techniques and different diversity protocols.

2.5.2 Detection algorithms for GSM

Some research can be found in the literature to find low-complexity detectors with near-ML performance. Among these, one can mention the following works.

In [7], a MRC algorithm is introduced to detect active antenna indices. After determining the antennas, PSK/QAM symbols are estimated. In [36], a decorrelator based detection algorithm is presented. Although it reduces the receiver complexity, BER performance is much higher than that of ML detection. In [37], an ordered block minimum mean square error (OB-MMSE) detector which is based on a block MMSE processing method is proposed. In [38], Gaussian approximation and QR projection is used to detect the antenna subset index. After detecting the antenna subset index, classical ML detector is used to identify the transmitted symbol. The advantage of Bayesian compressive sensing, which uses the inherent sparsity of GSM signals, is investigated in [39]. This algorithm, which is called enhanced Bayesian compressive sensing (EBCS), includes two stages where in the first stage, the active antenna indices are estimated with the help of hyper-parameter vector and the BCS algorithm reconstructs the transmitted signal. In the second stage, re-estimation is designed to check and correct the errors from the first stage. Additionally, in order to investigate low complexity algorithms, sphere decoding (SD) algorithms are investigated in [40], [41] and [42].

2.5.3 Cooperative SM systems

Combining the advantages of SM and cooperative communications has been recently studied in the literature [43]-[44]. The first study is performed by Serafimovski *et.al.* in [43] in which a dual-hop SM system is proposed. In this multi antenna dual-hop system, R uses DF protocol to support communication between S and D. In [44], BER performance of a dual-hop SSK system applying AF relaying is investigated for single receive antenna at both R and D. A real cooperative scenario in which S sents its information to R and D in the first time slot is considered in [45]. In this system, the multi antenna S transmits its data using SSK to N relays and D (all nodes have single transmit and receive antennas) in the first time slot and N relays amplify the incoming signal and retransmit to D in the following N time slots. In the same study, the use of DF strategy is investigated when the relays which correctly detect the source symbol are permitted to continue. Since the relays have single antenna, communication between R and D can not be performed with SSK. In [46], the same authors improved the dual-hop SSK system in [44] to an N-relay system considering opportunistic relaying to increase the spectral efficiency. Cooperative SM system with multi-antenna S, single transmit/receive antenna, multiple-R using DF strategy and single receive antenna D is considered in [47]. The first cooperative system in which multi antenna nodes use SSK with DF is introduced in [48]. However, the exact BER analysis for SSK-DF system with incremental relaying and selection combining at D is derived only for two transmit antennas at S and R.

Combining the advantages of both SM and STBC, the STBC-SM scheme is introduced in [49]. Applying the idea of STBC-SM to the cooperative systems is investigated in [50] where end-to-end pairwise error probability (PEP) analysis and optimal source-relay power allocation is presented.

Furthermore, the combination of SM/SSK and physical layer network coding (PLNC) is performed in [51], [52], [53]. Since a framework for PLNC is the cooperative system, our study in this thesis can be the precursor for PLNC-SM systems. PLNC is proposed to increase the spectral efficiency of cooperative communications where different users share the same relay to communicate with each other at the same time.

As can be seen from the above studies and to the best of our knowledge a comprehensive work on cooperative communications, where all nodes have multiple transmit and/or receive antennas, has not been done before.

3. OUTAGE PROBABILITY ANALYSIS OF SM

The capacity analysis of the SM scheme is different than that of the MIMO systems. Since the information is conveyed not only through the M-ary constellation domain but also through the antenna domain, the capacity of SM is expressed as the sum of the capacities of these two domains. Due to this summation, the outage probability of SM differs from the conventional systems. A detailed analysis of the outage probability of SM has not been given in the literature yet. In this chapter, we derive the outage probability performance of the classical SM system.

3.1 System Model

Consider a MIMO system consisting of N_t transmit and N_r receive antennas. A unit energy, $E[\mathbf{x}^H\mathbf{x}] = 1$, SM symbol can be regarded as $\mathbf{x} = [\underbrace{0,\ 0,\ \ldots,\ 0,}_{l-1}\ x_q,\ \underbrace{0,\ \ldots,\ 0}_{N_t-l}]^T$, where l is the active antenna index and x_q is the APM constellation symbol. The received signal is given as

$$\mathbf{y} = \mathbf{h}_l x_q + \mathbf{n} \tag{3.1}$$

where $\mathbf{h}_l \in \mathbb{C}^{N_r} \times 1$ vector is the l^{th} column of the MIMO channel matrix, $\mathbf{H} = [\ \mathbf{h}_1\ \ \mathbf{h}_2\ \ \ldots\ \ \mathbf{h}_{N_t}\ ]$, with each element being independent and identically distributed (i.i.d) with $\mathscr{CN}(0,1)$ and $\mathbf{n}$ is the $N_r \times 1$ additive white Gaussian vector whose entries are modeled as $\mathscr{CN}(0,N_0)$ with noise spectral density $N_0/2$ per dimension. The detector having ideal channel state information (CSI) estimates antenna index, l, and the APM signal, x_q, with the ML decision rule as follows:

$$[\hat{l},\hat{x}_q] = \arg\min_{l,x_q} \left\|\mathbf{y} - \mathbf{h}_l x_q\right\|^2. \tag{3.2}$$

3.2 SM Capacity And Outage Probability

Since the information is conveyed not only by classical APM symbols, but also by the space domain in SM systems, the SM capacity is computed as the capacity of conventional modulation plus the capacity of the space domain [31],

$$I(X_q, X_{ch}; Y) = I(X_{ch}; Y) + I(X_q; Y \mid X_{ch}) \tag{3.3}$$

where X_q and X_{ch} represents APM and antenna domain symbol spaces, respectively and Y is the output symbol space. If the space domain symbols are considered equiprobable for flat fading channel, the mutual information between APM symbol X_q and the output symbol Y conditioned on the antenna domain symbol X_{ch} is given as,

$$\begin{aligned} I(X_q; Y \mid X_{ch}) &= H(Y \mid X_{ch}) - H(Y \mid X_q, X_{ch}) \\ &= \frac{1}{N_t} \sum_{l=1}^{N_t} \log_2(1 + \rho \, \|\mathbf{h}_l\|^2) \end{aligned} \tag{3.4}$$

where ρ is the average signal to noise ratio (SNR) and the input symbols are considered as i.i.d. complex Gaussian ensembles. The first term in the right hand side of (3.3) can be calculated with the well-known differential entropy equation as,

$$I(X_{ch}; Y) = H(Y) - H(Y \mid X_{ch}). \tag{3.5}$$

Generally, a closed form expression for (3.5) is difficult to obtain and its value can be calculated numerically. However, when no antenna index error occurs at the receiver, the information carried by space domain is transferred fully and (3.5) becomes $I(X_{ch}; Y) = \log_2(N_t)$. Hence, the SM capacity can be bounded as [35]

$$\frac{1}{N_t} \sum_{l=1}^{N_t} \log_2 \left(1 + \rho \, \|\mathbf{h}_l\|^2\right) \leq C_{SM} \leq \frac{1}{N_t} \sum_{l=1}^{N_t} \log_2 \left(1 + \rho \, \|\mathbf{h}_l\|^2\right) + \log_2 (N_t). \tag{3.6}$$

In (3.6), the right hand side is the full SM capacity with no antenna index error.

SM capacity in (3.3) averages the mutual information of each link from all transmit antennas. However, since only one antenna is active for a given time interval, the outage event occurs over the activated link. Hence, the instantaneous capacity have to be derived for the outage probability analysis.

Consider the set of N_t channel vector norms, $\left\{\|\mathbf{h}_l\|^2\right\}_{l=1}^{N_t}$, an outage event occurs when $I(X_q, X_{ch=\min}; Y) \leq R$. Here, R is the pre-determined rate in bits/s/Hz and $X_{ch=\min}$ represents the space domain symbols when the channel with the lowest vector norm, $\|\mathbf{h}_{\min}\|^2$, is chosen. In this way, (3.4) and (3.5) respectively simplify to $\log_2 \left(1 + \rho \, \|\mathbf{h}_{\min}\|^2\right)$ and $H(Y) - H(Y \mid X_{ch=\min})$. Therefore, the instantenous mutual information for the $l = \min$ link is

$$I(X_q, X_{ch=\min}; Y) = \log_2 \left(1 + \rho \, \|\mathbf{h}_{\min}\|^2\right) + H(Y) - H(Y \mid X_{ch=\min}). \tag{3.7}$$

Here, entropies can be written as,

$$H(Y) = -\int_{\mathbf{y}} f_Y(\mathbf{y})\log_2 f_Y(\mathbf{y})d\mathbf{y} \tag{3.8}$$

$$\begin{aligned} H(Y \mid X_{ch=\min}) &= -\int_{\mathbf{y}} f_Y(\mathbf{y} \mid x_{ch=\min})\log_2 f_Y(\mathbf{y} \mid x_{ch=\min})d\mathbf{y} \\ &= \log_2 \det(\pi e \mathbf{K}_{\min}) \end{aligned} \tag{3.9}$$

where $f_Y(\mathbf{y}) = \frac{1}{N_t}\sum_{l=1}^{N_t}\frac{1}{\det(\pi\mathbf{K}_l)}\exp\left(-\mathbf{y}^H\mathbf{K}_l^{-1}\mathbf{y}\right)$ and $\mathbf{K}_l = \mathbf{h}_l\mathbf{h}_l^H\sigma_x^2 + N_0\mathbf{I}_{N_r}$. Besides, $f_Y(\mathbf{y} \mid x_{ch=\min}) = \frac{1}{\det(\pi\mathbf{K}_{\min})}\exp\left(-\mathbf{y}^H\mathbf{K}_{\min}^{-1}\mathbf{y}\right)$ and $\mathbf{K}_{\min} = \mathbf{h}_{\min}\mathbf{h}_{\min}^H\sigma_x^2 + N_0\mathbf{I}_{N_r}$ where σ_x^2 is the signal power. (3.8) can be computed with numerical integration. On the other hand, since (3.8) and (3.9) depend on the space domain symbols which are in fact the channel gains, $H(Y) - H(Y \mid X_{ch=\min})$ have to be averaged. After computing the $H(Y) - H(Y \mid X_{ch=\min})$, the exact outage probability may be expressed as

$$\begin{aligned} P_{out}^{SM} &= \Pr\left\{I(X_q, X_{ch=\min}; Y) \le R\right\} \\ &= \Pr\left\{\|\mathbf{h}_{\min}\|^2 \le \frac{2^{R-H(Y)+H(Y|X_{ch=\min})}-1}{\rho}\right\}. \end{aligned} \tag{3.10}$$

The pdf of the $\nu \triangleq \|\mathbf{h}_{\min}\|^2$ which is the minimum of the channel vector norms, $\ell \triangleq \mathbf{h}_l$, $l = 1,\ldots,N_t$, (Erlang random variables) can be calculated with the help of order statistics as

$$f_\nu(x) = N_t\left(1 - F_\ell(x)\right)^{N_t-1} f_\ell(x) \tag{3.11}$$

where the $F_\ell(x)$ and $f_\ell(x)$ denote the cdf and the pdf of the channel vector norms, respectively. Hence,

$$\begin{aligned} f_\nu(x) &= N_t\left(e^{-x}\sum_{k=0}^{N_r-1}\frac{x^k}{k!}\right)^{N_t-1}\frac{x^{N_r-1}e^{-x}}{(N_r-1)!} \\ &= \frac{N_t e^{-N_t x}}{(N_r-1)!}\sum_{k=0}^{(N_r-1)(N_t-1)} C(k)x^{N_r+k-1} \end{aligned} \tag{3.12}$$

where $C(k)$ is the coefficient of x^k in the expansion of $\left(\sum_{k=0}^{N_r-1}\frac{x^k}{k!}\right)^{N_t-1}$. Therefore, the outage probability for SM systems is given as

$$\begin{aligned} P_{out}^{SM} &= \int_0^{\rho^{th}} \frac{N_t e^{-N_t x}}{(N_r-1)!}\sum_{k=0}^{N} C(k)x^{N_r+k-1} \\ &= \sum_{k=0}^{N}\frac{C(k)\gamma(N_r+k, N_t\rho^{th})}{N_t^{N_r+k-1}(N_r-1)!}. \end{aligned} \tag{3.13}$$

In (3.13), $\rho^{th} = \frac{2^{R-H(Y)+H(Y|X_{ch=\min})}-1}{\rho}$, $\gamma(\cdot,\cdot)$ represents lower incomplete Gamma function and $N = (N_r - 1)(N_t - 1)$.

Furthermore, the lower and upper bounds for outage probability of SM can be written as

$$\Pr\left\{\log_2\left(1+\rho\,\|\mathbf{h}_{\min}\|^2\right) \leq R\right\} \geq P_{out} \geq \Pr\left\{\log_2\left(1+\rho\,\|\mathbf{h}_{\min}\|^2\right) + \log_2(N_t) \leq R\right\}. \tag{3.14}$$

3.3 Performance Evaluation

In this section, we present computer simulation results for the outage probability of SM. Monte Carlo simulations are realized for at least 10^7 channel uses as a function of received SNR (ρ) and obtained curves are compared with the analytical results. On the other hand, the SM system is compared with the conventional modulation techniques on an equal spectral efficiency basis.

SM with $R = 3$ and $R = 4$ bits/s/Hz and corresponding M-PSK systems are compared in Figure 3.1. The outage probabilities of SM systems with $N_t = 4$, BPSK and $N_t = 4$, QPSK are simulated. The upper and lower bounds are also depicted for comparison purposes. Additionally, they are compared with 8-PSK and 16-PSK systems. As seen from Figure 3.1, the exact outage probability lies between the upper and lower bounds, as expected. At low SNR values, it is highly probable that the receiver makes antenna index error which makes the exact curve closer to upper bound. At high SNR values, it is less likely to have antenna index error and the outage probability converges to lower bound.

Also note that the analytical curves obtained from (3.13) exactly match with the simulation results. The outage probability performance of SM system gets better at high SNR values as expected. On the other hand, since we selected the minimum channel gains for computing the outage probability, this performance is the worst case scenario. Hence, the better outage probability performance can be expected in practice.

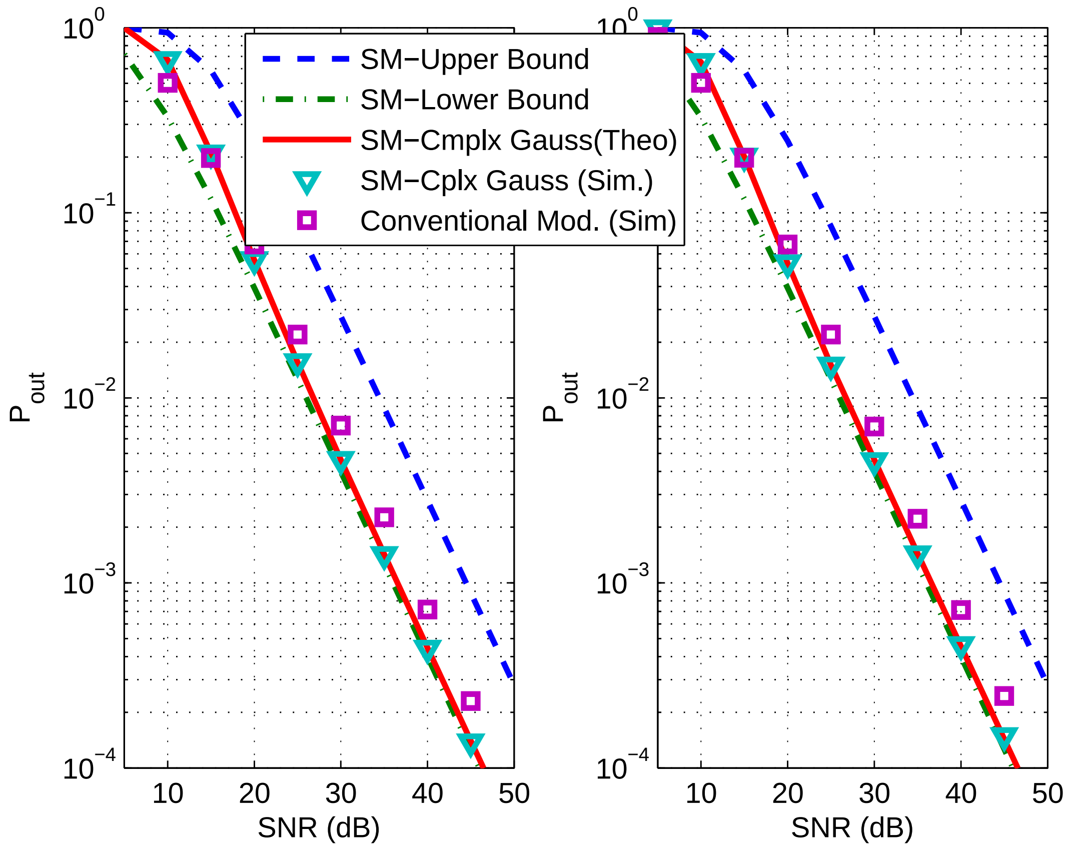

Figure 3.1 : Outage probability performance of $N_t = 4$, BPSK (left) and $N_t = 4$, QPSK (right) SM systems compared with SISO 8-PSK and 16-PSK ($N_r = 1$).

4. A NEW LOW-COMPLEXITY DETECTION ALGORITHM FOR GSM

As stated in Chapter 2, Generalized Spatial Modulation (GSM) is an extension of SM which is considered significant for the next generation communication systems. Optimal detection process for the GSM is the ML detection. However, the receiver is much more complicated than SM due to inter-antenna interference and/or increased number of combinations. Therefore, computational complexity grows with the number of transmit antennas and the signal constellation size. In the literature, sub-optimal techniques such as maximum ratio combining (MRC), zero forcing (ZF) and minimum mean square error (MMSE) are presented. In this chapter, we introduce a novel and simple detection algorithm which uses sub-optimal method to detect likely antenna combinations. Once the antenna indices are detected, ML detection is utilized to identify the transmitted symbol(s). Our proposed algorithm reduces the search complexity while achieving a near optimum solution. Computer simulation results show that the proposed algorithm performs close to the optimal (ML) detection resulting in a minimum increase of computational complexity.

4.1 System Model

As introduced in Section 2.3.2, the GSIM transmitted symbol can be written as

$$\mathbf{x}_{GSIM} = \begin{bmatrix} s_1 & 0 & s_2 & 0 & 0 \end{bmatrix}^T. \tag{4.1}$$

where the total information bits are $m = 7$ and first three bits are mapped to first and third antennas and remaining four bits are mapped to two 4-PSK/QAM symbols. When we consider the GSM symbol, total number of bits will be $m = 5$ and only one PSK/QAM symbol takes place in the transmitted symbol as

$$\mathbf{x}_{GSM} = \begin{bmatrix} s & 0 & s & 0 & 0 \end{bmatrix}^T. \tag{4.2}$$

The received signal for both transmissions can be written as

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{n} \tag{4.3}$$

where $\mathbf{n} \in \mathbb{C}^{N_r \times 1}$ is the AWGN vector with $\mathcal{CN}(0, N_0)$.

The detector at the receiver, which has the ideal channel state information, detects the GSM/GSIM symbols applying the ML decision rule as

$$\hat{\mathbf{x}} = \arg\min_{\mathbf{x} \in \mathbb{S}} \|\mathbf{y} - \mathbf{H}\mathbf{x}\|^2 \tag{4.4}$$

where $\mathbb{S}$ is the set of all possible antenna combinations and PSK/QAM symbols and $\|.\|$ is the Euclidean norm of the vector.

4.2 New Simple Detection Algorithms

ML detection in (4.4) can be re-written for GSIM and GSM respectively in another form as

$$\left[\hat{\imath}, \hat{\mathbf{s}}\right] = \arg\min_{i, \mathbf{s}} \|\mathbf{y} - \mathbf{H}_i \mathbf{s}\|^2 \tag{4.5a}$$

$$\left[\hat{\imath}, \hat{s}\right] = \arg\min_{i, s} \|\mathbf{y} - (\mathbf{h}_e)_i s\|^2 \tag{4.5b}$$

where $\mathbf{s} = [s_1, s_2, \ldots, s_{N_{act}}]^T$ is the transmitted symbol vector, i is the index of antenna combinations set where $i \in 1, 2, \ldots, N_c$. $\mathbf{H}_i$ is composed of active antenna columns of $\mathbf{H}$ and $(\mathbf{h}_e)_i$ is the sum of the columns of $\mathbf{H}_i$. (4.5) has an excessive computational complexity that grows exponentially with the number of transmit antennas. Furthermore, another exponential increase is caused by GSIM complexity where the antenna combinations carry different PSK/QAM symbols. One of the low complexity solution is the sub-optimal solution which is also called Zero Forcing (ZF) solution. This sub-optimal solution is based on the Least Squares solution of the ML detection,

$$\mathbf{s}_i^{LS} = \left(\mathbf{H}_i^H \mathbf{H}_i\right)^{-1} \mathbf{H}_i^H \mathbf{y} \tag{4.6}$$

where $(.)^H$ is the Hermitian operation. We have to emphasize that this solution is conditioned to the correct information about i.

In this chapter, we propose a new and simple reduced complexity detection algorithm based on the LS solution in (4.6).

4.2.1 Low complexity detection algorithm

Using the LS solution in (4.6), another version of the cost function in (4.5) can be given as

$$\left[\hat{\imath},\hat{\mathbf{s}}\right] = \arg\min_{i,\mathbf{s}} \left\|\mathbf{y}-\mathbf{H}_i\mathbf{s}_i^{LS}\right\|^2 + \left(\mathbf{s}_i^{LS}-\mathbf{s}\right)^H \mathbf{H}_i^H\mathbf{H}_i\left(\mathbf{s}_i^{LS}-\mathbf{s}\right). \tag{4.7}$$

Assuming that the LS solution would be close to the true solution, the second part of (4.7) can be ignored which leads to the reduced cost function

$$\left[\hat{\imath},\hat{\mathbf{s}}\right] = \arg\min_{i} \left\|\mathbf{y}-\mathbf{H}_i\mathbf{s}_i^{LS}\right\|^2 \tag{4.8}$$

which only depends on i. (4.8) can be expanded as

$$\begin{aligned}\left[\hat{\imath},\hat{\mathbf{s}}\right] &= \arg\min_{i} \left(\mathbf{y}-\mathbf{H}_i\mathbf{s}_i^{LS}\right)^H\left(\mathbf{y}-\mathbf{H}_i\mathbf{s}_i^{LS}\right) &(4.9)\\ &= \arg\min_{i}\left\{\|\mathbf{y}\|^2-\left(\mathbf{s}_i^{LS}\right)^H\mathbf{H}_i^H\mathbf{y}-\mathbf{y}^H\mathbf{H}_i\mathbf{s}_i^{LS}+\left(\mathbf{s}_i^{LS}\right)^H\mathbf{H}_i^H\mathbf{H}_i\mathbf{s}_i^{LS}\right\}. &(4.10)\end{aligned}$$

$\|\mathbf{y}\|^2$ in (4.10) has no effect in minimization and can be discarded. Since the minimizing a function over its argument is equivalent to maximizing that function over the same argument with a sign change, the minimization problem in (4.8), with using (4.6), can be re-expressed as

$$\hat{\imath}^{GSIM} = \arg\max_{i}\left\{\mathbf{y}^H\mathbf{H}_i\left(\mathbf{H}_i^H\mathbf{H}_i\right)^{-1}\mathbf{H}_i^H\mathbf{y}\right\} \tag{4.11a}$$

$$\hat{\imath}^{GSM} = \arg\max_{i}\frac{\left|(\mathbf{h}_e)_i^H\mathbf{y}\right|}{\|(\mathbf{h}_e)_i\|}. \tag{4.11b}$$

PSK/QAM symbols are then solved with the quantization (slicing) operation after the detection of the active antennas, $\hat{\mathbf{s}} = Q\left\{\mathbf{s}_{\hat{\imath}}^{LS}\right\}$.

In [7] and [36], detected antenna combinations are found as in (4.11) which selects only the maximum value. The ML detection, on the other hand, searches not only the maximum value but also all possible outputs which are as many as 2^{m_a} in addition to PSK/QAM symbol constellation points. Hence, the complexity grows with the number of transmit antennas and high modulation orders.

To reduce the ML complexity, one has to decrease the number of combinations to be tested. Accordingly, we take the N best estimate of (4.11) to find the optimum i

$$\tilde{\imath} = \arg\max_{i}\left\{\mathbf{y}^H\mathbf{H}_i\left(\mathbf{H}_i^H\mathbf{H}_i\right)^{-1}\mathbf{H}_i^H\mathbf{y}\right\}\Big|_1^N, \tilde{\imath} = 1,\ldots,N \tag{4.12}$$

where $\{\cdot\}|_1^N$ denotes the first N values of (4.12). Therefore, our modified ML detection can be written as for GSIM and GSM respectively as

$$\left[\hat{i}_{LC}, \hat{\mathbf{s}}_{LC}\right]^{GSIM} = \arg\min_{\tilde{i}} \|\mathbf{y} - \mathbf{H}_{\tilde{i}}\tilde{\mathbf{s}}_{\tilde{i}}\|^2 \tag{4.13a}$$

$$\left[\hat{i}_{LC}, \hat{s}_{LC}\right]^{GSM} = \arg\min_{\tilde{i}} \|\mathbf{y} - (\mathbf{h}_e)_{\tilde{i}}\tilde{s}_{\tilde{i}}\|^2 \tag{4.13b}$$

where $\tilde{\mathbf{s}}_{\tilde{i}} = Q\left\{\mathbf{s}_{\tilde{i}}^{LS}\right\}$. Hence, we can state Algorithm 1 as

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Algorithm 1: Low Complexity GSM/GSIM Detection Algorithm

Data: $\mathbf{y}$, $\mathbf{H}$, N_t, N_{act}
Initialization:
  $\mathbf{A}_i = \mathbf{H}_i^H\mathbf{H}_i$, $\mathbf{b}_i = \mathbf{H}_i^H\mathbf{y}$, $i \in \mathbb{I} = i_1, \ldots, i_{N_c}$
  solve the linear system $\mathbf{A}_i\mathbf{s}_i^{LS} = \mathbf{b}_i$
  $z_i = \mathbf{b}_i^H\mathbf{s}_i^{LS}$, $\mathbf{z} = [z_1, \ldots, z_{N_c}]^T$
  $[\tilde{i}(1), \ldots, \tilde{i}(N)] = \arg sort(\mathbf{z})\,|_1^N$
for $\tilde{i} \leftarrow 1$ **to** N **do**
  $\tilde{\mathbf{s}}_{\tilde{i}} = Q\left\{\mathbf{s}_{\tilde{i}}^{LS}\right\}$
  $d_{\tilde{i}} = \|\mathbf{y} - \mathbf{H}_{\tilde{i}}\tilde{\mathbf{s}}_{\tilde{i}}\|^2$
end
$\hat{i}_{LC} = \arg\min d_{\tilde{i}}$
$\hat{\mathbf{s}}_{LC} = \tilde{\mathbf{s}}_{\hat{i}_{LC}}$

Remark. For GSM, $\tilde{s}_{\hat{i}_{LC}}$ is the optimum solution for optimum $\hat{i}_{LC}$. On the other hand, for GSIM, the minimization of the second part of (4.7) must be carried out for every i to find the optimum solution. For this reason, $\tilde{\mathbf{s}}_{\hat{i}_{LC}}$ is defective to reach closer to the ML solution especially for long $\tilde{\mathbf{s}}_{\hat{i}_{LC}}$ vector size, i.e., high activated transmit antenna numbers.

4.2.2 Reduced complexity detection algorithm for GSIM

As stated in **Remark** and as will be observed in Section IV, $\tilde{\mathbf{s}}_{\hat{i}_{LC}}$ is not sufficient to find the near ML solution. A simple way to approach the ML performance is to use all possible symbol combinations in (4.13a) while keeping the N minimum

$$\left[\hat{i}_{RC}, \hat{\mathbf{s}}_{RC}\right]^{GSIM} = \arg\min_{\tilde{i},\mathbf{s}} \|\mathbf{y} - \mathbf{H}_{\tilde{i}}\mathbf{s}\|^2. \tag{4.14}$$

The algorithm for this new detection rule can be seen in Algorithm 2.

4.2.3 Reduced Complexity Detection Algorithm with SD

Algorithm 2: Reduced Complexity GSIM Detection Algorithm

Data: $\mathbf{y}$, $\mathbf{H}$, N_t, N_{act}, M
Initialization: Same as Alg. 1
for $\mathbf{s} \leftarrow 1$ **to** $M^{N_{act}}$ **do**
  for $\tilde{i} \leftarrow 1$ **to** N **do**
    $d_{\tilde{i},\mathbf{s}} = \|\mathbf{y} - \mathbf{H}_{\tilde{i}}\mathbf{s}\|^2$
  end
end
$\left[\hat{i}_{RC}, \hat{\mathbf{s}}_{RC}\right] = \arg\min d_{\tilde{i},\mathbf{s}}$

Although the Algorithm 2 reduces the search space in terms of the antenna indices, it checks every constellation point for PSK/QAM symbols. In order to take one more step and reduce the search space further, we can apply SD to the second part of (4.7). The sphere decoder examines only the points that lie inside a sphere with a radius r

$$\mathbf{s}_{\tilde{i}}^{SD} = \arg\min_{\mathbf{s}} \left(\mathbf{s}_{\tilde{i}}^{LS} - \mathbf{s}\right)^H \mathbf{H}_{\tilde{i}}^H \mathbf{H}_{\tilde{i}} \left(\mathbf{s}_{\tilde{i}}^{LS} - \mathbf{s}\right) \leq r^2. \tag{4.15}$$

(4.15) can be modified as $\mathbf{s}_{\tilde{i}}^{SD} = \arg\min_{\mathbf{s}} \|\mathbf{U}_{\tilde{i}} \left(\mathbf{s}_{\tilde{i}}^{LS} - \mathbf{s}\right)\|^2 \leq r^2$ where $\mathbf{U}_{\tilde{i}}$ is obtained from the Cholesky decomposition of $\mathbf{H}_{\tilde{i}}^H \mathbf{H}_{\tilde{i}}$ and it is implemented only for $\tilde{i} = 1, \ldots, N$. This culminates in Algorithm 3 as

Algorithm 3: Reduced Complexity GSIM Detection Algorithm with SD

Data: $\mathbf{y}$, $\mathbf{H}$, N_t, N_{act}, M, r
Initialization: Same as Alg. 1
  $\mathbf{U}_{\tilde{i}}^H \mathbf{U}_{\tilde{i}} = \mathbf{H}_{\tilde{i}}^H \mathbf{H}_{\tilde{i}}$ Cholesky decomp.
for $\tilde{i} \leftarrow 1$ **to** N **do**
  $\mathbf{s}_{\tilde{i}}^{SD} = \arg\min_{\mathbf{s}} \|\mathbf{U}_{\tilde{i}} \left(\mathbf{s}_{\tilde{i}}^{LS} - \mathbf{s}\right)\|^2 \leq r^2$
  $d_{\tilde{i}}^{SD} = \left\|\mathbf{y} - \mathbf{H}_{\tilde{i}}\mathbf{s}_{\tilde{i}}^{SD}\right\|^2$
end
$\hat{i}_{RC}^{SD} = \arg\min d_{\tilde{i}}^{SD}$
$\hat{\mathbf{s}}_{RC}^{SD} = \mathbf{s}_{\hat{i}_{RC}^{SD}}^{SD}$

Complexity of the Algorithm 1 can be calculated using the complexities of $\mathbf{A}_i$, $\mathbf{b}_i$, solving the linear system and $d_{\tilde{i}}$ over N times. Hence, total complexity is

$$C^{GSM} = N_r(N_{act} - 1) + 4N_r(N + 1) \tag{4.16a}$$

$$C^{GSIM} = 2N_{act}\left((N_r + 1)(N_{act} + 1)\right) + \frac{N_{act}^3}{3} + N_r^2 N(2N_{act} + N_r) \tag{4.16b}$$

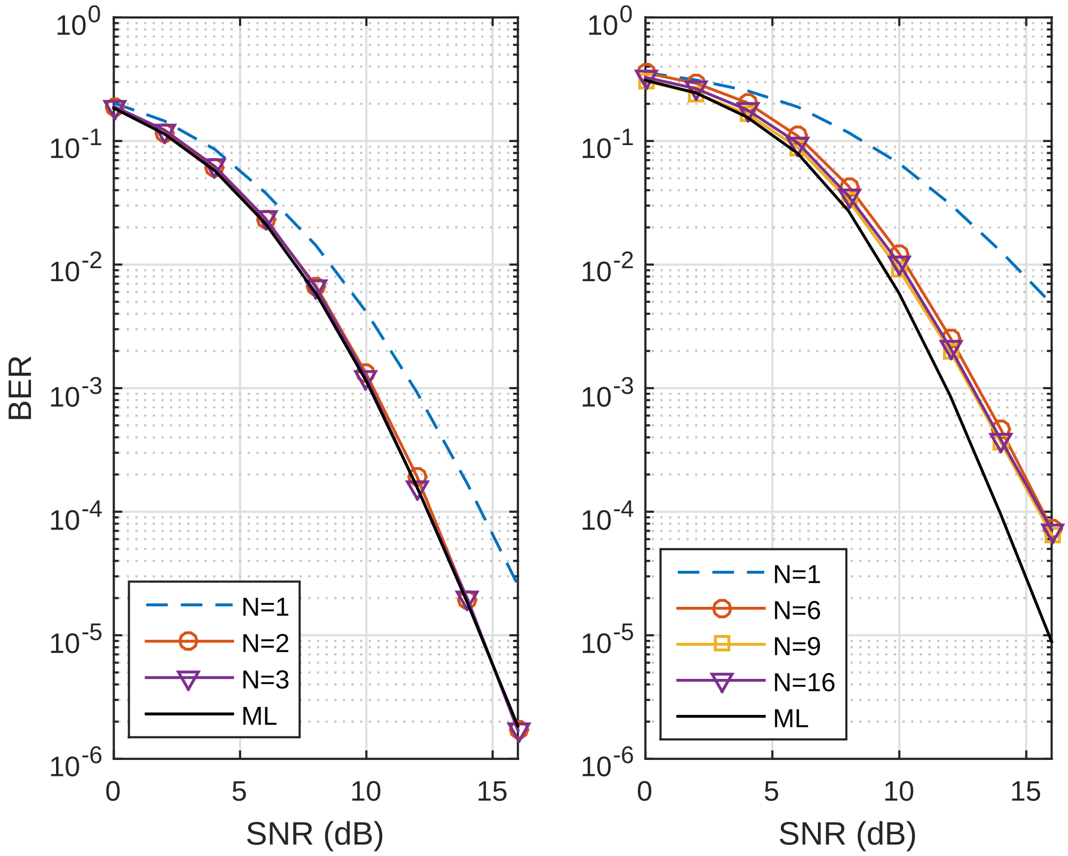

Figure 4.1 : BER performance comparison of GSM (left) and GSIM (right) with ML detection and proposed algorithm ($N_t = 6, N_{act} = 3, N_r = 6$, 4-QAM).

floating point operations (flops). For Algorithm 2 last term in (4.16b) have to be multiplied with $M^{N_{act}}$ and for Algorithm 3, extra N times $C(N_{act}, N_0, r^2)$ complexity which is defined in [54] are needed.

4.3 Performance Evaluation

In this section, computer simulation results for the BER of GSM and GSIM systems are presented. Monte Carlo simulations are performed for at least 10^6 channel uses as a function of the received SNR. Since the sub-optimal solution is based on the LS estimation, the channel matrix have to be an overdetermined system which corresponds that the number of receive antennas have to be grater than the number of activated transmit antennas, namely $N_r > N_{act}$. BER performance comparison of the GSM and GSIM systems with ML detection for Algorithm 1 is given in Figure 4.1. The number of transmit, activated and receive antennas are $N_t = 6, N_{act} = 3, N_r = 6$, respectively and modulation is 4-QAM. N represents the number of sub-optimal solutions. As seen

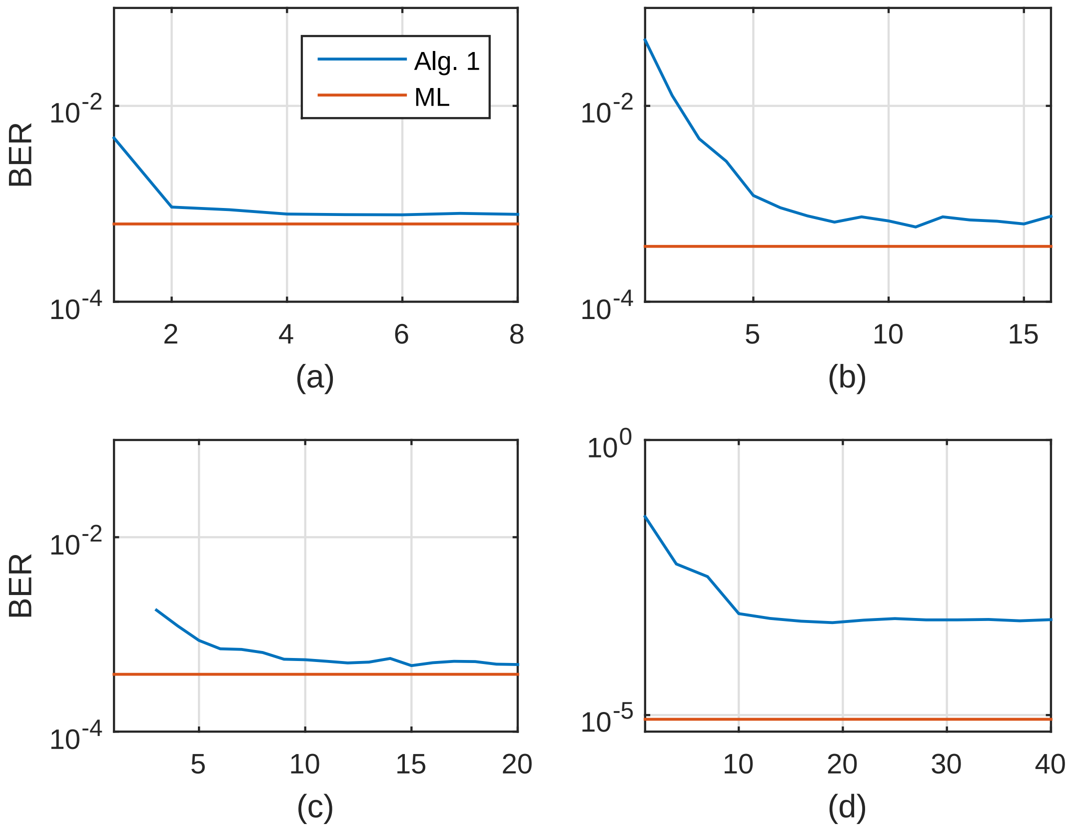

Figure 4.2 : BER vs. number of best estimate, N, for GSIM system (a) $N_t = 5, N_{act} = 2, N_r = 4$, 8-QAM, (b) $N_t = 6, N_{act} = 3, N_r = 6$, BPSK, (c) $N_t = 10, N_{act} = 2, N_r = 4$, QPSK, (d) $N_t = 8, N_{act} = 4, N_r = 6$, QPSK.

from Figure 4.1 (left), although the number of the best estimates are very low ($N = 2, 3$) with respect to the number of antenna combinations, $N_c = 16$, proposed algorithm for GSM system performs near the optimal solution. Furthermore, from Figure 4.1 (right), GSIM also shows satisfactory performance with low N numbers. For GSIM system, BER performance improves with increasing N until the acceptable N value. In order to find this acceptable value, GSIM systems with different configurations are investigated.

BER vs. number of best estimate, N, with different configured GSIM systems can be seen in Figure 4.2. When N is increased, the proposed algorithm becomes closer to ML detection. However, there is a point where performance increase is minimal. This number can be regarded for our algorithm as the acceptable N.

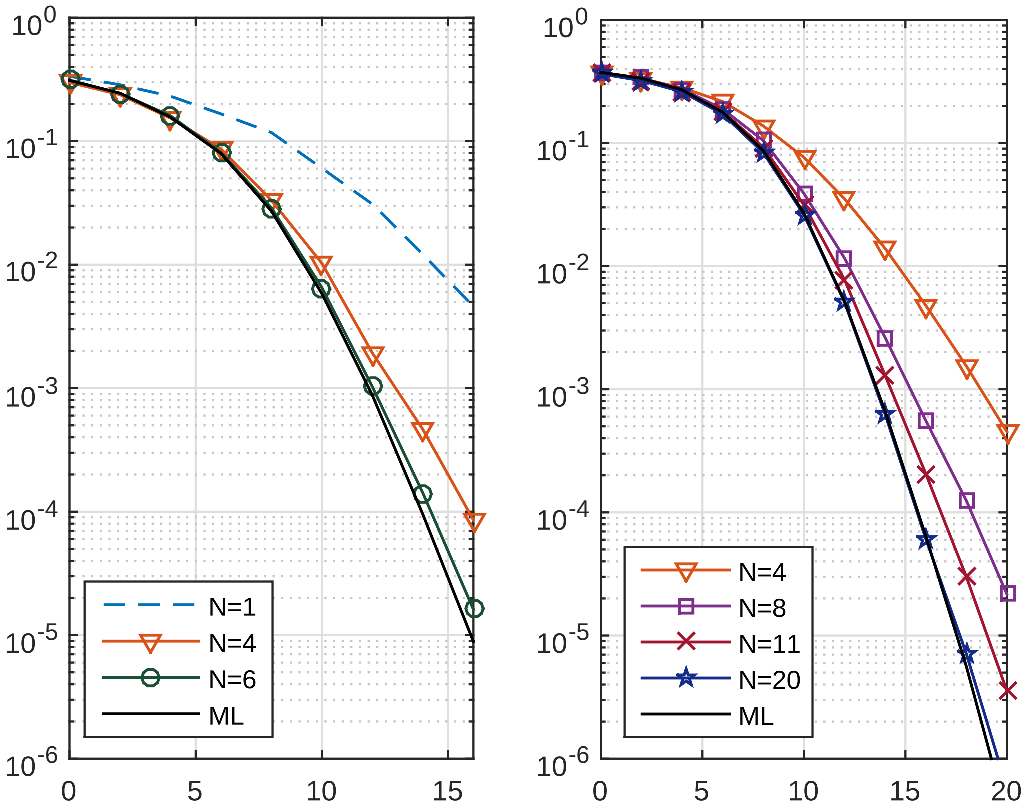

Figure 4.3 : BER performance comparison of reduced complexity algorithm (Alg. 2) with ML. Note that (a) $N_t = 6, N_{act} = 3, N_r = 6$, 4-QAM, (b) $N_t = 7, N_{act} = 4, N_r = 6$, 4-QAM.

Rule of Thumb: As can be seen from the Figure 4.2 and many other simulation results which are not shown here, the number for N can be chosen as $N \approx \frac{Nc}{3}$ for GSIM system.

As stated in **Remark** and can bee seen from Figure 4.2(d), Algorithm 1 can not achieve the near optimal solution especially for high number of activated transmit antennas. Figure 4.3 shows the BER performance of Algorithm 2. As can be seen from Figure 4.3, Algorithm 2 shows near optimal performance. When N is increased, this algorithm approaches to optimum solution. Although, the search space increases with this algorithm, the complexity is still low with respect to ML detection. Figure 4.4 shows the BER comparison of Algorithm 2 and 3. As can be seen from the figure that Algorithm 3 shows the same performance as Algorithm 2 while it reduces the complexity.

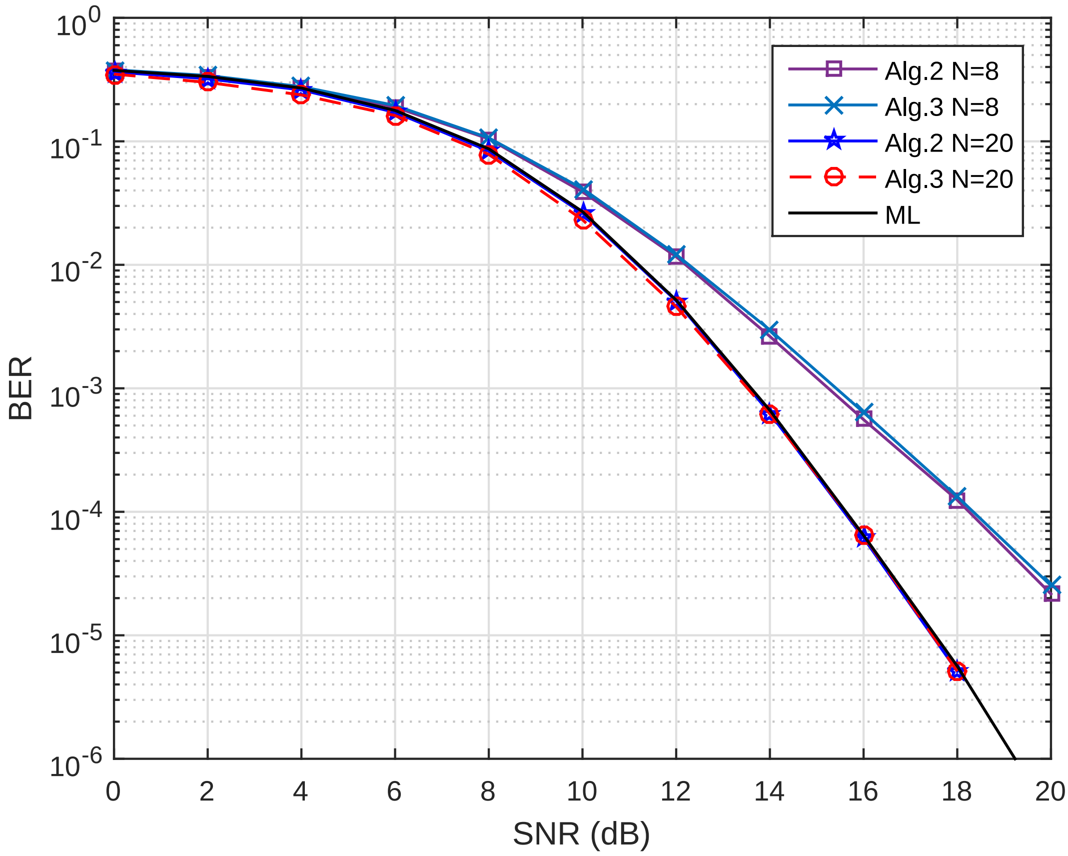

Figure 4.4 : BER performance comparison of Alg. 2 and Alg.3 for $N_t = 7, N_{act} = 4, N_r = 6$, 4-QAM.

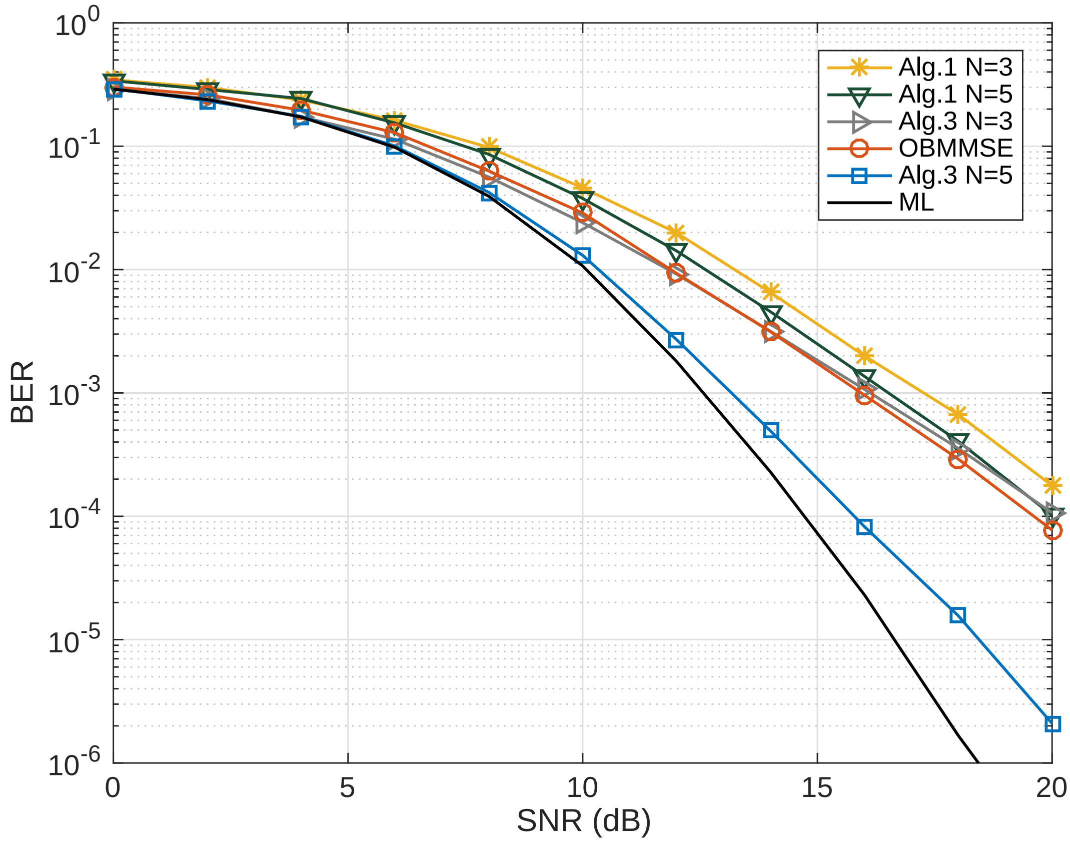

Figure 4.5 : BER performance comparison of proposed algorithms with OB-MMSE. $N_t = 6, N_{act} = 4, N_r = 6$, 4-QAM.

In [39], BER performance comparison to EBCS and OB-MMSE (and other algorithms) are presented where both algorithms show approximately same performance. In Figure 4.5 BER performance comparison of our proposed algorithms (Alg.1 and Alg.3) and OB-MMSE can be seen. Algorithm 1 and Algorithm 3 (with low N number) shows approximately same performance as OB-MMSE. Furthermore, Algorithm 3 with higher N number exhibits better performance than OB-MMSE (and hence EBCS).

5. OUTAGE PROBABILITY ANALYSIS OF COOPERATIVE SM

In this chapter, we extend the results of Chapter 3 to the cooperative scenarios under fixed, selective and incremental relaying techniques. It is shown that cooperative SM systems provide better performance compared to conventional modulation systems in terms of outage probability.

5.1 Outage Probability of Cooperative SM Systems

In a cooperative system, consisting of a source (S), a relay (R) and a destination (D), S and R have N_t^S, and N_t^R transmit antennas, while R and D have N_r^R and N_r^D receive antennas, respectively.

Fading coefficients of each link of the MIMO channels between S and R, $\mathbf{H}^{SR} \in \mathbb{C}^{N_r^R \times N_t^S}$, S and D, $\mathbf{H}^{SD} \in \mathbb{C}^{N_r^D \times N_t^S}$, R and D, $\mathbf{H}^{RD} \in \mathbb{C}^{N_r^D \times N_t^R}$, can be modeled as $\mathscr{CN}(0,1)$ and obey the flat Rayleigh fading channel conditions. In the first time slot, S transmits a unit energy, $E[\mathbf{x}^H\mathbf{x}] = 1$, SM symbol which can be regarded as $\mathbf{x} = [\underbrace{0, \ 0, \ \ldots, \ 0,}_{l-1} \ x_q, \ \underbrace{0, \ \ldots, \ 0}_{N_t-l}]^T$, to D and R as,

$$\mathbf{y}^{SD} = \mathbf{h}_l^{SD} x_q + \mathbf{n}^{SD} \tag{5.1}$$

$$\mathbf{y}^{SR} = \mathbf{h}_l^{SR} x_q + \mathbf{n}^{SR}, \tag{5.2}$$

respectively, where $\mathbf{h}_l^{SD(SR)}$ denotes the l^{th} column vector of the channel matrix $\mathbf{H}^{SD(SR)}$ and x_q is the APM constellation symbol. Based on the DF relaying protocol, R processes the incoming data using the ML detection and makes a decision according to

$$[\tilde{l}, \tilde{x}_q] = \arg\min_{l,q} \left\| \mathbf{y}^{SR} - \mathbf{h}_l^{SR} x_q \right\|^2 \tag{5.3}$$

and re-encodes to an SM signal and sends to D as

$$\mathbf{y}_{DF}^{RD} = \mathbf{h}_{\tilde{l}}^{RD} \tilde{x}_q + \mathbf{n}^{RD}. \tag{5.4}$$

On the other hand, in AF relaying protocol, R just amplifies the received signal and forwards to D as

$$\mathbf{y}_{AF}^{RD} = G\mathbf{H}^{RD}\mathbf{y}^{SR} + \mathbf{n}^{RD} \tag{5.5}$$

where $G = \sqrt{\frac{1}{\left\|\mathbf{h}_l^{SR}\right\|^2 + N_r^R N_0}}$ is the gain factor which is chosen to fix the transmitted energy of the relay to unity. $\mathbf{n}^{SD}, \mathbf{n}^{SR}$ and $\mathbf{n}^{RD}$ are the additive Gaussian noise vectors whose entries are distributed with $\mathscr{CN}(0, N_0)$. After simple manipulations, (5.5) can be written as

$$\begin{aligned} \mathbf{y}_{AF}^{RD} &= G\mathbf{H}^{RD}\left(\mathbf{h}_l^{SR}x_q + \mathbf{n}^{SR}\right) + \mathbf{n}^{RD} \\ &= G\mathbf{H}^{RD}\mathbf{h}_l^{SR}x_q + \tilde{\mathbf{n}}^{RD} \end{aligned} \tag{5.6}$$

where $\tilde{\mathbf{n}}^{RD}$ is the effective noise vector with a covariance matrix as $\mathbf{C} = G^2\mathbf{H}^{RD}\left(\mathbf{H}^{RD}\right)^H N_0 + N_0\mathbf{I}_{N_r^D}$.

The outage probability analysis for classical modulation techniques and different relay and diversity protocols can be found in [9]. In this study, we have used the results for these protocols and applied them to the cooperative SM systems. In the following, the antenna numbers are assumed to be equal, i.e., $N_t^S = N_t^R = N_t$ and $N_r^R = N_r^D = N_r$.

5.1.1 Fixed DF relaying

The mutual information of fixed DF-SM is given as

$$\begin{aligned} I_{DF_{SM}} = \min\Big\{ &\log_2\left(1 + \rho\left\|\mathbf{h}_{\min}^{SR}\right\|^2\right) + H(Y^{SR}) - H(Y^{SR} \mid X_{ch=\min}), \\ &\log_2\left(1 + \rho\left\|\mathbf{h}_{\min}^{SD}\right\|^2 + \left\|\mathbf{h}_{\min}^{RD}\right\|^2\right) + H(Y^{SD+RD}) - H(Y^{SD+RD} \mid X_{ch=\min})\Big\} \end{aligned} \tag{5.7}$$

and the corresponding outage probability is

$$P_{out}^{DF_{SM}} = \Pr\left\{\left\|\mathbf{h}_{\min}^{SR}\right\|^2 \le \mu\right\} + \Pr\left\{\left\|\mathbf{h}_{\min}^{SR}\right\|^2 > \mu\right\}\Pr\left\{\left\|\mathbf{h}_{\min}^{SD}\right\|^2 + \left\|\mathbf{h}_{\min}^{RD}\right\|^2 \le \mu\right\} \tag{5.8}$$

where $H(Y^{SR})$ and $H(Y^{SD+RD})$ can be computed as in (3.8) with variances $\sigma_{SR}^2 = \left\|\mathbf{h}_l^{SR}\right\|^2\sigma_x^2 + N_0$ and $\sigma_{SD+RD}^2 = \left(\left\|\mathbf{h}_l^{SD}\right\|^2 + \left\|\mathbf{h}_l^{RD}\right\|^2\right)\sigma_x^2 + N_0$. On the other hand, $H(Y^{SR} \mid X_{ch=\min})$ and $H(Y^{SD+RD} \mid X_{ch=\min})$ can be computed as in (3.9) with the above variances when $\mathbf{h}_l = \mathbf{h}_{\min}$. The threshold value is $\mu = \frac{1}{\rho}\left(2^{2R - H(Y^{SR/SD+RD}) + H(Y^{SR/SD+RD} \mid X_{ch=\min})} - 1\right)$. As seen from the classical SM systems (see. Figure 3.1), $H(Y) - H(Y \mid X_{ch=\min})$ equals to $\log_2(N_t)$ for mid to high SNR values. For cooperative communications, this value converges to $\log_2(N_t)$ faster than the classical systems. In this work, $H(Y)-H(Y \mid X_{ch=\min})$ is fixed to $\log_2(N_t)$ thus, the threshold value will be $\mu = \dfrac{2^{2R-\log_2(N_t)}-1}{\rho}$. In (5.8), the first term in the right hand side can be calculated from (3.13). The second term is the complement of the first term, i.e., $\Pr\left\{\|\mathbf{h}_{\min}^{SR}\|^2 \leq \mu\right\}$. Finally, the last term can be evaluated using the pdf of sum of two independent random variables which is the convolution of their pdfs. If the pdf of each term is given as in (3.12), the pdf of their sum is calculated as

$$\begin{aligned} f_Z(z) &= \int_0^z f_{\mathscr{H}}(z-x) f_{\mathscr{H}}(x)dx \\ &= \frac{N_t^2 e^{-N_t z}}{(N_r-1)!^2} \sum_{k=0}^{N} \sum_{n=0}^{N} C(k)C(n) z^{2N_r+k+n-1} B(N_r+k, N_r+n) \end{aligned} \tag{5.9}$$

where $B(.,.)$ is the Beta function. The outage probability is then expressed as,

$$\begin{aligned} \Pr\left\{\left\|\mathbf{h}_{\min}^{SD}\right\|^2 + \|\mathbf{h}_{\min}^{RD}\|^2 \leq \mu^{th}\right\} &= \int_0^{\mu^{th}} f_Z(z)dz. \\ &= \sum_{k=0}^{N} \sum_{n=0}^{N} \frac{C(k)C(n)B(N_r+k, N_r+n)}{(N_r-1)!^2 N_t^{2N_r+k+n-2}} \\ &\quad \times \gamma\left(2N_r+k+n, N_t \mu^{th}\right). \end{aligned} \tag{5.10}$$

5.1.2 DF selection relaying

The outage probability for DF-SR is given as

$$\begin{aligned} P_{out}^{DF-SR_{SM}} = \Pr\left\{\left\|\mathbf{h}_{\min}^{SR}\right\|^2 \leq \mu\right\} \Pr\left\{2\left\|\mathbf{h}_{\min}^{SD}\right\|^2 \leq \mu\right\} \\ + \Pr\left\{\left\|\mathbf{h}_{\min}^{SR}\right\|^2 > \mu\right\} \Pr\left\{\left\|\mathbf{h}_{\min}^{SD}\right\|^2 + \|\mathbf{h}_{\min}^{RD}\|^2 \leq \mu\right\} \end{aligned} \tag{5.11}$$

5.1.3 DF incremental relaying

In incremental relaying, if the S-D link SNR is not high enough for the appropriate communication, the relay participates in the communication and transmits the received signal. Otherwise, R remains silent (i.e., there is no second phase). Since the relay does not cooperate in every time slot, the spectral efficiency will be R when relay is silent and $R/2$ otherwise. Therefore, the averaged spectral efficiency can be expressed as

$$\bar{R}_l = R \Pr\left\{\left\|\mathbf{h}_{\min}^{SD}\right\|^2 > \mu'\right\} + \frac{R}{2} \Pr\left\{\left\|\mathbf{h}_{\min}^{SD}\right\|^2 \leq \mu'\right\} \tag{5.12}$$

where $\mu' = \frac{2^{R-log_2(N_t)-1}}{\rho}$. (5.12) can be evaluated using (3.13). The outage probability of DF-IR is written as,

$$\begin{aligned} P_{out}^{DF-IR_{SM}} &= \Pr\left\{ \left\| \mathbf{h}_{\min}^{SR} \right\|^2 \leq \bar{\mu} \right\} \Pr\left\{ \left\| \mathbf{h}_{\min}^{SD} \right\|^2 \leq \bar{\mu} \right\} \\ &\quad + \Pr\left\{ \left\| \mathbf{h}_{\min}^{SR} \right\|^2 > \bar{\mu} \right\} \Pr\left\{ \left\| \mathbf{h}_{\min}^{SD} \right\|^2 + \left\| \mathbf{h}_{\min}^{RD} \right\|^2 \leq \bar{\mu} \right\} \end{aligned} \tag{5.13}$$

where $\bar{\mu} = \frac{2^{\bar{R}-log_2(N_t)-1}}{\rho}$.

5.1.4 Fixed AF relaying

The outage probability for fixed AF relaying can be written as a function of fading coefficients as (The analysis for AF relaying is only given for $N_r = 1$)

$$P_{out}^{AF_{SM}} = \Pr\left\{ \left| h_{\min}^{SD} \right|^2 + \frac{1}{\rho} f\left(\rho \left| h_{\min}^{SR} \right|^2, \rho \left| h_{\min}^{RD} \right|^2 \right) \leq \mu \right\} \tag{5.14}$$

where $f(x,y) = \frac{xy}{x+y+1}$ [9]. In order to find the pdf of the left hand side of the inequality, we can employ the upper bound approach as performed in [55]. The expressed function can be upper bounded as,

$$f(x,y) = \frac{xy}{x+y+1} \leq \min(x,y). \tag{5.15}$$

Hence, the outage probability is re-expressed as

$$P_{out}^{AF_{SM}} \approx \Pr\left\{ \left| h_{\min}^{SD} \right|^2 + \min\left(\left| h_{\min}^{SR} \right|^2, \left| h_{\min}^{RD} \right|^2 \right) \leq \mu \right\}. \tag{5.16}$$

For $N_r = 1$, each $\left\{ |h_l|^2 \right\}_{l=1}^{N_t}$ is exponentially distributed. Hence $\xi \triangleq |h_{\min}|^2$ is also exponentially distributed with the pdf $f_\Xi(\xi) = N_t e^{-N_t \xi}$ and the cdf $F_\Xi(\xi) = 1 - e^{-N_t \xi}$. Therefore the pdf of $\omega = a + \min(b,c)$ is $f_\Omega(\omega) = 2N_t \left(e^{-N_t \omega} - e^{-2N_t \omega} \right)$. Finally, the outage probability of AF relaying SM is given as

$$P_{out}^{AF} \approx \int_0^\mu f_\Omega(\omega) dz = 1 - 2e^{-N_t \mu} + e^{-2N_t \mu}. \tag{5.17}$$

5.1.5 AF incremental relaying

The outage probability of AF-IR can be written as,

$$\begin{aligned} P_{out}^{AF-IR_{SM}} &= \Pr\left\{ \left| h_{\min}^{SD} \right|^2 \leq \bar{\mu} \right\} \Pr\left\{ \left| h_{\min}^{SD} \right|^2 + \min\left(\left| h_{\min}^{SR} \right|^2, \left| h_{\min}^{RD} \right|^2 \right) \leq \bar{\mu} \,\middle|\, \left| h_{\min}^{SD} \right|^2 \leq \bar{\mu} \right\} \\ &= \Pr\left\{ \left| h_{\min}^{SD} \right|^2 + \min\left(\left| h_{\min}^{SR} \right|^2, \left| h_{\min}^{RD} \right|^2 \right) \leq \bar{\mu} \right\}. \end{aligned} \tag{5.18}$$

The analytical expression for (5.18) is the same as in (5.17), but the only difference is μ replaced by $\bar{\mu}$.

5.2 Performance Evaluation

In this section, we present computer simulation results for the outage probability of cooperative SM systems. Monte Carlo simulations are realized for at least 10^7 channel uses as a function of received SNR (ρ) and obtained curves are compared to the analytical results. All cooperative SM systems are compared with the conventional modulation techniques on an equal spectral efficiency basis.

The outage probability performance of fixed DF, DF-SR, DF-IR and fixed AF, AF-IR of an $N_t = 4$, QPSK SM system is presented in Figure 5.1. As seen from Figure 5.1, the simulation results and the analytical curves have exact match. In [9] and [55], it is stated that the AF relaying is superior to DF relaying in terms of outage probability. This result is observed in here as well. The performance of cooperative SM and the corresponding PSK systems is presented in Figure 5.2 for $R = 4$ bits/s/Hz. As seen from Figure 5.2, the cooperative SM system provides better performance than corresponding conventional modulated systems for both DF and AF protocols.

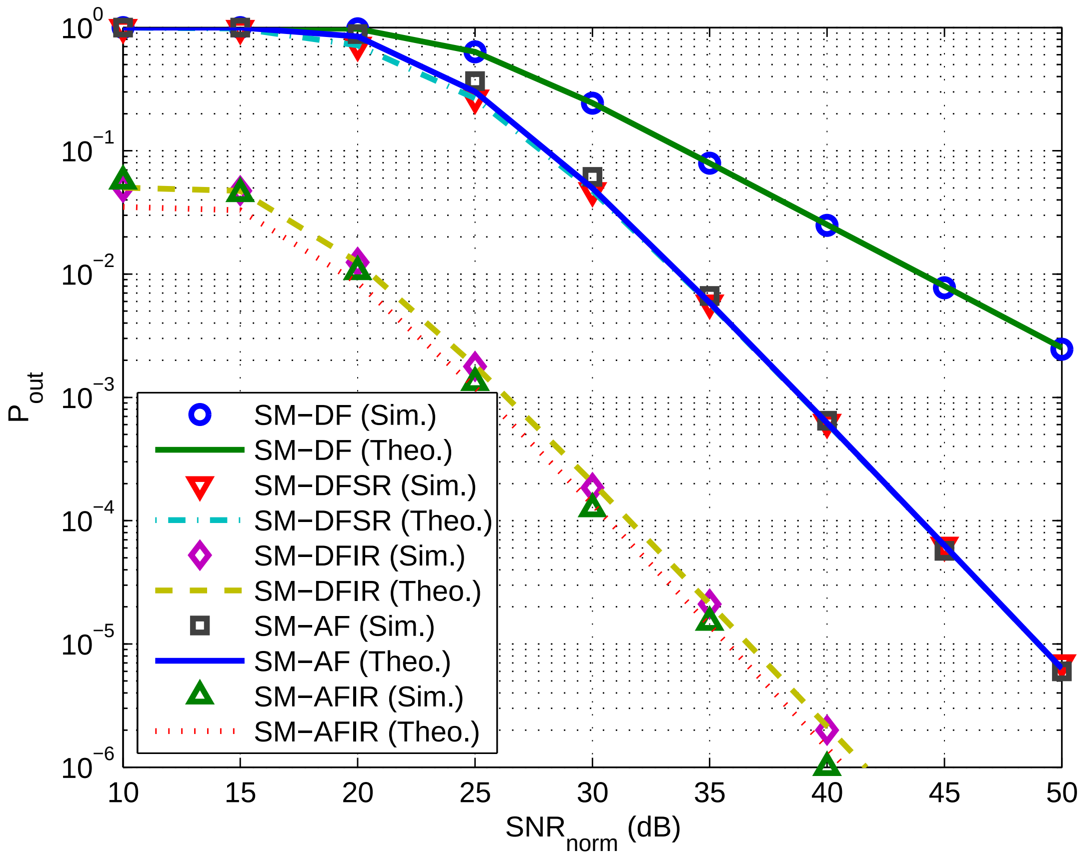

Figure 5.1 : Outage probability performance of $N_t = 4$ QPSK AF vs. DF Cooperative SM system ($N_r = 1$).

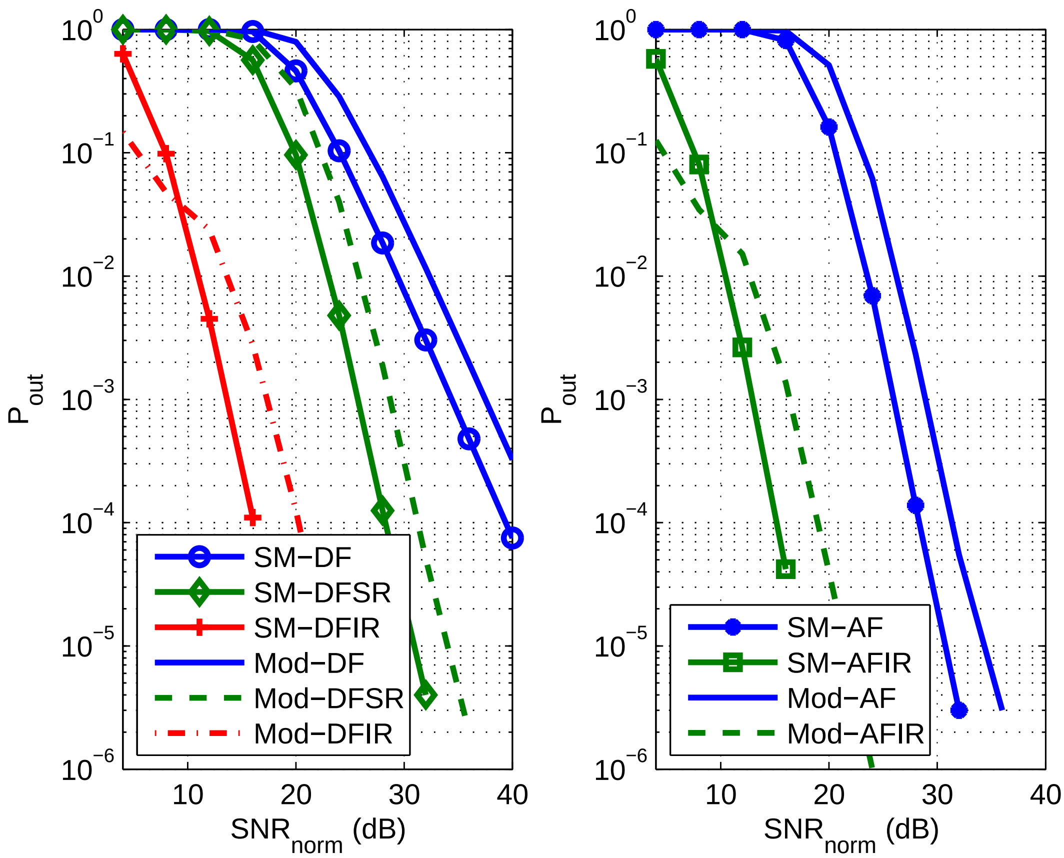

Figure 5.2 : Outage probability performance of $N_t = 4$ QPSK Cooperative SM system with respect to SIMO 16-PSK ($N_r = 2$).

6. BER ANALYSIS OF MIMO COOPERATIVE SM SYSTEMS

In this chapter, we propose novel cooperative SM systems with two main relaying techniques (amplify-and-forward, AF, and decode-and-forward, DF) where all nodes have multiple transmit and/or receive antennas. Most of the studies in the literature of cooperative SM systems, which combine the advantages of cooperative communications and SM systems, consider only the SSK scheme with single receive/transmit antenna at relay and destination. Since the error performance of SM highly depends on the number of receive antennas and more flexible cooperative communications systems can be obtained by using SM with multiple antennas, it is essential to investigate multi-antenna cooperative SM systems. We derive analytical expressions of the average bit error probability (ABEP) for both the newly proposed cooperative SM-DF and cooperative SM-AF systems and validate with the computer simulation results. Furthermore, we present the BER comparison of considered systems with classical M-PSK/QAM cooperative systems. Computer simulation results indicate that multiple antennas cooperative SM systems provide better error performance than classical cooperative systems for both relaying techniques.

6.1 System Model

The considered cooperative communications system for SM MIMO-DF and MIMO-AF consisting of a single relay is given in Figs. 6.1 and 6.2. In these systems, S and R have N_t^S and N_t^R transmit antennas while R and D have N_r^R and N_r^D receive antennas, respectively. The channel matrices composed of channel fading coefficients between S-R, S-D and R-D can be given as $\mathbf{H}^{SR} \in \mathbb{C}^{N_r^R \times N_t^S}$, $\mathbf{H}^{SD} \in \mathbb{C}^{N_r^D \times N_t^S}$, and $\mathbf{H}^{RD} \in \mathbb{C}^{N_r^D \times N_t^R}$, respectively. Each entry of the above matrices is modeled as an independent and identically distributed rv. with $\mathcal{CN}(0, \sigma_h^2)$ and the channel obeys the Rayleigh flat fading model, where σ_h^2 is equal to σ_{SR}^2, σ_{SD}^2 and σ_{RD}^2 for the corresponding three channel matrices, respectively. To take into account the path loss, the variances are defined as $\sigma_{SR}^2 \triangleq d_{SR}^{-\alpha}$, $\sigma_{SD}^2 \triangleq d_{SD}^{-\alpha}$ and $\sigma_{RD}^2 \triangleq d_{RD}^{-\alpha}$ where d_{SR}, d_{SD}

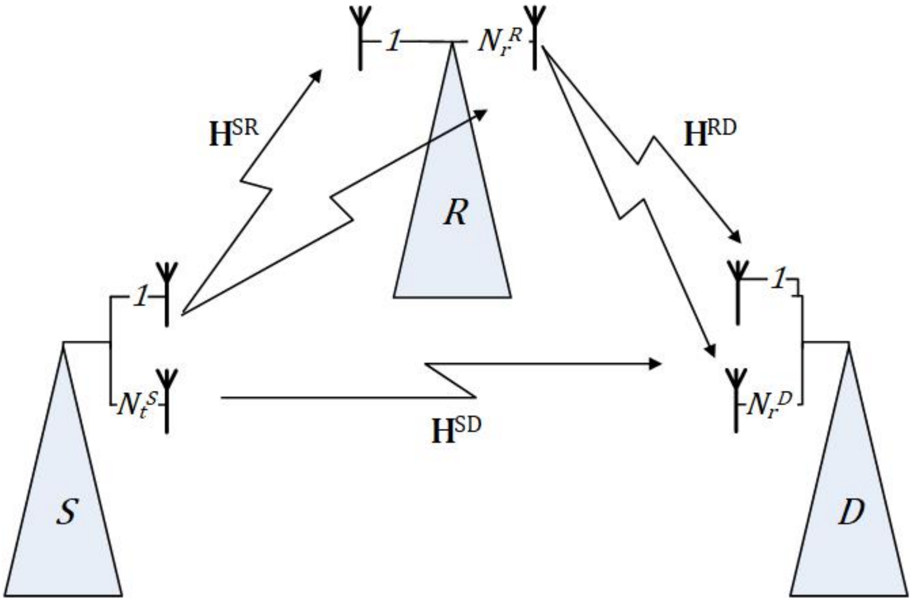

Figure 6.1 : Cooperative communications scenario with SM MIMO-DF.

and d_{RD} are the distances between S-R, S-D and R-D, respectively and α is the path loss exponent [9]. SNR parameter is defined as received SNR at D.

An SM symbol is given as $\mathbf{x} = [\underbrace{0 \ \ 0 \ \ \cdots \ \ 0}_{l-1} \ x_q \ \underbrace{0 \ \ \cdots \ \ 0}_{N_t^S - l}]^T$, where l is the index of the activated transmit antenna, x_q is the M-PSK/QAM constellation symbol and it is assumed that $E\left\{\mathbf{x}^H\mathbf{x}\right\} = 1$. In the first time slot, S transmits an SM symbol to R and D as

$$\mathbf{y}^{SR} = \mathbf{H}^{SR}\mathbf{x} + \mathbf{n}^{SR} \tag{6.1}$$

$$\mathbf{y}^{SD} = \mathbf{H}^{SD}\mathbf{x} + \mathbf{n}^{SD} \tag{6.2}$$

respectively, where $\mathbf{n}^{SR(SD)} \in \mathbb{C}^{N_r^R(N_r^D)\times 1}$ is the AWGN vector with $\mathscr{CN}(0, N_0)$.

6.1.1 Decode-and-forward cooperative SM

In the second time slot, the detector at R, which has the ideal channel state information (CSI), detects the SM symbol applying the ML decision rule as

$$(\tilde{l}, \tilde{x}_q) = \arg\min_{l, x_q} \left\| \mathbf{y}^{SR} - \mathbf{H}^{SR}\mathbf{x} \right\|^2 \tag{6.3}$$

and re-encodes into a new SM symbol by considering $\tilde{l}$ and $\tilde{x}_q$ as $\tilde{\mathbf{x}} = [\underbrace{0 \ \ 0 \ \ \cdots \ \ 0}_{\tilde{l}-1} \ \tilde{x}_q \ \underbrace{0 \ \ \cdots \ \ 0}_{N_t^R - \tilde{l}}]^T$ and sends it to D, which is received as

$$\mathbf{y}^{RD} = \mathbf{H}^{RD}\tilde{\mathbf{x}} + \mathbf{n}^{RD} \tag{6.4}$$

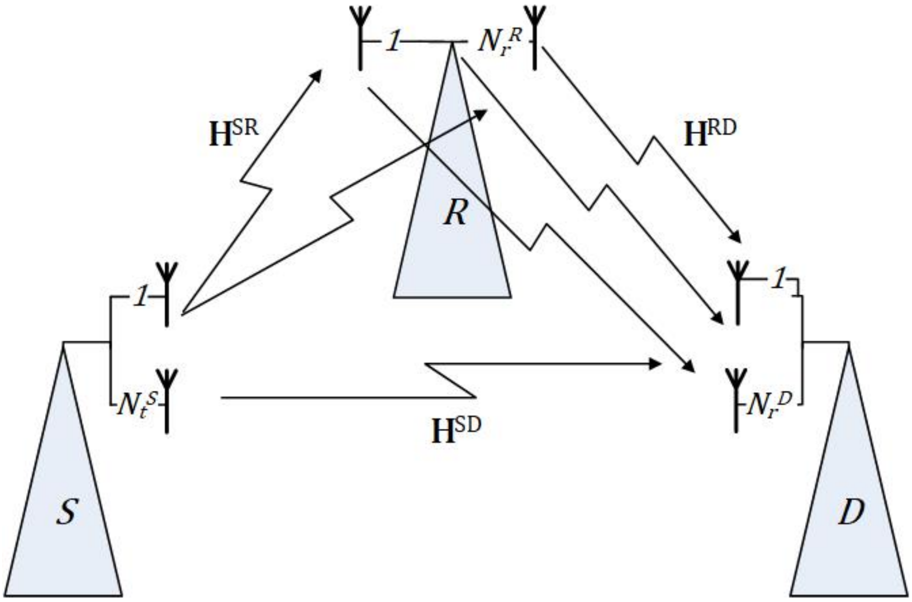

Figure 6.2 : MIMO-AF configuration ($N_t^R = N_r^R = N^R$).

where $\mathbf{n}^{RD} \in \mathbb{C}^{N_r^D \times 1}$ is the AWGN vector with $\mathscr{CN}(0, N_0)$. From (6.2) and (6.4), the ML detection rule at D is given by

$$(\hat{l}, \hat{x}_q) = \arg\min_{l, x_q} \left(\left\| \mathbf{y}^{SD} - \mathbf{H}^{SD}\mathbf{x} \right\|^2 + \left\| \mathbf{y}^{RD} - \mathbf{H}^{RD}\mathbf{x} \right\|^2 \right). \tag{6.5}$$

6.1.2 Amplify-and-forward cooperative SM

In MIMO-AF system, R amplifies the received signal at each receive antenna and sends to D from the same antenna as seen from Figure 6.2 ($N_t^R = N_r^R = N^R$). The received signal vector at D becomes

$$\begin{aligned} \mathbf{y}^{RD} &= G\mathbf{H}^{RD}\mathbf{y}^{SR} + \mathbf{n}^{RD} \\ &= G\mathbf{H}^{RD}\mathbf{H}^{SR}\mathbf{x} + \mathbf{n}_{\text{MIMO}} \end{aligned} \tag{6.6}$$

where $G = \sqrt{\frac{1}{N_t^R(\sigma_{SR}^2 + N_0)}}$ (in G, $\sqrt{\frac{1}{N_t^R}}$ is the scaling factor for the normalization of the transmitted energy at R), $\mathbf{n}_{\text{MIMO}}$ is the colored Gaussian noise vector and can be expressed as

$$\mathbf{n}_{\text{MIMO}} = G\mathbf{H}^{RD}\mathbf{n}^{SR} + \mathbf{n}^{RD} \tag{6.7}$$

with the conditional covariance matrix

$$\begin{aligned} \mathbf{C} &= E\left\{ \mathbf{n}_{\text{MIMO}}\mathbf{n}_{\text{MIMO}}^H \mid \mathbf{H}^{RD} \right\} \\ &= G^2\mathbf{H}^{RD}\left(\mathbf{H}^{RD}\right)^H N_0 + N_0\mathbf{I}_{N_r^D}. \end{aligned} \tag{6.8}$$

ML detection rule at D is given for the MIMO-AF system as [56]

$$(\hat{l},\hat{x}_q)=\arg\min_{l,q}\left(\left\|\mathbf{y}^{SD}-\mathbf{H}^{SD}\mathbf{x}\right\|^2/N_0+\left\|\mathbf{C}^{-1/2}\left(\mathbf{y}^{RD}-G\mathbf{H}^{RD}\mathbf{H}^{SR}\mathbf{x}\right)\right\|^2\right). \tag{6.9}$$

6.2 Average Bit Error Probability Analysis For DF Relaying

The ABEP of the cooperative SM system can be evaluated using the average pairwise error probability (APEP) which is computed next.

APEP at D for the MIMO-DF cooperative SM system is upper bounded as

$$P_D^{DF}(\mathbf{x}\to\hat{\mathbf{x}})\le P_R^c(\mathbf{x})P_D(\mathbf{x}\to\hat{\mathbf{x}}\mid R:\mathbf{x})+\sum_{\substack{\tilde{\mathbf{x}}\\ \tilde{\mathbf{x}}\neq\mathbf{x}}}P_R(\mathbf{x}\to\tilde{\mathbf{x}})P_D(\mathbf{x}\to\hat{\mathbf{x}}\mid R:\tilde{\mathbf{x}}) \tag{6.10}$$

where $P_R(\mathbf{x}\to\tilde{\mathbf{x}})$ is the APEP at R when $\mathbf{x}$ is transmitted and it is erroneously detected as $\tilde{\mathbf{x}}$, $P_R^c(\mathbf{x})$ is the probability of correct decision at R, $P_D(\mathbf{x}\to\hat{\mathbf{x}}\mid R:\mathbf{x})$ is the APEP at D when R detects the SM symbol correctly and $P_D(\mathbf{x}\to\hat{\mathbf{x}}\mid R:\tilde{\mathbf{x}})$ is the APEP at D when R makes a decoding error.

$P_D(\mathbf{x}\to\hat{\mathbf{x}}\mid R:\tilde{\mathbf{x}})$ can be calculated for all possible $\tilde{\mathbf{x}}$ as

$$\begin{aligned}P_D(\mathbf{x}\to\hat{\mathbf{x}}\mid R:\tilde{\mathbf{x}})=E\Bigg\{\Pr\Bigg\{&\left\|\mathbf{y}^{SD}-\mathbf{H}^{SD}\mathbf{x}\right\|^2+\left\|\mathbf{y}^{RD}-\mathbf{H}^{RD}\mathbf{x}\right\|^2\\ &\ge\left\|\mathbf{y}^{SD}-\mathbf{H}^{SD}\hat{\mathbf{x}}\right\|^2+\left\|\mathbf{y}^{RD}-\mathbf{H}^{RD}\hat{\mathbf{x}}\right\|^2\mid\mathbf{H}^{SD},\mathbf{H}^{RD}\Bigg\}\Bigg\}\end{aligned} \tag{6.11}$$

$$\begin{aligned}=E\Bigg\{\Pr\Bigg\{&\left\|\left(\mathbf{H}^{SD}\mathbf{x}+\mathbf{n}^{SD}\right)-\mathbf{H}^{SD}\mathbf{x}\right\|^2+\left\|\left(\mathbf{H}^{RD}\tilde{\mathbf{x}}+\mathbf{n}^{RD}\right)-\mathbf{H}^{RD}\mathbf{x}\right\|^2\\ &\ge\left\|\left(\mathbf{H}^{SD}\mathbf{x}+\mathbf{n}^{SD}\right)-\mathbf{H}^{SD}\hat{\mathbf{x}}\right\|^2\\ &\quad+\left\|\left(\mathbf{H}^{RD}\tilde{\mathbf{x}}+\mathbf{n}^{RD}\right)-\mathbf{H}^{RD}\hat{\mathbf{x}}\right\|^2\mid\mathbf{H}^{SD},\mathbf{H}^{RD}\Bigg\}\Bigg\}\end{aligned} \tag{6.12}$$

which simplifies to

$$\begin{aligned}&P_D(\mathbf{x}\to\hat{\mathbf{x}}\mid R:\tilde{\mathbf{x}})\\ &=E\left\{Q\left(\frac{\left\|\mathbf{H}^{SD}(\mathbf{x}-\hat{\mathbf{x}})\right\|^2+\left\|\mathbf{H}^{RD}(\tilde{\mathbf{x}}-\hat{\mathbf{x}})\right\|^2-\left\|\mathbf{H}^{RD}(\tilde{\mathbf{x}}-\mathbf{x})\right\|^2}{\sqrt{2N_0\left(\left\|\mathbf{H}^{SD}(\mathbf{x}-\hat{\mathbf{x}})\right\|^2+\left\|\mathbf{H}^{RD}(\tilde{\mathbf{x}}-\hat{\mathbf{x}})\right\|^2+\left\|\mathbf{H}^{RD}(\tilde{\mathbf{x}}-\mathbf{x})\right\|^2\right)}}\right)\right\}.\end{aligned} \tag{6.13}$$

When the transmitted SM symbol $\mathbf{x}$ is erroneously detected as

$$\hat{\mathbf{x}}=[\,\underbrace{0\ \ 0\ \ \dots\ \ 0}_{\hat{l}-1}\ \hat{x}_q\ \underbrace{0\ \ \dots\ \ 0}_{N_t-\hat{l}}\,]^T$$

at both R and D, i.e., for the case of $\tilde{\mathbf{x}} = \hat{\mathbf{x}}$, (6.13) can be written as

$$P_D(\mathbf{x} \to \hat{\mathbf{x}} \mid R : \hat{\mathbf{x}}) = E\left\{ Q\left(\frac{\left\|\mathbf{H}^{SD}(\mathbf{x}-\hat{\mathbf{x}})\right\|^2 - \left\|\mathbf{H}^{RD}(\hat{\mathbf{x}}-\mathbf{x})\right\|^2}{\sqrt{2N_0\left(\left\|\mathbf{H}^{SD}(\mathbf{x}-\hat{\mathbf{x}})\right\|^2 + \left\|\mathbf{H}^{RD}(\hat{\mathbf{x}}-\mathbf{x})\right\|^2\right)}} \right) \right\}. \quad (6.14)$$

Since the Q-function strictly decreases, (6.14) is greater than (6.13). Moreover, at high SNR values ($N_0 \to 0$), (6.14) can be approximated as [57]

$$P_D(\mathbf{x} \to \hat{\mathbf{x}} \mid R : \hat{\mathbf{x}}) \approx \Pr\left(\left\|\mathbf{H}^{SD}(\mathbf{x}-\hat{\mathbf{x}})\right\|^2 < \left\|\mathbf{H}^{RD}(\mathbf{x}-\hat{\mathbf{x}})\right\|^2 \mid \mathbf{H}^{SD}, \mathbf{H}^{RD} \right). \quad (6.15)$$

Since the right and left hand sides of the inequality in (6.15) follow the same distribution, this probability equals 0.5. This implies that ABEP of MIMO-DF cooperative SM system is dominated by the case of $\tilde{\mathbf{x}} = \hat{\mathbf{x}}$. Therefore, (6.10) can be approximated as

$$P_D^{DF}(\mathbf{x} \to \hat{\mathbf{x}}) \approx P_R^c(\mathbf{x}) P_D(\mathbf{x} \to \hat{\mathbf{x}} \mid R : \mathbf{x}) + P_R(\mathbf{x} \to \hat{\mathbf{x}}) P_D(\mathbf{x} \to \hat{\mathbf{x}} \mid R : \hat{\mathbf{x}}). \quad (6.16)$$

$P_R(\mathbf{x} \to \hat{\mathbf{x}})$ is the APEP of the conventional SM and can be formulated as

$$\begin{aligned} P_R(\mathbf{x} \to \hat{\mathbf{x}}) &= E\left\{ \Pr\left\{ \mathbf{x} \to \hat{\mathbf{x}} \mid \mathbf{H}^{SR} \right\} \right\} \\ &= E\left\{ Q\left(\sqrt{\frac{\left\|\mathbf{H}^{SR}(\mathbf{x}-\hat{\mathbf{x}})\right\|^2}{2N_0}} \right) \right\}. \end{aligned} \quad (6.17)$$

$P_R^c(\mathbf{x})$ is also dominated by the case of $\tilde{\mathbf{x}} = \hat{\mathbf{x}}$ and can be approximated as $P_R^c(\mathbf{x}) \approx (1 - P_R(\mathbf{x} \to \hat{\mathbf{x}}))$.

Since $P_D(\mathbf{x} \to \hat{\mathbf{x}} \mid R : \mathbf{x})$ is the APEP at D when R detects the SM symbol correctly, the antenna index and the data symbol are detected properly, i.e., $\tilde{l} = l$ and $\tilde{x}_q = x_q$. For this case, by substituting $\tilde{\mathbf{x}} = \mathbf{x}$ in (6.13), $P_D(\mathbf{x} \to \hat{\mathbf{x}} \mid R : \mathbf{x})$ simplifies to

$$P_D(\mathbf{x} \to \hat{\mathbf{x}} \mid R : \mathbf{x}) = E\left\{ Q\left(\sqrt{\frac{\left\|\mathbf{H}^{SD}(\mathbf{x}-\hat{\mathbf{x}})\right\|^2 + \left\|\mathbf{H}^{RD}(\mathbf{x}-\hat{\mathbf{x}})\right\|^2}{2N_0}} \right) \right\}. \quad (6.18)$$

To obtain the APEP, the pdf.s of the rv.s in Q-function of (6.17) and (6.18) have to be computed. Let $\gamma^{SR} \triangleq \frac{\rho}{2}\left\|\mathbf{H}^{SR}(\mathbf{x}-\hat{\mathbf{x}})\right\|^2$ with the pdf $f_{\gamma^{SR}}(\gamma), \quad \gamma^{SR} \geq 0$ where $\rho = \frac{1}{N_0}$. The APEP at R can be calculated as

$$P_R(\mathbf{x} \to \hat{\mathbf{x}}) = \int_0^\infty Q\left(\sqrt{\gamma^{SR}}\right) f_{\gamma^{SR}}(\gamma) d\gamma \quad (6.19)$$

which can be computed with Craig's formula [58]. Using the alternative form of the Q-function in (6.19) yields

$$P_R(\mathbf{x} \to \hat{\mathbf{x}}) = \frac{1}{\pi} \int_0^{\frac{\pi}{2}} M_{\gamma^{SR}} \left(-\frac{1}{2\sin^2\theta} \right) d\theta \tag{6.20}$$

where $M_{\gamma^{SR}}(s)$ is the MGF of γ^{SR}. For the Rayleigh flat fading channel model, γ^{SR} follows the gamma distribution (in this special case, in fact, it follows Erlang distribution) with *Gamma* $\left(N_r^R, \frac{\rho\lambda_x\sigma_{SR}^2}{2}\right)$ where λ_x is the single eigenvalue of $(\mathbf{x}-\hat{\mathbf{x}})(\mathbf{x}-\hat{\mathbf{x}})^H$ and is equal to

$$\lambda_x = \begin{cases} \left|x_q - \hat{x}_q\right|^2 & \text{if } l = \hat{l} \\ \left|x_q\right|^2 + \left|\hat{x}_q\right|^2 & \text{if } l \neq \hat{l}. \end{cases} \tag{6.21}$$

The MGF of γ^{SR} is obtained as [58]

$$M_{\gamma^{SR}}(s) = \left(1 - \frac{\rho\lambda_x\sigma_{SR}^2}{2}s\right)^{-N_r^R}. \tag{6.22}$$

(6.20) can be computed using Eq. (5A.4a) of [58] as

$$P_R(\mathbf{x} \to \hat{\mathbf{x}}) = \frac{1}{2}\left[1 - \mu \sum\nolimits_{j=0}^{N_r^R-1} \binom{2j}{j} \left(\frac{1-\mu^2}{4}\right)^j\right] \tag{6.23}$$

where $\mu = \sqrt{\frac{\rho\lambda_x\sigma_{SR}^2/4}{(\rho\lambda_x\sigma_{SR}^2/4)+1}}$. The same procedures can also be followed for the calculation of $P_D(\mathbf{x} \to \hat{\mathbf{x}} \mid R : \mathbf{x})$.

Let us consider

$$\begin{aligned} \gamma^{DF} &= \frac{\rho}{2}\left(\left\|\mathbf{H}^{SD}(\mathbf{x}-\hat{\mathbf{x}})\right\|^2 + \left\|\mathbf{H}^{RD}(\mathbf{x}-\hat{\mathbf{x}})\right\|^2\right) \\ &= \gamma^{SD} + \gamma^{RD}. \end{aligned} \tag{6.24}$$

(6.24) is the sum of two Gamma rv.s with different scale parameters. The MGF of (6.24) can be written as

$$M_{\gamma^{DF}}(s) = \left(1 - \frac{\rho\lambda_x\sigma_{SD}^2}{2}s\right)^{-N_r^D} \left(1 - \frac{\rho\lambda_x\sigma_{RD}^2}{2}s\right)^{-N_r^D}. \tag{6.25}$$

Therefore, $P_D(\mathbf{x} \to \hat{\mathbf{x}} \mid R : \mathbf{x})$ can be written as

$$P_D(\mathbf{x} \to \hat{\mathbf{x}} \mid R : \mathbf{x}) = \frac{1}{\pi} \int_0^{\frac{\pi}{2}} M_{\gamma^{DF}} \left(-\frac{1}{2\sin^2\theta} \right) d\theta \tag{6.26}$$

which can be calculated with the help of Appendix 5A.58 of [58].

On the other hand, the pdf of γ^{DF} follows the distribution of the sum of two gamma rv.s as $Gamma\left(2N_r^D, \frac{\rho\lambda_x\sigma^2}{2}\right)$ when $\sigma_{SD}^2 = \sigma_{RD}^2 = \sigma^2$. Therefore, its MGF is given as

$$M_{\gamma^{DF}}(s) = \left(1 - \frac{\rho\lambda_x\sigma^2}{2}s\right)^{-2N_r^D}. \tag{6.27}$$

As a result, $P_D(\mathbf{x} \to \hat{\mathbf{x}} \mid R : \mathbf{x})$ is obtained as

$$P_D(\mathbf{x} \to \hat{\mathbf{x}} \mid R : \mathbf{x}) = \frac{1}{2}\left[1 - \mu \sum_{j=0}^{2N_r^D-1} \binom{2j}{j} \left(\frac{1-\mu^2}{4}\right)^j\right] \tag{6.28}$$

where μ is as defined in (6.23).

After computing the APEP, the ABEP can be averaged as [11]

$$P_b^{DF} \approx \frac{1}{N_t M \log_2(N_t M)} \sum_{\mathbf{x}} \sum_{\substack{\hat{\mathbf{x}} \\ \hat{\mathbf{x}} \neq \mathbf{x}}} n(\mathbf{x} \to \hat{\mathbf{x}}) P_D^{DF}(\mathbf{x} \to \hat{\mathbf{x}}) \tag{6.29}$$

where $n(\mathbf{x} \to \hat{\mathbf{x}})$ is the number of bit errors between the SM symbols $\mathbf{x}$ and $\hat{\mathbf{x}}$.

An upper limit can be obtained when $\theta = \pi/2$ is used in the integrand function of (6.20) and (6.26). Therefore, $P_R(\mathbf{x} \to \hat{\mathbf{x}})$ and $P_D(\mathbf{x} \to \hat{\mathbf{x}} \mid R : \mathbf{x})$ are upper bounded as

$$P_R(\mathbf{x} \to \hat{\mathbf{x}}) \leq \left(1 + \frac{\rho\lambda_x\sigma_{SR}^2}{4}\right)^{-N_r^R} \tag{6.30}$$

$$P_D(\mathbf{x} \to \hat{\mathbf{x}} \mid R : \mathbf{x}) \leq \left(1 + \frac{\rho\lambda_x\sigma_{SD}^2}{4}\right)^{-N_r^D} \left(1 + \frac{\rho\lambda_x\sigma_{RD}^2}{4}\right)^{-N_r^D}. \tag{6.31}$$

We can easily see from (6.16), (6.30) and (6.31) that when SNR $\to \infty$, i.e., $\frac{\rho\lambda_x\sigma_{SR,RD,SD}^2}{4} \to \infty$, the diversity order can be achieved as $\min\left\{N_r^R, 2N_r^D\right\}$. Note that, when $\sigma_{SR}^2 = d_{SR}^{-\alpha} \gg \sigma_{SD}^2, \sigma_{RD}^2$, i.e., R close to S, R detects the signal correctly with a high probability and (6.31) dominates compared to (6.30). Therefore, the diversity order will be $2N_r^D$.

6.3 Average Bit Error Probability Analysis For AF Relaying

The APEP at D for the SM MIMO-AF system can be calculated as

$$\begin{aligned} P_D^{AF}(\mathbf{x} \to \hat{\mathbf{x}}) = E\Big\{\Pr\Big\{ & \left\|\mathbf{y}^{SD} - \mathbf{H}^{SD}\mathbf{x}\right\|^2 / N_0 + \left\|\mathbf{C}^{-1/2}\left(\mathbf{y}^{RD} - G\mathbf{H}^{RD}\mathbf{H}^{SR}\mathbf{x}\right)\right\|^2 \\ & \geq \left\|\mathbf{y}^{SD} - \mathbf{H}^{SD}\hat{\mathbf{x}}\right\|^2 / N_0 + \left\|\mathbf{C}^{-1/2}\left(\mathbf{y}^{RD} - G\mathbf{H}^{RD}\mathbf{H}^{SR}\hat{\mathbf{x}}\right)\right\|^2 \mid \mathbf{H}^{SD}, \mathbf{H}^{SR}, \mathbf{H}^{RD}\Big\}\Big\}. \end{aligned} \tag{6.32}$$

Following the same steps as in (6.12) and after simplifications, (6.32) can be rewritten as

$$P_D^{AF}(\mathbf{x} \to \hat{\mathbf{x}}) = E\left\{Q\left(\sqrt{\gamma^{SD} + \gamma^{SRD}}\right)\right\} \tag{6.33}$$

where γ^{SD} is same as in (6.24) and γ^{SRD} is given as

$$\gamma^{SRD} = \frac{G^2 \left\| \mathbf{C}^{-1/2} \mathbf{H}^{RD} \mathbf{H}^{SR} (\mathbf{x} - \hat{\mathbf{x}}) \right\|^2}{2}. \tag{6.34}$$

The MGF of γ^{SD} can be calculated as in (6.22). On the other hand, the MGF of γ^{SRD} is given as (see Appendix A)

$$M_{\gamma^{SRD}}(s) = \frac{\det(\Phi)}{\prod_{n=1}^{r_{\min}} \Gamma(r_{\max} - n + 1) \Gamma(r_{\min} - n + 1)} \tag{6.35}$$

where $r_{\min} = \min\{N^R, N_r^D\}$, $r_{\max} = \max\{N^R, N_r^D\}$ and Φ is an $r_{\min} \times r_{\min}$ Hankel matrix whose (n,m)th entry is obtained as

$$\Phi_{n,m} = z^{-\eta}\Gamma(\eta) U\left(\eta, \eta, \frac{1}{z}\right) + G^2 z^{-\eta-1}\Gamma(\eta+1) U\left(\eta+1, \eta+1, \frac{1}{z}\right) \tag{6.36}$$

where $z = G^2\left(1 - \frac{\rho\lambda_x s \sigma_{SR}^2}{2}\right)$, $\eta = r_{\max} - r_{\min} + n + m - 1$ and $U(\cdot,\cdot,\cdot)$ is the confluent hypergeometric function of the second kind.

Using Craig's formula again, (6.33) is expressed as

$$P_D^{AF}(\mathbf{x} \to \hat{\mathbf{x}}) = \frac{1}{\pi} \int_0^{\frac{\pi}{2}} M_{\gamma^{SD}}\left(-\frac{1}{2\sin^2\theta}\right) M_{\gamma^{SRD}}\left(-\frac{1}{2\sin^2\theta}\right) d\theta \tag{6.37}$$

where numerical integration is needed. When the numerical integration is not used, (6.37) can be simplified and upper bounded as

$$P_D^{AF}(\mathbf{x} \to \hat{\mathbf{x}}) < \frac{1}{\pi} M_{\gamma^{SD}}\left(-\frac{1}{2}\right) M_{\gamma^{SRD}}\left(-\frac{1}{2}\right). \tag{6.38}$$

ABEP can be obtained using the union bound as [11]

$$P_b^{AF} \le \frac{1}{N_t M \log_2(N_t M)} \sum_{\mathbf{x}} \sum_{\substack{\hat{\mathbf{x}} \\ \hat{\mathbf{x}} \neq \mathbf{x}}} n(\mathbf{x} \to \hat{\mathbf{x}}) P_D^{AF}(\mathbf{x} \to \hat{\mathbf{x}}). \tag{6.39}$$

On the other hand, when SNR$\to \infty$, i.e., $\lambda_x \sigma_{SD(SR)}^2 / N_0 \to \infty$, (6.38) is approximated to

$$P_D^{AF}(\mathbf{x} \to \hat{\mathbf{x}}) \approx \frac{1}{\pi}\left(1 + \frac{\rho\lambda_x\sigma_{SD}^2}{4}\right)^{-N_r^D} \left(G^2\left(1 + \frac{\rho\lambda_x\sigma_{SR}^2}{4}\right)\right)^{-r_{\min}} \tag{6.40}$$

where we use $U(a,b,z) \approx z^{1-b}$ for small z [59]. Finally, the diversity order of the MIMO-AF system can be observed from (6.40) as $N_r^D + \min\{N^R, N_r^D\}$.

6.4 Performance Evaluation

In this section, analytical and computer simulation results for the BEP of cooperative SM systems are presented. Moreover, BER performance comparisons of the cooperative SM (both AF and DF) and classical cooperative systems are performed. In these comparisons, classical cooperative systems are selected as in [9] where the considered modulation is classical M-ary PSK or QAM. On the other hand, modulation order is chosen to provide the same spectral efficiency as in the cooperative SM systems. Since, only a single transmit antenna is active in the SM systems, a single antenna is employed for classical M-PSK/QAM systems ($N_t^S = 1$). At receiver, MRC is considered for classical cooperative systems. Monte Carlo simulations are performed for at least 10^6 channel uses as a function of the received SNR at D and compared with the analytical results. The path loss exponent is chosen as $\alpha = 3$.

6.4.1 Results for DF relaying

In order to obtain the same spectral efficiency, the number of transmit antennas and modulation orders are taken identical for S-R and R-D links, i.e., $N_t^S = N_t^R = N_t$ and $M^S = M^R = M$ for MIMO-DF relaying.

BER performance of the cooperative SM system with DF relaying is given in Figure 6.3. The number of receive antennas for R and D are considered as the same, i.e., $N_r^R = N_r^D = 2$. The computer simulations are evaluated for different spectral efficiencies, by considering different number of transmit antennas and modulation orders. As seen from Figure 6.3, theoretical curves closely match with computer simulation results at high SNR due to the use of BER union bound.

In Figure 6.4, the effect of number of receive antennas on the BER performance is investigated. Computer simulation results are depicted for $N_t^S = N_t^R = 2$, $M = 2$ (BPSK) and different number of receive antennas at R and D. As seen from Figure 6.4, since the lower modulation orders and number of transmit/receive antennas are chosen, the analytical curves and computer simulation results are in close match also at lower SNR values. On the other hand, the slope of the curves, i.e., the diversity order, depends on the number of receive antennas at R and D, i.e., $\min\left\{N_r^R, 2N_r^D\right\}$. In Figures 6.3 and 6.4, unit distances are considered between all nodes, i.e., $d_{SD} = d_{SR} = d_{RD} = 1$.

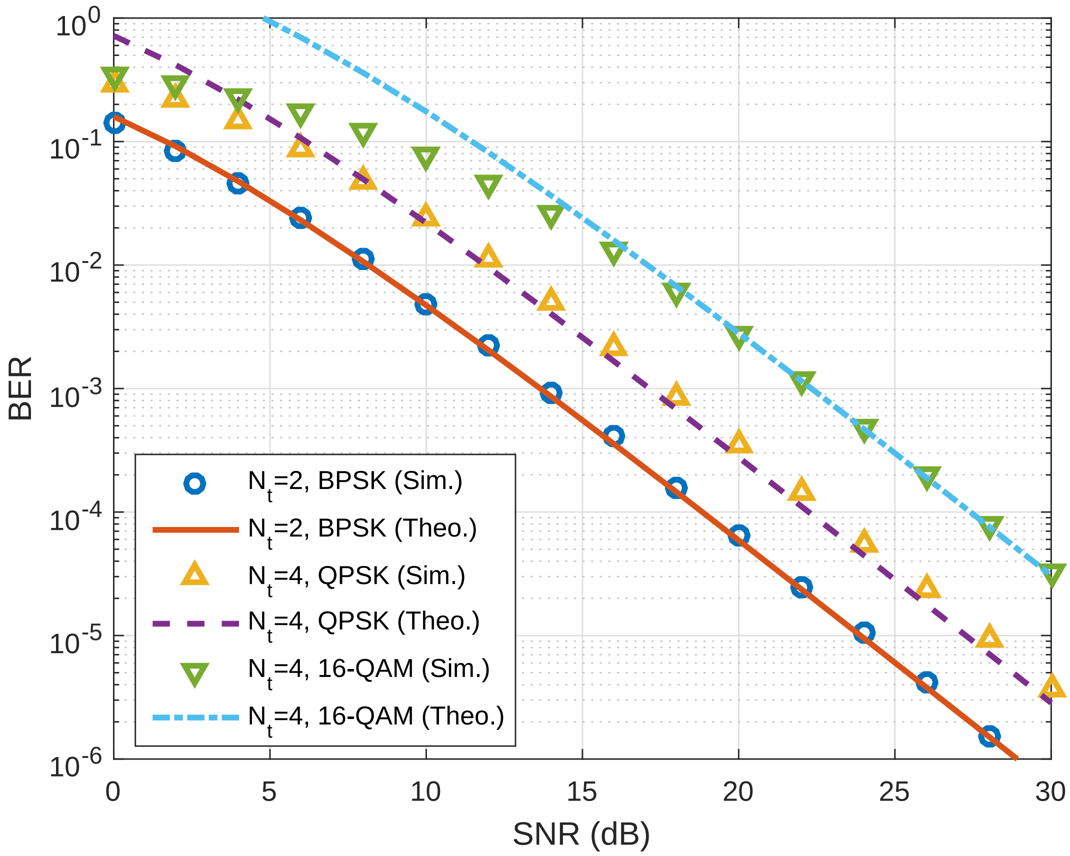

Figure 6.3 : BER performance of cooperative SM system with DF relaying for different number of transmit antennas and modulation orders, $N_t^S = N_t^R = N_t$ and $N_r^R = N_r^D = 2$.

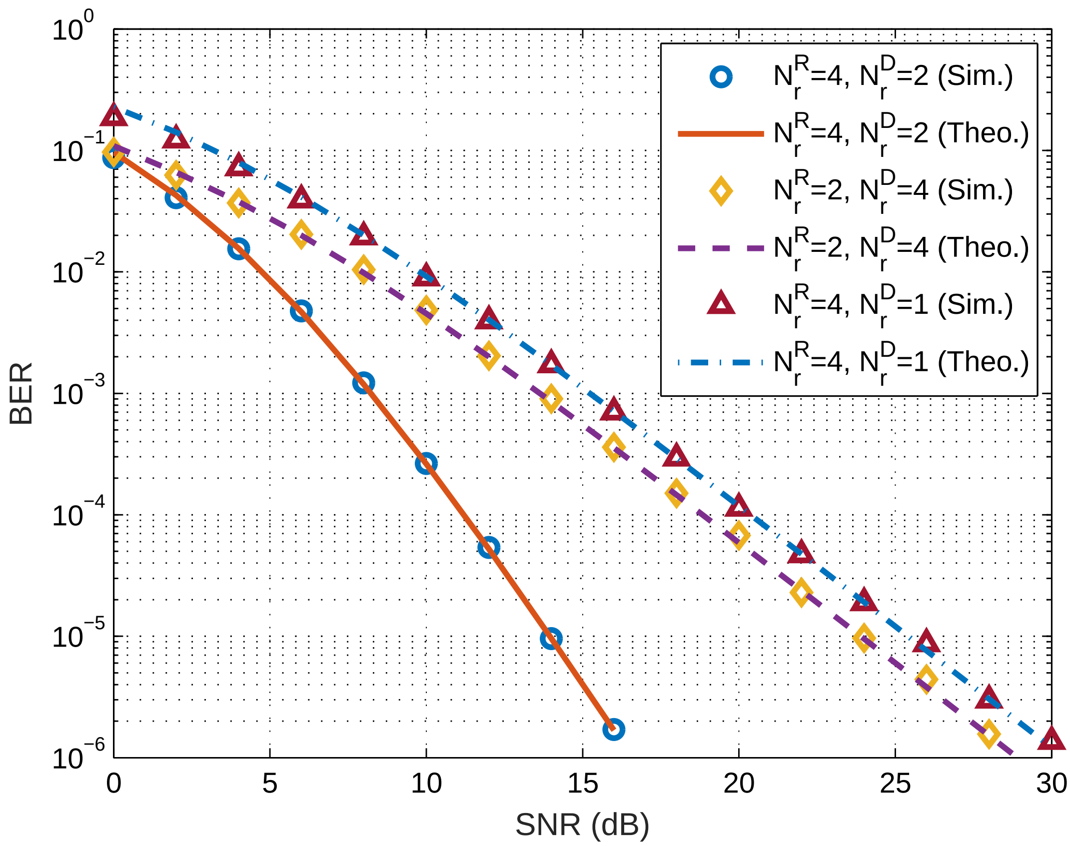

Figure 6.4 : BER performance of cooperative SM system with DF relaying for different number of receive antennas at R and D, $N_t^S = N_t^R = N_t = 2$, $M = 2$ (BPSK).

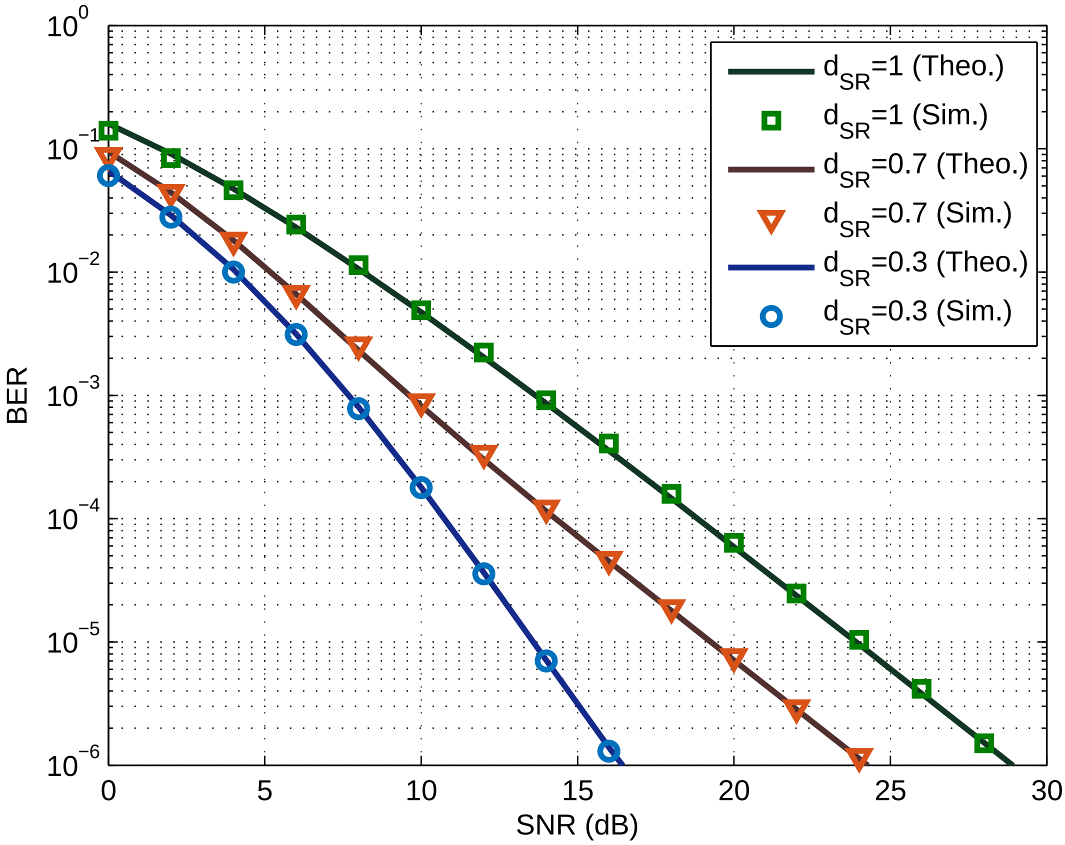

Figure 6.5 : BER performance of cooperative SM system with DF relaying for different relay locations ($d_{SD} = d_{RD} = 1$).

In Figure 6.5, the effect of different distances between nodes on the BER performance is evaluated both analytically and theoretically. The SM parameters are chosen as $N_t^S = 2$, $M = 2$ (BPSK) and the number of receive antennas at R and D is $N_r^R = N_r^D = 2$. A unit distance is assumed between S to D and R to D ($d_{SD} = d_{RD} = 1$), and the following three different cases are considered for the distances between S and R: $d_{SR} = 0.3$, $d_{SR} = 0.7$ and $d_{SR} = 1$. It can be seen from Figure 6.5 that the BER performance of DF relaying, more specifically the diversity order, improves when R gets closer to S. For this case, σ_{SR}^2 becomes larger than σ^2 and when the SNR goes to infinity, (6.31) dominates the diversity order that converges to $2N_r^D$.

The BER performance comparison of cooperative SM and classical cooperative M-ary modulated systems for $R = 3$, 4 and 5 bits/s/Hz spectral efficiencies is given in Figure 6.6, where $R = \log_2(N_t M)$ assuming two receive antennas at R and D. As seen from Figure 6.6, cooperative SM system provides approximately 2 dB SNR gain over the corresponding classical APM modulated cooperative system when R and D have two receive antennas.

6.4.2 Results for AF relaying

Computer simulations are performed as a function of the power spent at S to noise ratio (P_S/N_0) for AF relaying. The power spent at R is taken equal to P_S, i.e., $P_S = P_R$. In all computer simulations, unit power is assumed at each nodes.

Theoretical and computer simulation results for AF relaying are given in Figure 6.7, where three different SM configurations are considered: i) $N_t^S = 2$, $N_t^R = N_r^R = N^R = 2$, $N_r^D = 2$, ii) $N_t^S = 2$, $N^R = 4$, $N_r^D = 2$, and iii) $N_t^S = 4$, $N^R = 4$, $N_r^D = 4$. For all configurations QPSK is considered. As seen from Figure 6.7, theoretical results exactly match with the computer simulation results at high SNR values.

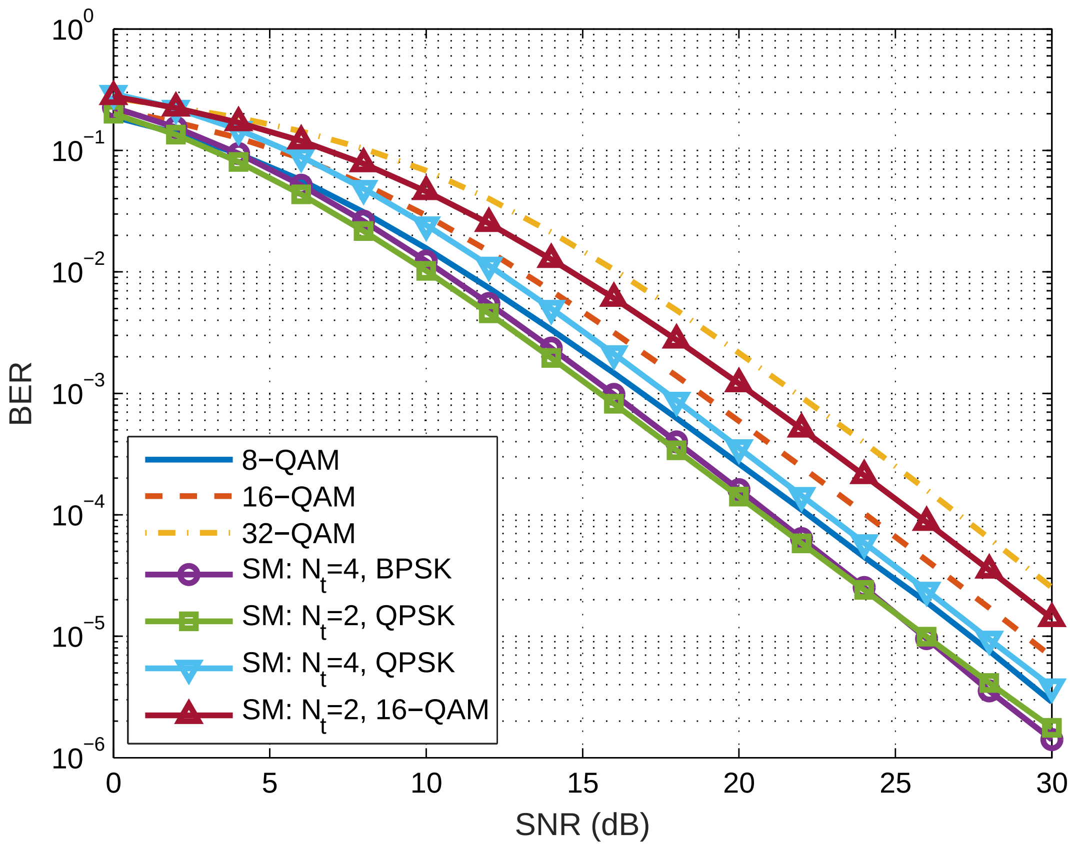

Figure 6.6 : BER performance comparison of cooperative SM with classical cooperative M-ary modulation for DF relaying systems ($N_r^R = N_r^D = N_r = 2$).

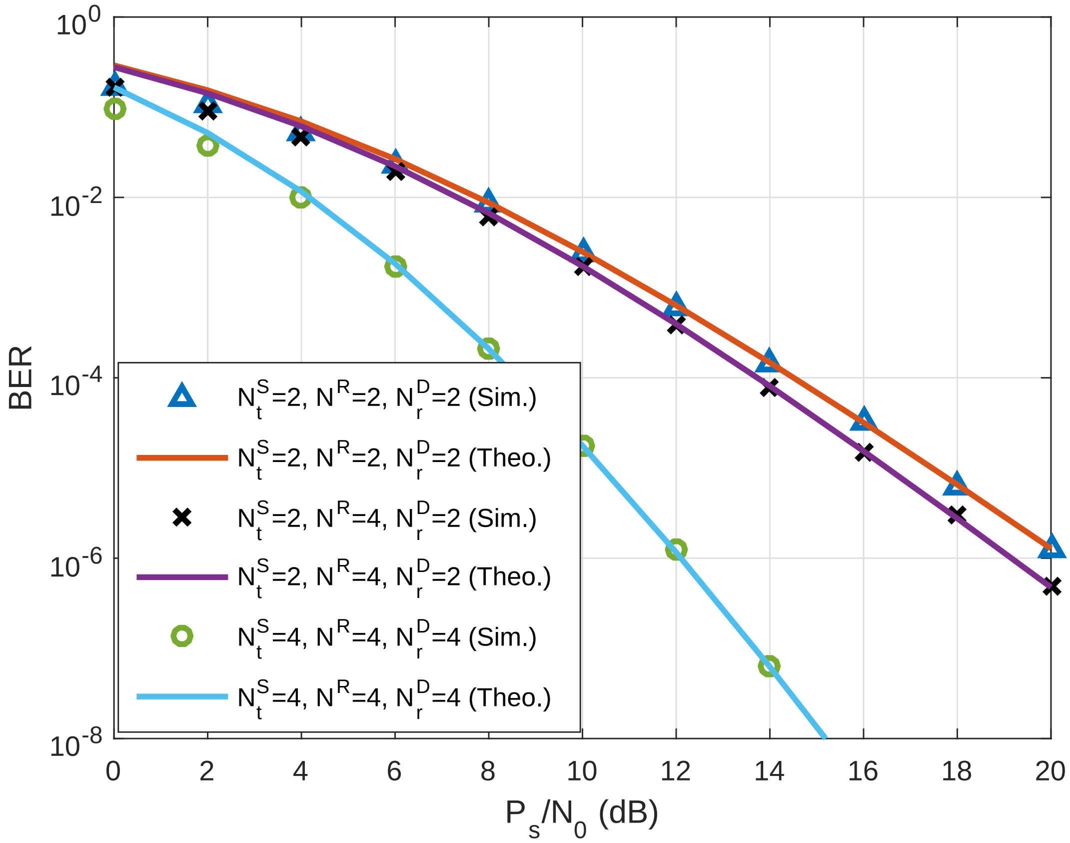

Figure 6.7 : BER performance of SM MIMO-AF relaying with different configurations. $N_t^R = N_r^R = N^R$ with QPSK.

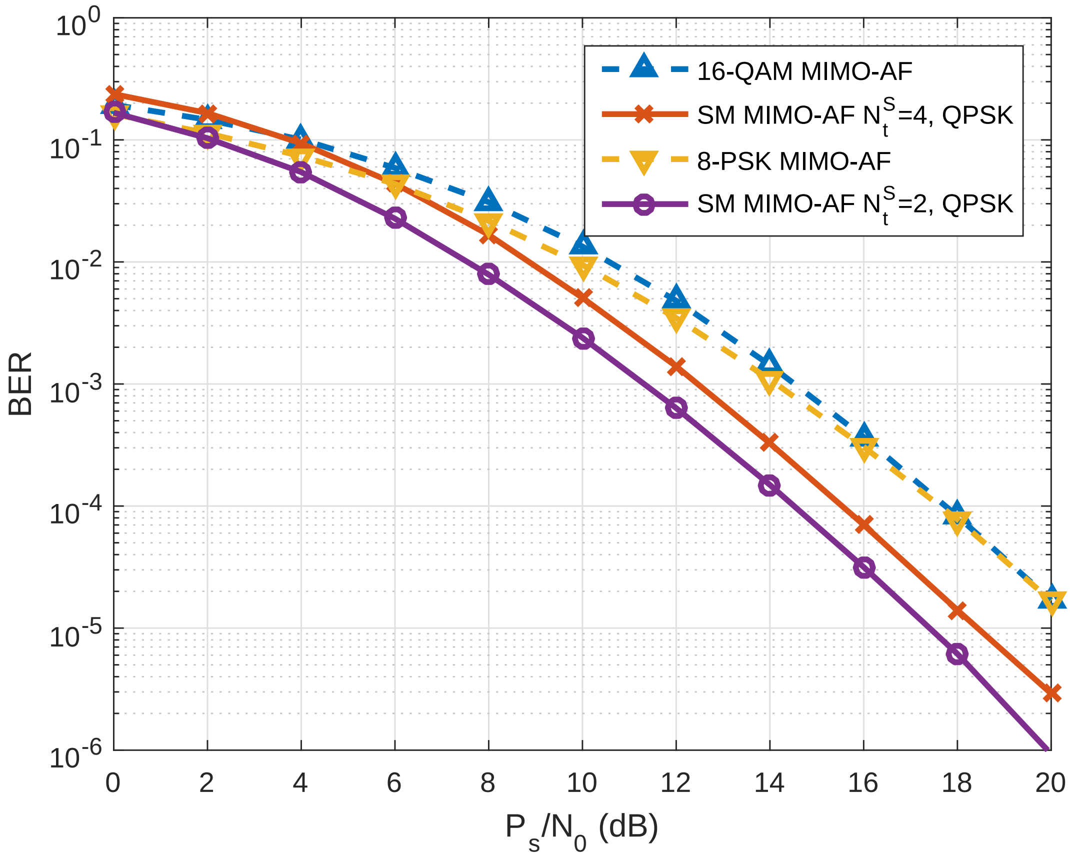

Figure 6.8 : BER comparison of MIMO-AF cooperative SM system with classical M-PSK/QAM modulated cooperative system ($N_t^R = N_r^R = N_r^D = 2$).

In Figure 6.8, the BER comparison of MIMO-AF cooperative SM system and classical M-PSK/QAM modulated cooperative system with ML decision at D is given for $R = 3$ and $R = 4$ bits/s/Hz spectral efficiencies. The classical modulated systems also use all of the available antennas at R, denoted by M-PSK/QAM MIMO-AF. Since only a single transmit antenna is active in the SM system, number of transmit antennas for classical M-PSK/QAM system is chosen as $N_t^S = 1$. The number of transmit/receive antennas for R and D is $N_r^R = N_r^D = 2$ for both systems. As seen from Figure 6.8, the cooperative SM scheme provides better error performance than classical cooperative systems for the same spectral efficiency. It can be observed that the SM system provides approximately 3 and 4 dB SNR gains for $R = 3$ and $R = 4$ bits/s/Hz spectral efficiencies, respectively.

6.4.3 Comparison of MIMO-DF and MIMO-AF systems

The comparison results of MIMO-DF and MIMO-AF cooperative SM systems can be found in Figs. 6.9 and 6.10. SM parameters are chosen as $N_t^S = 2$, $M = 2$ (QPSK) for both figures. In Figure 6.9, the effect of different number of receive antennas at R and D is investigated. When the number of receive antennas at R is lower, the error propagation has influence on the BER of the DF system so that MIMO-AF system has better error performance. Otherwise, MIMO-DF system outperforms MIMO-AF.

Since the location of R directly affects the BER performance of the DF system, its impact is examined in Figure 6.10. As can be seen from Figure 6.10, when the links S to R and R to D have $d_{SR} = d_{RD} = 0.5$, MIMO-AF system provides better error performance above the SNR value of 5 dB. When R gets closer to S, i.e., $d_{SR} = 0.4$, MIMO-DF system exhibits better BER performance than MIMO-AF up to the SNR value of 12 dB. This shows that error propagation effect is still significant at these distances.

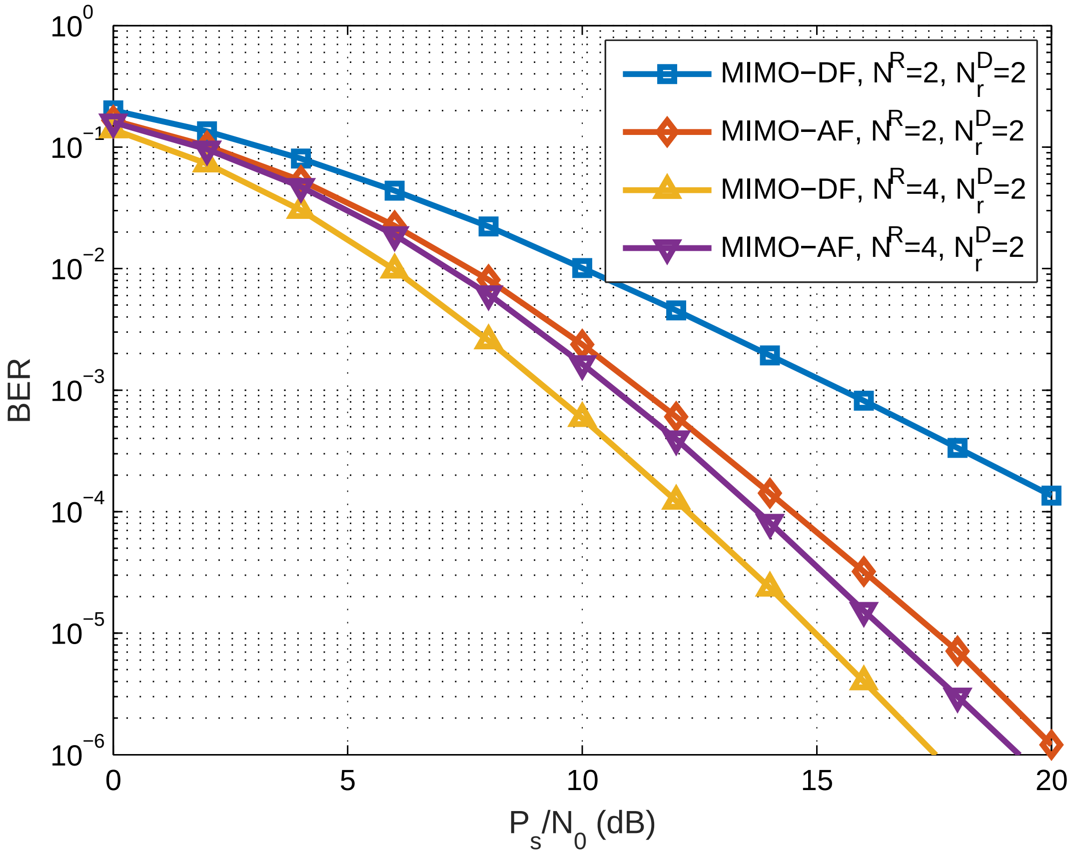

Figure 6.9 : BER comparison of MIMO-AF and MIMO-DF cooperative SM systems under different number of receive antennas, $N_t^S = 2$, $M = 4$ (QPSK).

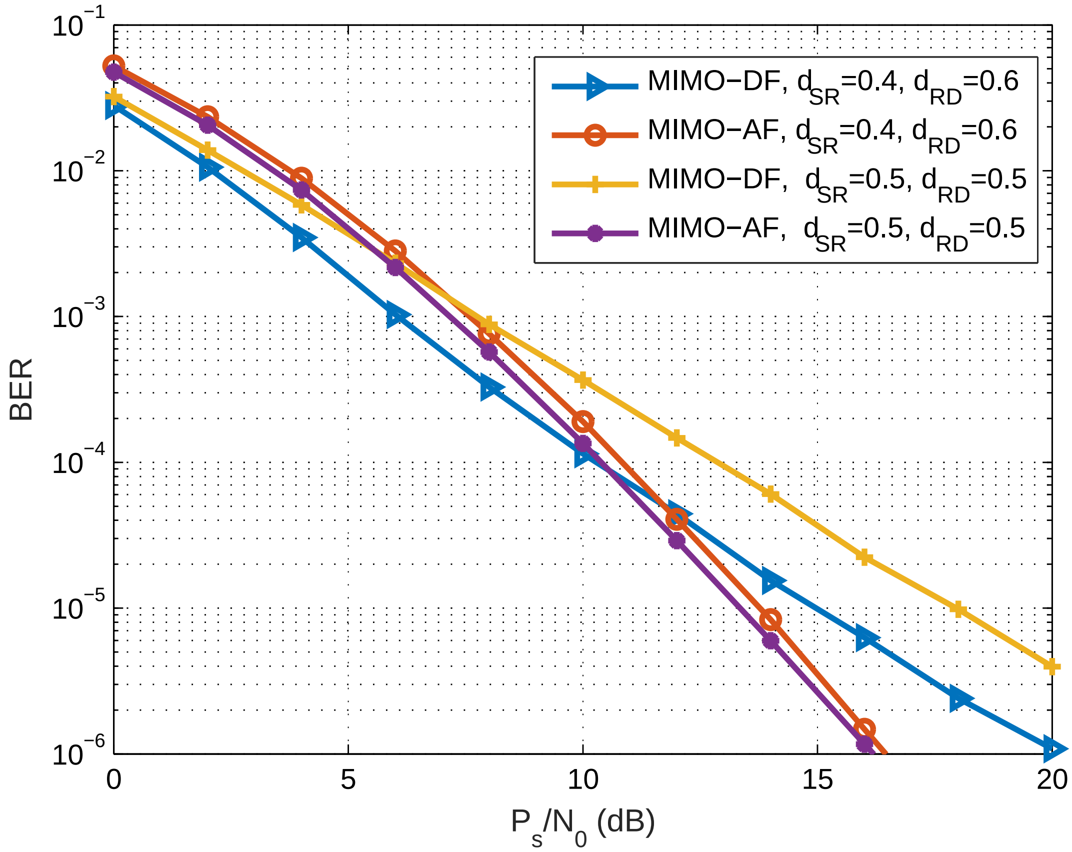

Figure 6.10 : BER comparison of MIMO-AF and MIMO-DF cooperative SM systems under different R locations, $N_t^S = 2$, $M = 4$ (QPSK), $N^R = N_r^D = 2$, $d_{SD} = 1$.

7. CONCLUSIONS AND RECOMMENDATIONS

In this thesis, we have evaluated the outage probability performance of both classical SM and cooperative SM systems. Due to the fact that SM conveys information not only by the modulated symbols, but also by the antenna indices, the capacity calculation of SM differs from the conventional systems. The corresponding capacity can be calculated as the sum of the capacities of APM and the space domains. Interrelated with the capacity, the outage probability performance of an SM system, which is not thoroughly examined in the literature, is deeply affected by the antenna index errors. In this work, we have derived the outage probability performance of point to point SM and the cooperative SM systems with classical DF and AF relaying, as well as dynamic relaying techniques such as SR and IR, for the worst case scenario. In both systems, although the minimum channel gains are selected, a better outage performance is achieved compared to the classical systems.

We also have investigated the ABEP performance of the cooperative SM scheme for AF and DF relaying. Most of the studies on cooperative SM systems in the literature consider only the SSK scheme with a single receive antenna at R and/or D. In this work, we have derived analytical expressions for the ABEP of AF and DF multi-antenna cooperative communications systems that employ SM. We also have presented a diversity order analysis and have demonstrated the effect of relay location to DF relaying. Since the BER performance is derived analytically from the union bound, computer simulations and analytical results have demonstrated that the derived expressions for the ABEP exactly match with the computer simulation results in high SNR region. Furthermore, we have presented the comparison results of MIMO-DF and MIMO-AF relaying with classical M-PSK/QAM systems. For both relaying systems, SM technique provides error performance gain over conventional cooperative APM systems. Finally, the comparison results of MIMO-AF with MIMO-DF cooperative SM systems have been introduced. Investigation of real practical problems such as signaling design and synchronization, channel estimation errors, resource management, interference, etc., has been left as a future study.

Additionally, we have investigated new and simple algorithms to reduce the computational complexity of optimal solution of GSM and GSIM systems. Although the GSM and GSIM increase the overall spectral efficiency, the receiver complexity also increases with the number of transmit antennas. The optimal solution of GSM/GSIM is the ML detection process which has greater complexity with the exhaustive search. In this work, we propose new and simple algorithms which are based on the N best estimate values of a reduced cost function to achieve a near ML solution with lower computational complexity. Our proposed algorithm for GSM system shows approximately same performance with respect to ML with low N values. On the other hand, for high activated transmit antenna numbers, GSIM shows relatively unsatisfactory performance compared to the optimal solution. In order to overcome this problem, another low complexity algorithm using sphere decoding is presented. Complexity calculations show that proposed algorithms have lower complexity even if the values of N are high.

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APPENDICES

APPENDIX A.1 : Derivation of (6.35)

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To find $M_{\gamma^{SRD}}$, we extend γ_{SRD}, given in (6.34), as

$$\gamma_{SRD} = \frac{G^2}{2} \operatorname{tr}\left(\left(\mathbf{H}^{RD}\right)^H \mathbf{C}^{-1} \mathbf{H}^{RD} \mathbf{H}^{SR} (\mathbf{x}-\hat{\mathbf{x}}) (\mathbf{x}-\hat{\mathbf{x}})^H \left(\mathbf{H}^{SR}\right)^H \right). \tag{A.1}$$

Using the eigenvalue decomposition for the Hermitian matrix $(\mathbf{x}-\hat{\mathbf{x}})(\mathbf{x}-\hat{\mathbf{x}})^H$, we have

$$(\mathbf{x}-\hat{\mathbf{x}})(\mathbf{x}-\hat{\mathbf{x}})^H = \mathbf{V}\Lambda\mathbf{V}^H. \tag{A.2}$$

Because the rank of the matrix $(\mathbf{x}-\hat{\mathbf{x}})(\mathbf{x}-\hat{\mathbf{x}})^H$ is equal to one, we have a single eigenvalue λ_x and the corresponding eigenvector $\mathbf{v}$ where λ_x is as in (6.21) with the indices defined in $\hat{\mathbf{x}}$. Since the elements of the channel matrices are Gaussian r.v.'s, the elements of the $\mathbf{H}^{SR}\mathbf{v}$ vector are also Gaussian. Let us define the vector $\mathbf{d} \triangleq \mathbf{H}^{SR}\mathbf{v}$, hence we obtain

$$\gamma_{SRD} = \frac{G^2\lambda_x}{2} \left(\mathbf{d}^H \left(\mathbf{H}^{RD}\right)^H \mathbf{C}^{-1} \mathbf{H}^{RD} \mathbf{d} \right). \tag{A.3}$$

The MGF of γ^{SRD} is given as

$$\begin{aligned} M_{\gamma^{SRD}}(s) &= E\left\{ e^{s\gamma^{SRD}} \right\} \\ &= E_{\mathbf{H}^{RD},\mathbf{d}}\left\{ \exp\left(\frac{G^2\lambda_x s}{2} \left(\mathbf{d}^H \left(\mathbf{H}^{RD}\right)^H \mathbf{C}^{-1} \mathbf{H}^{RD} \mathbf{d} \right) \right) \right\}. \end{aligned} \tag{A.4}$$

As mentioned earlier, $\mathbf{d}$ is a complex Gaussian vector with the covariance matrix $\mathbf{R_d} = \sigma_{SR}^2 \mathbf{I}_{N^R}$ where $N_r^R = N_t^R = N^R$. Therefore, the MGF of γ^{SRD} given $\mathbf{H}^{RD}$ can be obtained from [60, Theorem D.1]

$$M_{\gamma^{SRD}|\mathbf{H}^{RD}}(s) = \frac{1}{\det\left(\mathbf{I}_{N^R} - \frac{G^2\lambda_x\sigma_{SR}^2 s}{2} \left(\mathbf{H}^{RD}\right)^H \mathbf{C}^{-1} \mathbf{H}^{RD} \right)}. \tag{A.5}$$

It is difficult to integrate (A.5) with respect to $\mathbf{H}^{RD}$. On the other hand, the eigenvalue decomposition of the Hermitian matrix $\mathbf{H}^{RD}\left(\mathbf{H}^{RD}\right)^H$ can be used, i.e., $\mathbf{H}^{RD} = \mathbf{V}\Lambda\mathbf{V}^H$ where $\mathbf{V} \in \mathbb{C}^{N^R \times N^R}$ is a unitary matrix and $\Lambda \in \mathbb{R}^{N_r^D \times N^R}$ is a diagonal matrix with ordered eigenvalues $\lambda_1^{RD} > \lambda_2^{RD} > \cdots > \lambda_{r_{\min}}^{RD}$ along its main diagonal, which yields [61]

$$\begin{aligned} M_{\gamma^{SRD}|\Lambda}(s) &= \frac{1}{\det\left(\mathbf{I}_{N_r^D} - \frac{G^2\lambda_x\sigma_{SR}^2 s}{2N_0} \Lambda \left(\mathbf{I}_{N_r^D} + G^2\Lambda \right)^{-1} \right)} \\ &= \prod_{n=1}^{r_{\min}} \left(\frac{1}{1 - \frac{G^2\lambda_x\sigma_{SR}^2 s \lambda_n^{RD}}{2N_0\left(1+G^2\lambda_n^{RD}\right)}} \right) \end{aligned} \tag{A.6}$$

where $r_{\min} = \min\{N^R, N_r^D\}$. The joint pdf of ordered eigenvalues is [62]

$$f_\Lambda(\lambda) = \frac{\prod_{n<m}^{r_{\min}} (\lambda_m - \lambda_n)^2 \prod_{p=1}^{r_{\min}} \lambda_p^{r_{\max}-r_{\min}} e^{-\lambda_p}}{\prod_{n=1}^{r_{\min}} \Gamma(r_{\max} - n + 1) \Gamma(r_{\min} - n + 1)} \tag{A.7}$$

where $r_{\max} = \max\{N^R, N_r^D\}$. The MGF of γ^{SRD} can be derived using (A.6) and (A.7) as

$$M_{\gamma^{SRD}}(s) = \int_{\lambda_1^{RD}} \cdots \int_{\lambda_{r\min}^{RD}} \left(1 - \frac{G^2 \lambda_x \sigma_{SR}^2 s \lambda_n^{RD}}{2N_0 (1 + G^2 \lambda_n^{RD})}\right)^{-1} f_\Lambda(\lambda^{RD}) \quad d\lambda_1^{RD} \cdots d\lambda_{r\min}^{RD}. \tag{A.8}$$

(A.8) can be calculated using *Corollary 2* in [61] as

$$M_{\gamma^{SRD}}(s) = \mathscr{K}^{-1} \det\left(\left[\int_0^\infty \left(1 - \frac{G^2 \lambda_x \sigma_{SR}^2 s \lambda_n^{RD}}{2N_0 (1 + G^2 \lambda_n^{RD})}\right)^{-1} \times \lambda^{(r_{\max} - r_{\min} + n + m - 2)} e^{-\lambda} d\lambda\right]_{n,m=1,\cdots,r_{\min}}\right). \tag{A.9}$$

where $\mathscr{K}^{-1} = \prod_{n=1}^{r_{\min}} \Gamma(r_{\max} - n + 1) \Gamma(r_{\min} - n + 1)$. Performing some algebraic manupulations in (A.9), the result in (6.35) can be obtained. This completes the derivation.

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