CONCLUSION AND FUTURE WORKS

Thesis contributions:

The thesis contributions are presented as follows

1. Proposingthree low complexity detectors and a tighter BEP upper bound for HRSM systems;

2. Proposing a Spatially Modulated Space Time Block Coding (DT-SM) scheme for MIMO systems equipped with the number of transmit antennas greater than 4. Simulation and analytical results show that the DT-SM scheme have potential applications for MIMO systems requiring high spectral efficiency;

3. Proposing Diagonal Space Time Coded Spatial Modulation (DS-SM) scheme for MIMO systems equipped with the even number of transmit antennas. This scheme achieves fourth order transmit diversity. As a result, this is suitable for MIMO systems requiring high commumnication quality.

Future works:

1. Extending the reasearch of other low complexity detectors such as lattice reduction algorithm, soft detectors as well as PIC ones for the HRSM scheme;

2. Studying optimal estimation algorithms for MIMO-SM schemess;

3. Evaluating the HRSM performance under correlated MIMO channel and imperfect CSI at the receiver as well as deriving a tighter BEP upper bound for the HRSM scheme at low SNR region;

4. Studing the SM applications in massive MIMO systems as well as investigating the index modulation technique.

INTRODUCTION

1. Research backgound:

Mesleh *et. al* proposed a Spatial Modulation (SM) technique to take the MIMO advantages and overcome the existing MIMO challenges. At each symbol period only one transmit antenna is activated to transmit a modulated symbol. Therefore, the ICI problem at the receiver and the IAS at the transmitter are completely avoided in the MIMO-SM system. However, the SM technique is lack of transmit diversity. A combination of the Space Time Block Codes (STBC) and the SM is an effective way to increase transmit diversity. Several combined schemes are proposed such as a Space Time Block Coded Spatial Modulation (STBC-SM) [5], a Spatially Modulated Orthogonal Space Time Block Coding (SM-OSTBC) [35], a Spatially Modulated Diagonal Space Time Code (SM-DC) [70]. Therefore, this combination can build new MIMO-SM schemes having transmit diversity and improving spectral efficiency. This is also a potential research direction that has attrated scientists over the world.

2. Thesis contributions:

1. Proposing modified low complexity detectors and a tighter new BEP upper bound at high SNR region for HRSM systems;
2. Proposing a Spatially Modulated Double Space Time Coding (DT-SM) scheme for MIMO systems equipped with the number of transmit antennas greater than 4. This scheme obtains second order transmit diversity and has low detection complexity;
3. Proposing a Diagonal Space Time Block Coded Spatial Modulation (DS-SM) scheme for MIMO systems equipped with the even number of transmit antennas. This scheme attains fourth order transmit diversity and has a reasonable detection complexity.

3. Thesis outline:

The thesis includes 130 pages and is presented in 4 chapters except for the introduction, conclusion, and references.

Chapter 1

BACKGROUND OF THE SPATIAL MODULATION AND SPACE TIME CODES

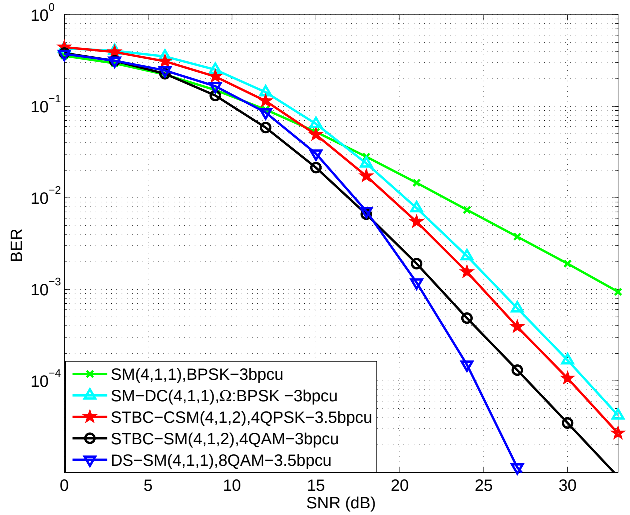
1.1 The spatial modulation principle

In the SM transmiter, each block of data bits is splitted into two different parts. The first part is used to activate one transmit antenna in the transmit antenna array (the spatial information). The remaining bits are mapped into the signal constellation of M-QAM or PSK technique to choose a modulated symbol (the signal information). Finally, the modulated symbol is transmitted from the active antenna. Therefore, the conveyed information including the spatial information (the antenna indices) and the signal information (the modulated symbol) is presented as a 3-D constellation diagram.

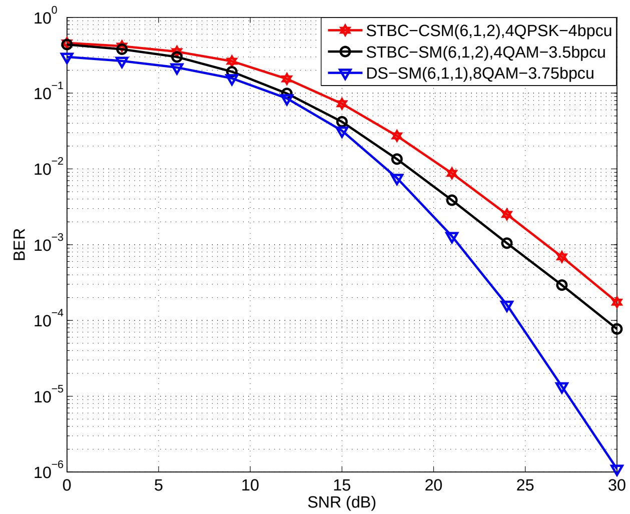
1.2 The MIMO-SM system model

A MIMO-SM system equipped with n_T transmit antennas, n_R receive antennas, and a M-ary modulation technique is considered. At the transmitter, each block of $(\log_2 n_T + \log_2 M)$ information bits is separated into two different parts: $\log_2 n_T$ and $\log_2 M$ bits. The $\log_2 n_T$ bits are used to activate one antenna in the transmit antenna array while the remaining $\log_2 M$ bits are mapped into a M-ary modulation technique to create a modulated symbol. Finally, this symbol is transmitted from the active antenna.

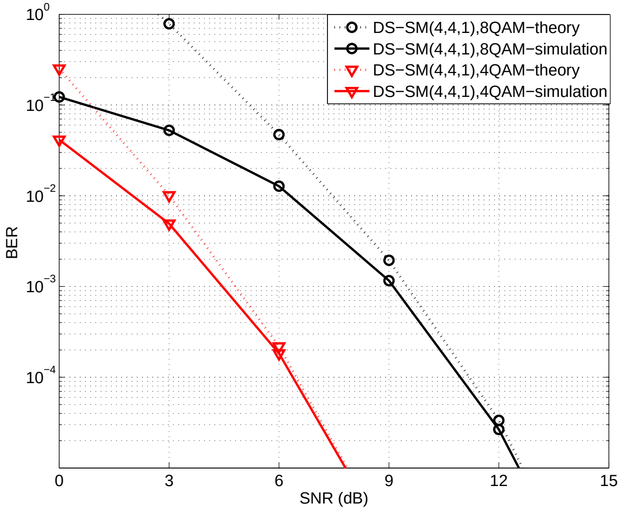
Fig. 1.1 decribes the SM scheme operation where this model is equipped with four transmit antennas, four receive antennas, and a 4-QAM technique. In this model, each block of four data bits is processed by the transmitter. The first two bits are used to activate a transmit antenna while the last two bits are mapped into the 4-QAM technique to create a modulated symbol. For example, the first block of data is 0011 where two bits, 00, are used to activate the first antenna in four transmit antennas while two bits, 11, are mapped into the 4-QAM technique to generate the modulated symbol, i.e., $1 + j$. Finally, this symbol is transmitted from the first transmit antenna. At the receiver, four data bits, 0011, are exactly recovered by a ML detector.



Hình 4.4: Performance comparison of the DS-SM with the SM, STC-SM, SM-DC, and STBC-CSM when $n_R = 1$ at spectral efficiency 3 bpcu.



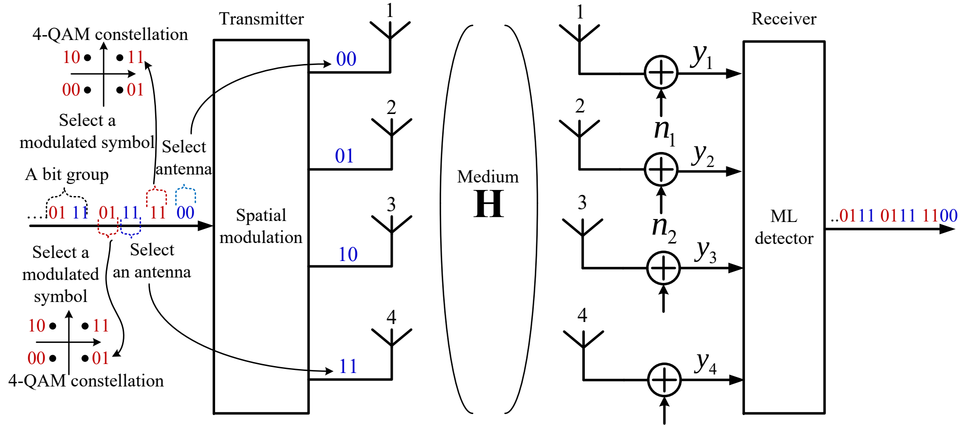
Hình 4.5: Performance comparison of the DS-SM, STBC-SM, and STBC-CSM when using 6 transmit, 1 receive antennas at spectral efficiency 4 bpcu.



Hình 4.3: Simulation and theoretical results of the (4,4) DS-SM scheme using 4-QAM and 8-QAM.

4.5 Conclusion

In this chapter, a new MIMO-SM scheme for MIMO systems equipped with the even number of transmit antennas greater than 4, called DS-SM, is proposed. This scheme obtains fourth order transmit diversity and has a reasonable detection complexity.



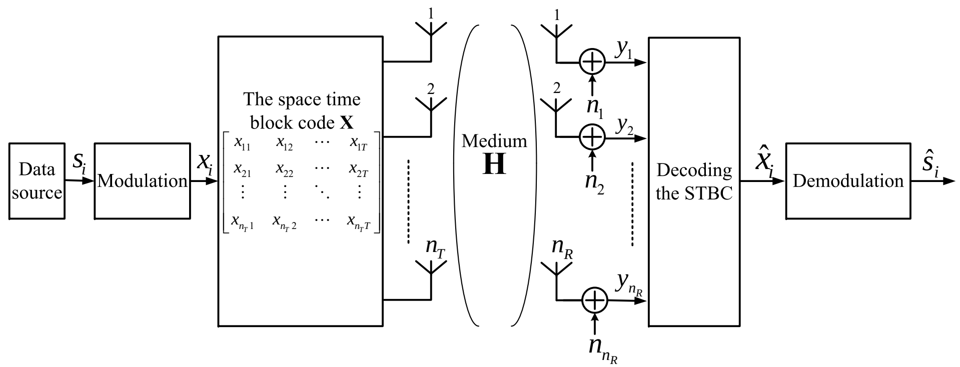
Hình 1.1: A MIMO-SM system, 4 transmit antennas, 4-QAM.

The MIMO-SM spectral efficiency is presented as given

$$C_{\mathrm{SM}} = \log_2 n_T + \log_2 M \ \ (\text{bpcu})\,,$$

where bpcu, denoting bits per channel use, expresses the number of transmitted bits in once channel utilization.

1.3 The space time codes



Hình 1.2: A general diagram of the STBC system.

Fig. 1.2 decribes a MIMO system ulitized the STBC technique. In this model, information bits s_i are mapped into M-ary modulation technique to create N_s modulated symbols, $x_i, i = 1, 2, ..., N_s$. Then, these symbols are arranged to make a matrix block including n_T rows and T columns as given

$$\mathbf{X} = \begin{bmatrix} x_{11} & x_{12} & \cdots & x_{1T} \\ x_{21} & x_{22} & \cdots & x_{2T} \\ \vdots & \vdots & \ddots & \vdots \\ x_{n_T 1} & x_{n_T 2} & \cdots & x_{n_T T} \end{bmatrix},$$

where $x_{v,t}$ is the value or the complex congugate of the incoming symbol sequence x_i, $i = 1, 2, ..., N_s$. $x_{v,t}$ is transmitted from the v-th transmit anttenna and the t-th time slot.

Because N_s symbols are transmitted in T symbol periods, the STBC code rate is $R = \frac{N_s}{T}$. Besides, two criteria in the STBC design to achieve full diveristy and coding gain are proposed. From two different coding matrices $\mathbf{X_1}$ and $\mathbf{X_2}$, the difference matrix is defined as given

$$\mathbf{A}(\mathbf{X}_1, \mathbf{X}_2) = \mathbf{X}_1 - \mathbf{X}_2. \tag{1.1}$$

The determinant of the difference matrix is calculated as follows

$$\mathbf{B}(\mathbf{X}_1, \mathbf{X}_2) = \mathbf{A}^H \mathbf{A}. \tag{1.2}$$

- **Rank criterion:** In order to achieve full diversity $n_T n_R$ for the STBC $\mathbf{X}$, with two arbitrary matrices $\mathbf{X}_i$ and $\mathbf{X}_q$, $i \neq q$, the difference matrix $\mathbf{A}(\mathbf{X}_i, \mathbf{X}_q)$ has to be full rank;
- **Determinant criterion:** To increase coding gain for a full diversity code, the minimum determinant $\mathbf{B}(\mathbf{X}_i, \mathbf{X}_q)$ has to be maximized.

1.4 Conclusion

Basic knowledge directly related to research subjects including the spatial modulation technique, the space time codes, and the MIMO-SM sytem research background is presented in detail in the first chapter. This knowledge will be used as a theoretical background supportting the research issues in the later chapter.

4.3.2 *Theoretical BEP upper bound for the DS-SM system*

The BEP upper bound of the DS-SM scheme can be derived as given

$$P_b \leq \frac{1}{N} \sum_{i=1}^{N} \sum_{q=1}^{N} \frac{P(\mathbf{C}_i \to \mathbf{C}_q) w_{i,q}}{\log_2 N}, \tag{4.14}$$

where $N = KM^4$ và $w_{i,q}$ is the number of bits in error between the matrices $\mathbf{C}_i$ and $\mathbf{C}_q$.

The conditional PEP of the DS-SM system is calculated as

$$P(\mathbf{C}_i \to \mathbf{C}_q | \mathbf{H}) = Q\left(\sqrt{\frac{\gamma}{2}} d^2(\mathbf{C}_i, \mathbf{C}_q)\right), \tag{4.15}$$

where $Q(x) = (1/2\pi) \int\limits_x^\infty e^{-y^2/2} dy$. The PEP is given as

$$P(\mathbf{C}_i \to \mathbf{C}_q) = \frac{1}{\pi} \int\limits_0^{\frac{\pi}{2}} \left(\frac{1}{1 + \frac{\gamma \lambda_{i,q,1}}{4\sin^2\phi}}\right)^{n_R} \left(\frac{1}{1 + \frac{\gamma \lambda_{i,q,2}}{4\sin^2\phi}}\right)^{n_R} \cdots \left(\frac{1}{1 + \frac{\gamma \lambda_{i,q,4}}{4\sin^2\phi}}\right)^{n_R} d\phi, \tag{4.16}$$

where $\lambda_{i,q,1}, \lambda_{i,q,2}, \lambda_{i,q,3}, \lambda_{i,q,4}$ is the eigenvalues of the codeword distance $(\mathbf{C}_i - \mathbf{C}_q)(\mathbf{C}_i - \mathbf{C}_q)^H$.

4.4 Simulation results

Fig. 4.3 illustrates the theoretical and simulation results for the BER performance of the DS-SM scheme in two spectral efficiencies such as 2,5 and 3,5 bpcu. We can see that that both results are coincident at the high SNR region. This result comfirms the accuracy of the DS-SM performance.

Fig. 4.4 illustrates the DS-SM $(4, n_R, 1)$, SM $(4, n_R, 1)$, STBC-SM $(4, n_R, 2)$, and STBC-CSM $(4, n_R, 2)$ with the receive antennas $n_R = 1$ at spectral efficiency 3 bpcu. The DS-SM scheme outperforms the existing ones at sufficiently high SNR region. Particularly, at $\text{BER} = 10^{-3}$, the DS-SM scheme achieves SNR gains about 1.1 dB, 2.7 dB, 4.7 dB, and 11.5 dB over the STBC-SM, STBC-CSM, SM-DC, and SM ones, respectively.

Similarly, Fig. 4.5 compare the performance of the DS-SM $(6, n_R, 1)$ with STBC-SM $(6, n_R, 2)$, and STBC-CSM $(6, n_R, 2)$ with the receive antennas $n_R = 1$. We can see that the DS-SM scheme outperforms the STBC-SM and STBC-CSM ones. At $\text{BER} = 10^{-3}$, the DS-SM scheme obtains 0.2 dB and 2 dB SNR gains over the STBC-SM and STBC-CSM ones, respectively.

4.3.1 Complexity analysis

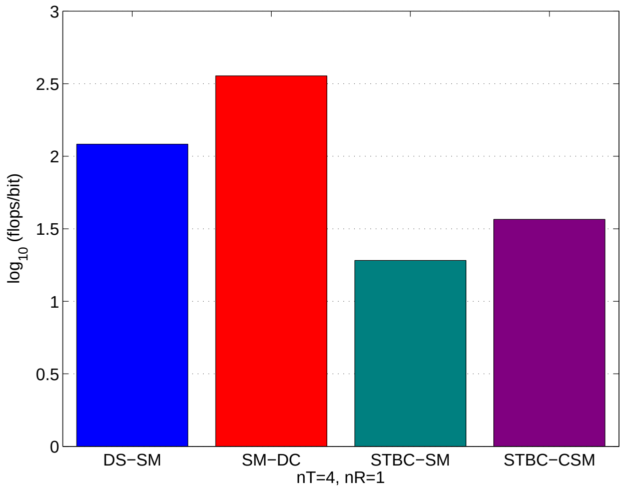
Similarly, the complexity of the pre-processing state is calculated as given

$$\rho_{\text{pre}} = \frac{4}{T}\left(32 n_T n_R + 2040 n_R - 36\right) K + \left(143 n_R + 7\right) K. \tag{4.12}$$

Therefore, the DS-SM complexity is genralized as follows

$$\rho = \frac{\rho_{\text{pre}} + \rho_s}{4m + \lfloor \log_2 (2n_T) \rfloor}, \tag{4.13}$$

where ρ_s is the number of average operations within SD searching stage.



Hình 4.2: Complexity comparison of the DS-SM, SM-DC, STBC-SM, and STBC-CSM at the spectral efficiency 3 bpcu, SNR 9dB, 4 transmit and 1 receive antenna, $T = 80$ symbol periods.

Fig. 4.2 compares the detection complexity of the DS-SM with the existing MIMO-SM schemes such as SM-DC, STBC-SM, và STBC-CSM, in a MIMO configuration with 4 transmit and a receive antennas at spectral efficiency 3 bpcu. It can see that the DS-SM complexity is lower than the SM-DC one. However, the DS-SM scheme offers higher detection complexity than the STBC-SM và STBC-CSM ones. In return for this disadvantage, the DS-SM spectral efficiency is higher 0.5 bpcu than that of the STBC-SM and STBC-CSM ones.

Chapter 2

LOW COMPLEXITY DETECTION ALGORITHMS AND A NEW THEORETICAL UPPER BOUND FOR THE HIGH RATE SPATIAL MODULATION SCHEME

2.1 The modified MMSE-VBLAST detector

Considering the HRSM system as an SDM one, the MMSE filter matrix at the receiver can be computed as follows

$$\mathbf{G}_{\text{MMSE}} = \tilde{\mathbf{H}}^H \left(\tilde{\mathbf{H}} \tilde{\mathbf{H}}^H + \frac{1}{E_s} \mathbf{I}_{n_R} \right)^{-1}, \tag{2.1}$$

where $\tilde{\mathbf{H}} = \sqrt{\frac{\gamma}{n_T E_s}} \mathbf{H}$. The modified BLAST (MBLAST) detection algorithm is presented in Table 2.1.

2.2 The modified MMSE-SQRD detector

An $(n_T + n_R) \times n_T$ extended channel matrix and an extended received vector $\mathbf{z}$ are defined as given

$$\mathbf{D} = \begin{bmatrix} \tilde{\mathbf{H}} \\ \frac{1}{\sqrt{E_s}} \mathbf{I}_{n_T} \end{bmatrix}, \mathbf{z} = \begin{bmatrix} \mathbf{y} \\ \mathbf{0} \end{bmatrix}.$$

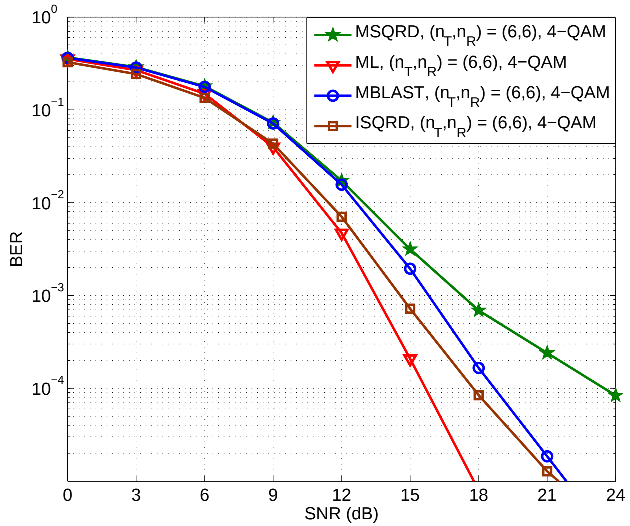
Doing QR decomposition of the matrix $\mathbf{D} = \mathbf{QR}$ where $\mathbf{Q}^H \mathbf{Q} = \mathbf{I}_{n_T}$. $\mathbf{R}$ is an $n_T \times n_T$ upper triangular matrix. Multiplying $\mathbf{z}$ with $\mathbf{Q}^H$, we have $\mathbf{v} = \mathbf{Rc} + \mathbf{w}$. The modified SQRD (MSQRD) algorithm is summarized in Table 2.2.

2.3 The improved SQRD detection algorithm

The HRSM system equation is re-expresses as given

$$\mathbf{t}_x = \bar{\mathbf{H}} \bar{\mathbf{c}} + \mathbf{n}, \tag{2.2}$$

where $\mathbf{t}_x = \mathbf{y} - \tilde{\mathbf{h}}_1 x$, $\bar{\mathbf{H}} = \begin{bmatrix} \tilde{\mathbf{h}}_2 & \tilde{\mathbf{h}}_3 & \cdots & \tilde{\mathbf{h}}_{n_T} \end{bmatrix}$, $\tilde{\mathbf{h}}_k$ is the k-th column of $\tilde{\mathbf{H}}$. $\bar{\mathbf{c}}$ includes the $n_T - 1$ remaining entries of the $\mathbf{c}$ vector (remove the first entry). The improved SQRD (ISQRD) detection algorithm is summarized in Table 2.3.



Hình 2.1: BER comparisons of ML, MBLAST, MSQRD, ISQRD in the HRSM (6,6) using 4-QAM.

2.4 Simulation results

Fig. 2.1 shows thats the ML algorithm offers the best performance while the MSQRD one has the worst performance. Particularly, at $\text{BER} = 10^{-4}$ in Fig. 2.1, the MSQRD detector loses about 8.5 dB, 5.5 dB, 4.8 dB SNR gains compared with ML, ISQRD, and MBLAST ones, respectively. At $\text{BER} = 10^{-5}$, the MBLAST and ISQRD detectors have lower performance about 4.5 dB and 4 dB than the ML one.

2.5 A new BEP upper bound of the HRSM system

we define the concept of "nearest symbol" as follows: Let $x, x' \in \Omega_x$, $x' \neq x$ where Ω_x is a QAM constellation. Then x and x' are said to be "nearest symbol" if and only if the Euclidean distance between x and x' is minimum.

Theorem: Given a transmitted HRSM codeword matrix $\mathbf{c}_i = \mathbf{s}_k x_n$, $i = 1,2,...,N$; $k = 1,2,...,K$; and $n = 1,2,...,M$, Let us define V_n is the set of nearest symbols of x_n and $\bar{V}_n = \Omega_x - V_n$ is the set of remaining symbols that are not nearest symbols of x_n. In addition, we define the set of SC codewords from the nearest symbols as $\Omega_i = \Omega_{\mathbf{c}} - \left\{\mathbf{c} \in \Omega_{\mathbf{c}} | \mathbf{c} = \mathbf{s}_k x_v, x_v \in \bar{V}_n\right\}$ where $\Omega_{\mathbf{c}}$ is the set of the total N HRSM codewords $\mathbf{c}$. The new upper bound for the BEP of the HRSM

1. For given even n_T transmit antennas, arbitrary M modulation level. using Table 4.1 to choose the suitable value of θ and generate the basic set of 4 SC codewords.
2. Adding a $(n_T - 4) \times 4$ zeros matrix $\mathbf{0}$ under the columns of the basic SC codewords from the equation (4.2). Then, these matrices are cycled in two rows to create $2n_T$ new SC codewords.
3. Depending on the information bits to choose the suitable number of SC codewords $\mathbf{S}_k, \forall k = 1,2,..., \lfloor \log_2 (2n_T) \rfloor$.

4.3 The DS-SM signal detection

For a given matrix $\mathbf{S}_k, k = 1,2,...,K$, the $n_R \times 4$ equivalent matrix is created as $\tilde{\mathbf{H}}_k = \sqrt{\frac{\gamma}{E_s}} \mathbf{H}\mathbf{S}_k$. Therefore, the system equation in (4.3) can be re-written as

$$\mathbf{Y} = \tilde{\mathbf{H}}_k \mathbf{X} + \mathbf{N}. \tag{4.7}$$

Based on the $\mathbf{X}$ diagonal structure, the formula (4.7) can be re-expressed as given

$$\mathbf{y} = \mathbf{H}_{e,k} \tilde{\mathbf{x}} + \mathbf{n}, \tag{4.8}$$

where $\mathbf{H}_{e,k} = \left[\ \text{diag}(\tilde{\mathbf{h}}_1) \quad \text{diag}(\tilde{\mathbf{h}}_2) \quad \cdots \quad \text{diag}(\tilde{\mathbf{h}}_{n_R}) \ \right]^T$. $\tilde{\mathbf{h}}_k, k = 1,...,n_R$, is the k-th row of $\tilde{\mathbf{H}}_k$. $\mathbf{y} = \text{vec}\left(\mathbf{Y}^T\right)$, $\mathbf{n} = \text{vec}\left(\mathbf{N}^T\right)$. $\text{vec}\left(\mathbf{A}\right)$ denotes vectorial stacking operation of the matrix $\mathbf{A}$. Converting the formula (4.8) into the equivalent real system-equation, we have

$$\mathbf{v} = \mathbf{M}_k \mathbf{s} + \mathbf{w}, \tag{4.9}$$

where $\mathbf{M}_k = \begin{bmatrix} \Re\left(\mathbf{H}_{e,q}\right) & -\Im\left(\mathbf{H}_{e,q}\right) \\ \Im\left(\mathbf{H}_{e,q}\right) & \Re\left(\mathbf{H}_{e,q}\right) \end{bmatrix} \mathbf{W}$, $\mathbf{s} = \left[\Re\left(\mathbf{u}\right), \Im\left(\mathbf{u}\right)\right]^T$, $\mathbf{w} = \left[\Re\left(\mathbf{n}\right), \Im\left(\mathbf{n}\right)\right]^T$, $\mathbf{v} = \left[\Re\left(\mathbf{y}\right), \Im\left(\mathbf{y}\right)\right]^T$. The system equation in (4.9) is now similar to that of a conventional spatial multiplexing scheme. Therefore, the SD algorithm can be used to detect $\mathbf{s}$ for a given $\mathbf{S}_k$ as follows

$$\left(\hat{\mathbf{s}}\right)_k = \arg\min_{\mathbf{s}} \left\|\mathbf{t}_k - \mathbf{R}_k \mathbf{s}\right\|^2, \tag{4.10}$$

where $\mathbf{t}_k = \mathbf{Q}_k^H \mathbf{v}$ và $\mathbf{Q}_k \mathbf{R}_k$ is matrices from QR decomposition of $\mathbf{M}_k$. After that, the index k of the transmitted SC codeword is determined as given

$$\hat{k} = \arg\min_{k} \left\|\mathbf{t}_k - \mathbf{R}_k \left(\hat{\mathbf{s}}\right)_k\right\|^2 + \mathbf{v}^H \mathbf{v} - \mathbf{t}_k^H \mathbf{t}_k. \tag{4.11}$$

Finally, information bits are recovered from a pair of the detected SC codeword and the detected signal vector $\left(\hat{\mathbf{S}}_k, \hat{\mathbf{u}}_k\right)$.

where $w_{tk} = \cos\left(\frac{\pi}{4n}(4t-1)(2k-1)\right)$ với $1 \le t, k \le n$ and $n = 8$. The rotated signal vector $\tilde{\mathbf{u}}$ is arranged as follows

$$\tilde{\mathbf{x}} = [\begin{array}{cccc} \tilde{u}_1 + j\tilde{u}_5 & \tilde{u}_2 + j\tilde{u}_6 & \tilde{u}_3 + j\tilde{u}_7 & \tilde{u}_4 + j\tilde{u}_8 \end{array}]^T. \quad (4.2)$$

The 4×4 diagonal STBC matrix $\mathbf{X}$ is then obtained by arranging the vector $\tilde{\mathbf{x}}$ as $\mathbf{X} = \mathrm{diag}(\tilde{\mathbf{x}})$ The DS-SM codeword $\mathbf{C}$ is created as $\mathbf{C} = \mathbf{SX}$. The channel is assumed to be quasi static flat Rayleigh fading, the $n_R \times 4$ received signal $\mathbf{Y}$ is presented as given

$$\mathbf{Y} = \sqrt{\frac{\gamma}{E_s}}\mathbf{HC} + \mathbf{N} = \sqrt{\frac{\gamma}{E_s}}\mathbf{HSX} + \mathbf{N} \quad (4.3)$$

4.2 SC codeword design

In a DS-SM scheme using 4 transmit antennas, a basic set of 4 SC codewords is proposed as given

$$\mathbf{S}_1 = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 \end{bmatrix}; \mathbf{S}_2 = \begin{bmatrix} 0 & e^{j\theta} & 0 & 0 \\ 0 & 0 & e^{j\theta} & 0 \\ 0 & 0 & 0 & e^{j\theta} \\ j & 0 & 0 & 0 \end{bmatrix}; \quad (4.4)$$
$$\mathbf{S}_3 = \mathbf{S}_2^2; \mathbf{S}_4 = \mathbf{S}_2^3.$$

The angle θ is searched in the $[0, \pi/2]$ interval to find optimal angle value, θ_o having maximum CDG, $\delta_{\min}(\theta)$ for a pair of DS-SM codewords $\mathbf{C} \ne \mathbf{C}'$ as given

$$\delta_{\min} = \min_{\mathbf{C} \ne \mathbf{C}'} \det\left[(\mathbf{C} - \mathbf{C}')^H (\mathbf{C} - \mathbf{C}')\right], \quad (4.5)$$

$$\theta_o = \arg\max_{\theta \in [0, \pi/2]} \delta_{\min}(\theta). \quad (4.6)$$

Finally, the angle and CDG results for different modulation techniques are summarized in Table 4.1.

Bảng 4.1: Optimal values of θ and correspoding CDGs for basic SC codewords

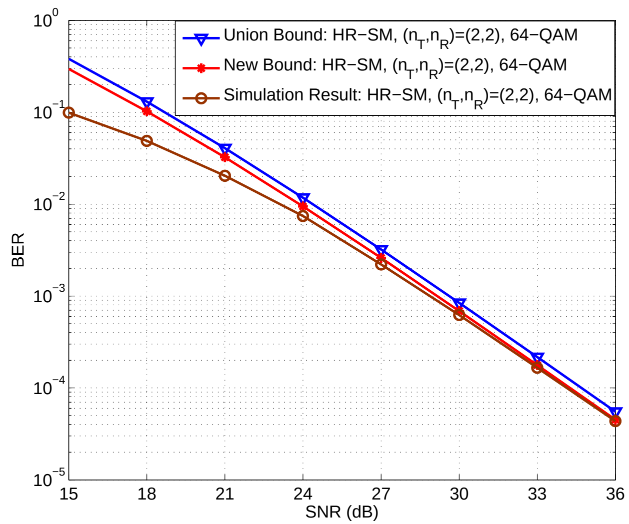
Modulation	BPSK	4QAM	8QAM	16QAM
θ_o(rad)	0.52	1.36	0.2	0.4
$d_{\min}$	0.11	0.037	$3.6\,10^{-3}$	7.10^{-4}

Generally, an extended spatial constellation for DS-SM systems equipped with $n_T = 2n > 4$ transmit antennas is presented as given

scheme is given as follows

$$P_e \le \frac{1}{(l+m)N} \sum_{i=1}^{N} \sum_{\mathbf{c}_q \in \Omega_i} P(\mathbf{c}_i \to \mathbf{c}_q)\, w_{i,q}. \quad (2.3)$$

2.6 Result analysis



Hình 2.2: The new upper bound, the union bound and simulation curve for the BER of the (2,2) HRSM system using 64-QAM.

Fig. 2.2 shows the new upper bound, the union bound and simulation results for the BER of the (2,2) HRSM system using 64-QAM. As indicated in Fig. 2.2, we can see that the union bound and the new bound are very close to the simulation curve in sufficiently high SNR region, SNR $\ge$ 30 dB, however, in low SNR region they are still loose. We observe that the new bound is closer to the simulation curve. Particularly, at $\mathrm{BER} = 10^{-3}$, the new bound is tighter than the union bound by an SNR gain of approximately 0.5 dB.

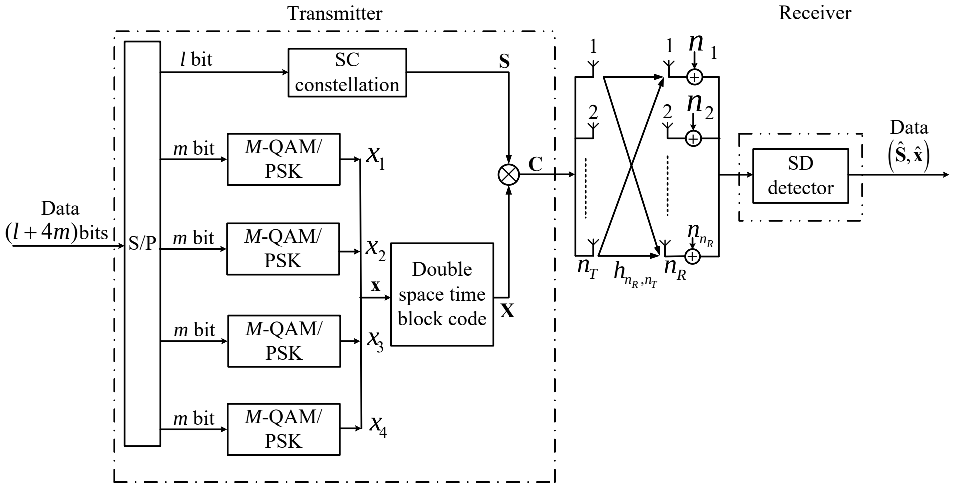
2.7 Conclusion

Three low complexity detectors and a tighter new BEP upper bound for HRSM systems are proposed in this chapter.

Chapter 3

A NEW MIMO-SM SCHEME ACHIEVING HIGH SPECTRAL EFICIENCY

3.1 The DT-SM system model



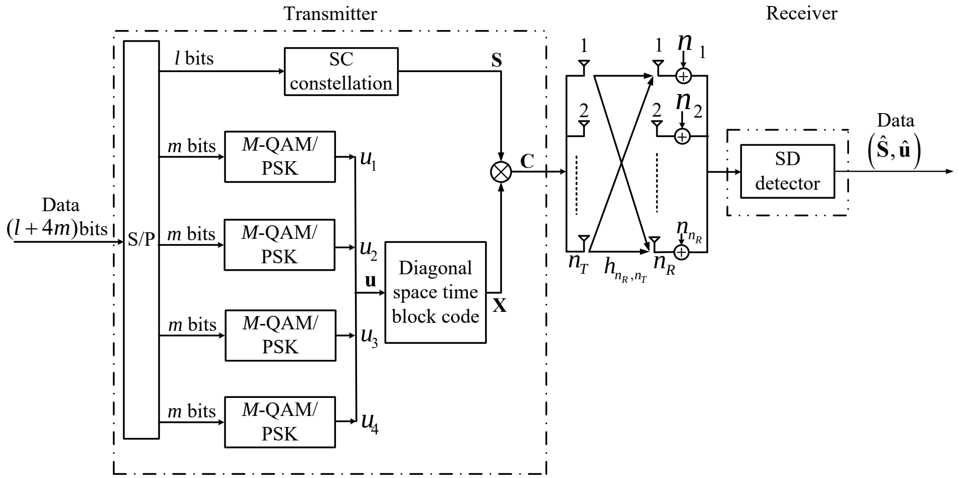
Hình 3.1: A DT-SM system diagram.

Fig. 3.1 presents a DT-SM scheme equipped with n_T transmit and n_R receive antennas. At each symbol period, 4 out of n_T transmit antennas are activated to emit signals into the air. Assuming that data are fed into the DT-SM transmitter in blocks including $(l + 4m)$ bits. The first part, l bits, are mapped into a SC codeword $\mathbf{S}$ which is selected from the set of K SC codewords $(K = 2^l)$ in the spatial constellation Ω_S. The last part, $4m$ bits, are modulated by M-QAM/PSK modulators to get a 4×1 signal vector $\mathbf{x} = [\ x_1 \quad x_2 \quad x_3 \quad x_4\ ]^T$. The DSTTD matrix $\mathbf{X}$ is created as $\mathbf{X} = \begin{bmatrix} x_1 & x_2 & x_3 & x_4 \\ -x_2^* & x_1^* & -x_4^* & x_3^* \end{bmatrix}^T$. The DT-SM codeword $\mathbf{C}$ is created as $\mathbf{C} = \mathbf{SX}$. the channel is assumed to be quasi static

Chapter 4

A NEW MIMO-SM SCHEME ACHIEVING HIGH ORDER TRANSMIT DIVERSITY

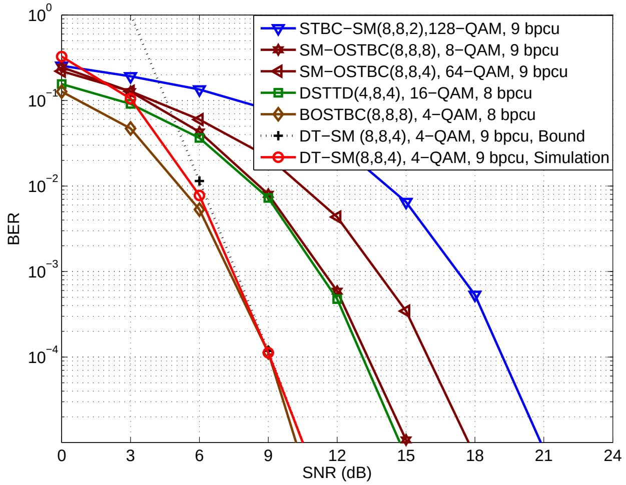
4.1 The DS-SM system model



Hình 4.1: Block diagram of the DS-SM scheme.

Spatial mapping is similar to the DT-SM scheme. Meanwhile, The last part, $4m$ bits, are modulated by M-QAM/PSK modulators, $(M = 2^m)$, to make a 4×1 modulated symbol vector $\mathbf{u} = [\ u_1 \quad u_2 \quad u_3 \quad u_4\ ]^T$. At DSTBC block, a rotated signal vector is created as $\tilde{\mathbf{u}} = \mathbf{W}[\Re(\mathbf{u}), \Im(\mathbf{u})]^T$. The rotation matrix $\mathbf{W}$ is presented as given

$$\mathbf{W} = \sqrt{\frac{2}{n}} \begin{bmatrix} w_{11} & w_{12} & \cdots & w_{1n} \\ w_{21} & w_{22} & \cdots & w_{2n} \\ \vdots & \vdots & \ddots & \vdots \\ w_{n1} & w_{n2} & \cdots & w_{nn} \end{bmatrix}, \tag{4.1}$$



Hình 3.5: BER performances of the DT-SM, STBC-SM, SM-OSTBC $C(8,8,8)$, SM-OSTBC $C(8,8,4)$ and BOSTC with $n_T = 8$, DSTTD với $n_T = 4$ at spectral efficiency of 8 bpcu and 9 bpcu. all schemes adopt $n_R = 8$ receive antennas.

tennna configuration and BER curve of the DSTTD scheme in $(n_T, n_R) = (4, 8)$ antenna configuration. As we can see from Fig. 3.5, the upper bound guarantees that our simulation results are reliable. Specifically, at $\text{BER} = 10^{-5}$, the proposed scheme offers around 3.8 dB, 4.3 dB and 7 dB gain over the DSTTD, $C(8,8,8)$, and $C(8,8,4)$, respectively. Note, however, that the proposed scheme uses 4 transmit antenna elements more than the DSTTD but 4 RF-chains fewer than the $C(8,8,8)$. Compared to the BOSTC, the proposed DT-SM has higher spectral efficiency about 1 bpcu and reduces 4 RF chains.

3.8 Conclusion

A Spatially Modulated Double Space Time Block Coding scheme, called DT-SM, is proposed. This scheme attains second order transmit diversity, high spectral efficiency and has low detection complexity at its receiver.

Rayleigh fading, the received signal $\mathbf{Y}$ is presented as given

$$\mathbf{Y} = \sqrt{\frac{\gamma}{4E_s}}\mathbf{HC} + \mathbf{N} = \sqrt{\frac{\gamma}{4E_s}}\mathbf{HSX} + \mathbf{N} \tag{3.1}$$

3.2 SC codeword design for 4 transmit antennas

3.2.1 Basic SC codeword design

Considering a DT-SM system with 4 transmit antennas, a basic set of 4 SC codewords is proposed as given

$$\mathbf{S}_1 = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ 0 & 0 & e^{j\theta} & 0 \\ 0 & 0 & 0 & e^{-j\theta} \end{bmatrix}; \mathbf{S}_2 = \begin{bmatrix} 0 & 0 & e^{j\theta} & 0 \\ 0 & 0 & 0 & e^{-j\theta} \\ 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \end{bmatrix};$$
$$\mathbf{S}_3 = \begin{bmatrix} 0 & 0 & e^{j2\theta} & 0 \\ 0 & e^{-j\theta} & 0 & 0 \\ e^{j\theta} & 0 & 0 & 0 \\ 0 & 0 & 0 & e^{-j2\theta} \end{bmatrix}; \mathbf{S}_4 = \begin{bmatrix} e^{j\theta} & 0 & 0 & 0 \\ 0 & 0 & 0 & e^{-j2\theta} \\ 0 & 0 & e^{j2\theta} & 0 \\ 0 & e^{-j\theta} & 0 & 0 \end{bmatrix}. \tag{3.2}$$

where θ is a rotation angle. In order for the DT-SM system to obtain second order transmit diversity, the angle θ must be optimized according to the rank and determinant criteria of the STCs. Then, an 4×4 appropriate precoding matrix is used as given

$$\mathbf{E} = \begin{bmatrix} 0 & e^{j\alpha} & 0 & 0 \\ 0 & 0 & e^{j\alpha} & 0 \\ 0 & 0 & 0 & e^{j\alpha} \\ 1 & 0 & 0 & 0 \end{bmatrix} \tag{3.3}$$

where α needs to be optimized. 12 new SC codewords are created as given

$$\mathbf{S}_{5:8} = \mathbf{S}_{1:4}\mathbf{E} \tag{3.4}$$
$$\mathbf{S}_{9:12} = \mathbf{S}_{1:4}\mathbf{E}^2 \tag{3.5}$$
$$\mathbf{S}_{13:16} = \mathbf{S}_{1:4}\mathbf{E}^3. \tag{3.6}$$

As a result, we have 16 SC codewords for DT-SM systems equipped with 4 transmit antennas. The α is optimized similar to the θ. The values of θ_o and α_o corresponding to minimum CDGs using different modulations are summarized in Table 3.1.

Bảng 3.1: The optimal values of θ_o, α_o, and CDGs corresponding to 16 SC codewords using different modulations.

Điều chế	BPSK	4QAM	8QAM	16QAM
θ_o(rad)	1.01	1.068	1.068	1.17
α_o(rad)	1.4	1.43	0.9	1.29
$\delta_{\min}$	1.05	0.73	0.19	0.035

3.3 SC codeword design for an arbitrary transmit antennas

the design procedure for SC codewords is summerized as follows

1. Select suitable optimum rotation angles θ_o and α_o according to Table 3.1; generate 16 basic SC codewords based on (3.2), and (3.4)-(3.6); Define $\bar{n}_T = n_T - 4$ as the number of inactive transmit antennas at a symbol period.
2. Compute the number of antenna combinations $N = \binom{n_T}{\bar{n}_T}$, and set $l = 1$.
3. Generate the l-th combination of $\bar{n}_T$ zero vectors, $\mathbf{0}$, with the 4 rows of a 4×4 matrix $\mathbf{Z}$, i.e., generate the combination vector $\mathbf{g}_l$ that indicates the indices of active and inactive antennas (making sure that $\mathbf{g}_l \neq \mathbf{g}_n$ for $n = 1, 2, \cdots, l-1$); and set $q = 1$.
4. Set $\mathbf{Z} = \mathbf{S}_q$; generate a new SC codeword using the combination vector $\mathbf{g}_l$; and multiply the newly generated SC codeword by the weight factor w_l.
5. Let $q = q + 1$; if $q \leq 16$, then go to Step 4.
6. Let $l = l + 1$; if $l \leq N$, then go to Step 3.
7. Select the first $K = 2^{\lfloor \log_2(16N) \rfloor}$ generated codewords as the desired SC codewords for the DT-SM system.

3.4 The DT-SM performance evaluation

An upper bound on the average BEP of the proposed DT-SM scheme is given by the following union bound

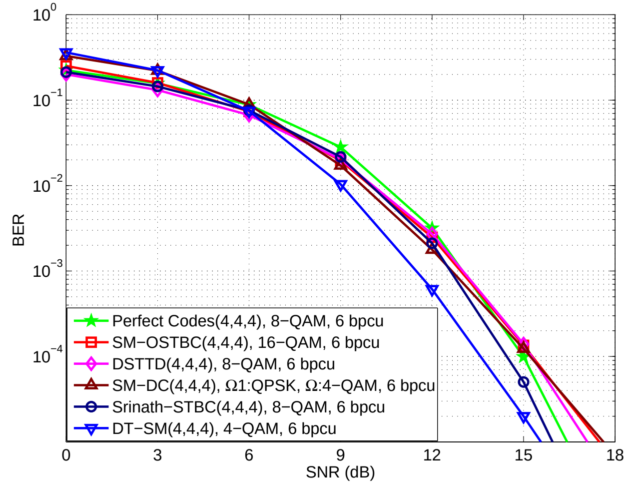
$$P_b \leq \frac{1}{N} \sum_{i=1}^{N} \sum_{q=1}^{N} \frac{P(\mathbf{C}_i \to \mathbf{C}_q)\, w_{i,q}}{b}, \tag{3.7}$$

where $P(\mathbf{C}_i \to \mathbf{C}_q)$ is a PEP of two codewords $\mathbf{C}_i$ and $\mathbf{C}_q$. $w_{i,q}$ is the Hamming distance between the two codewords $\mathbf{C}_i$ and $\mathbf{C}_q$. When the channel occurs correlation, $\mathrm{vec}(\mathbf{H}_{\mathrm{corr}}) = \mathbf{R}^{\frac{1}{2}} \mathrm{vec}(\mathbf{H})$. Then, The PEP is determined as given

$$P(\mathbf{C}_i \to \mathbf{C}_q | \mathbf{H}_{\mathrm{corr}}) = Q\left(\sqrt{\frac{1}{2}\left(\frac{\gamma}{4E_s}\right) d^2(\mathbf{C}_i, \mathbf{C}_q)}\right), \tag{3.8}$$

Perfect Codes, the SM-OSTBC and the DSTTD, it can provide better performance through the BER parameter, as shown later. Note that in Fig. 3.3 although the DT-SM exhibits a little higher detection complexity than the BOSTC, it actually provides higher spectral efficiency.

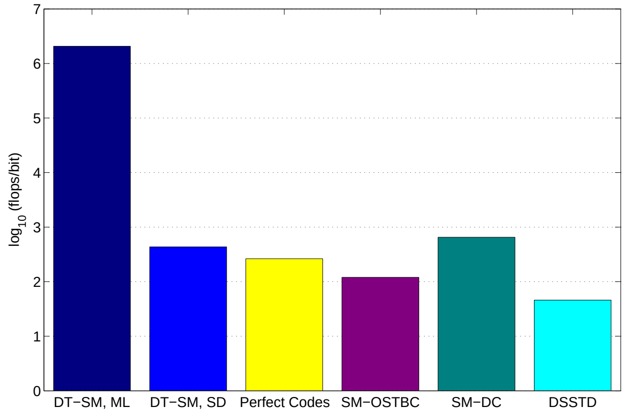
3.7 Simulation results



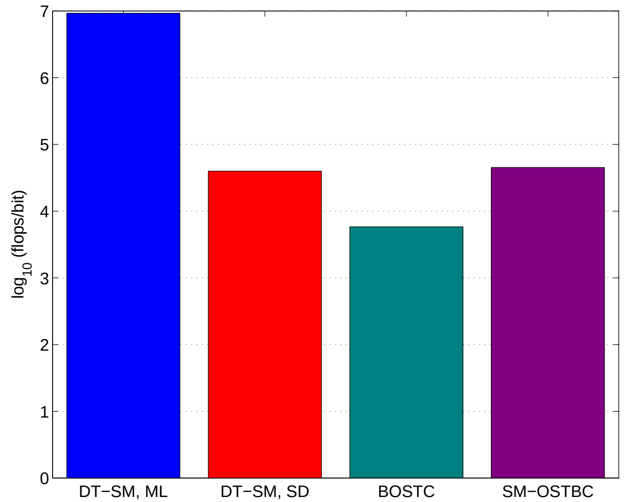
Hình 3.4: BER performances of the DT-SM, SM-OSTBC $C(4,4,4)$, SM-DC, DSTTD, Srinath-STBC, and Perfect Codes at spectral efficiency of 6 bpcu.

Fig. 3.4 compares the BER performance of the DT-SM with those of the SM-OSTBC $C(4,4,4)$, the rate-2 Srinath-STBC, the rate-2 Perfect Codes, the SM-DC, and the DSTTD at 6 bpcu for $(4,4,4)$ MIMO configuration. The DT-SM outperforms all related schemes for sufficiently high SNR $\in [6, 16]$ dB. Specifically, at $\mathrm{BER} = 10^{-3}$, the proposed scheme offers about 1.1 dB gain over the Srinath-STBC and the SM-DC, and around 1.5 dB over the SM-OSTBC, the DSTTD and the Perfect Codes. The gaps between the BER curve of the DT-SM and those of the Srinath-STBC and the Perfect Codes becomes smaller at $\mathrm{BER} = 10^{-5}$. This is due to the fact that both the Srinath-STBC and the Perfect Codes achieve higher transmit diversity order than the proposed DT-SM.

Fig. 3.5 illustrates the BER curves of the DT-SM, STBC-SM, SM-OSTBC $C(8,8,8)$, SM-OSTBC $C(8,8,4)$, and BOSTC schemes in $(n_T, n_R) = (8,8)$ an-



Hình 3.2: Detection complexities of DT-SM, Perfect Codes, SM-OSTBC $C(4,4,4)$, SM-DC, and DSTTD with $n_T = 4, n_R = 4$ at SNR $= 9$ dB and 8 bpcu; $T = 80$ symbol periods.



Hình 3.3: Detection complexities of DT-SM, SM-OSTBC $C(8,8,8)$ at SNR $= 9$ dB and 9 bpcu, and BOSTC at SNR $= 9$ dB and 8 bpcu; with $n_T = 8, n_R = 8$; $T = 80$ symbol periods.

SD, detection complexity of the DT-SM scheme is even lower than that of the SM-DC. Although, the DT-SM scheme has slightly higher complexity than the

where $Q(y) = \frac{1}{\sqrt{2\pi}} \int\limits_{y}^{\infty} e^{-\frac{x^2}{2}} dx$ is the Gaussian tail probability. PEP $P(\mathbf{C}_i \to \mathbf{C}_q)$ can be obtained by averaging the conditional PEP in the equation (3.8) over all realizations of $\mathbf{H}_{\text{corr}}$ is given as

$$P(\mathbf{C}_i \to \mathbf{C}_q) = \frac{1}{\pi} \int_0^{\frac{\pi}{2}} \Phi\left(-\frac{\gamma}{16E_s \sin^2\theta}\right) d\theta, \tag{3.9}$$

where $\Phi(s)$ is the moment generating function (MFG) of the random variable $d^2(\mathbf{C}_i, \mathbf{C}_q)$, and $P(\mathbf{C}_i \to \mathbf{C}_q)$ is evaluated at $s = -\frac{\gamma}{16E_s \sin^2\theta}$.

Through many transformations and counting $P(\mathbf{C}_i \to \mathbf{C}_q) = P(\mathbf{C}_q \to \mathbf{C}_i)$, $w_{i,q} = w_{q,i}$, và $w_{i,i} = 0$, we finally get

$$P_b \le \frac{2}{bN} \sum_{i=1}^{N-1} \sum_{q=i}^{N} \frac{w_{i,q}}{\pi} \int_0^{\frac{\pi}{2}} \prod_{l=1}^{2n_R} \left(1 + \frac{\gamma \kappa_{i,q,l}}{16E_s \sin^2\theta}\right)^{-1} d\theta. \tag{3.10}$$

3.5 Signal detection for the DT-SM system

For a given matrix $\mathbf{S}_k, k = 1, \ldots, K$, an $n_R \times 4$ equivalent matrix is constructed as $\tilde{\mathbf{H}}_k = \sqrt{\frac{\gamma}{4E_s}} \mathbf{H}\mathbf{S}_k$. Therefore, the system equation in (3.1) can be re-written as

$$\mathbf{Y} = \tilde{\mathbf{H}}_k \mathbf{X} + \mathbf{N}. \tag{3.11}$$

The structure of the DSTTD codeword $\mathbf{X}$ allows us to re-express (3.11) as

$$\mathbf{u} = \mathbf{H}_{e,k}\mathbf{x} + \mathbf{n}, \tag{3.12}$$

where $\mathbf{x} = \begin{bmatrix} x_1 & x_2 & x_3 & x_4 \end{bmatrix}^T$, $\mathbf{H}_{e,k} = \begin{bmatrix} \tilde{\mathbf{h}}_1 & \tilde{\mathbf{h}}_2 & \tilde{\mathbf{h}}_3 & \tilde{\mathbf{h}}_4 \\ \tilde{\mathbf{h}}_2^* & -\tilde{\mathbf{h}}_1^* & \tilde{\mathbf{h}}_4^* & -\tilde{\mathbf{h}}_3^* \end{bmatrix}$ with $\tilde{\mathbf{h}}_{\mathbf{q}}, q = 1, 2, .., 4$, is q-th column of $\tilde{\mathbf{H}}_k$; $\mathbf{u} = \begin{bmatrix} \mathbf{y}_1 & \mathbf{y}_2^* \end{bmatrix}^T$; $\mathbf{n} = \begin{bmatrix} \mathbf{n}_1 & \mathbf{n}_2^* \end{bmatrix}^T$; $\mathbf{y}_q$ and $\mathbf{n}_q, q = 1, 2$, respectively is q-th column of $\mathbf{Y}$ and $\mathbf{N}$. Converting the complex system equation (3.12) into the equivalent real-valued one, we have as given

$$\mathbf{v} = \mathbf{M}_k \mathbf{s} + \mathbf{w}, \tag{3.13}$$

where $\mathbf{M}_k = \begin{bmatrix} \Re(\mathbf{H}_{e,k}) & -\Im(\mathbf{H}_{e,k}) \\ \Im(\mathbf{H}_{e,k}) & \Re(\mathbf{H}_{e,k}) \end{bmatrix}$, $\mathbf{s} = \begin{bmatrix} \Re(\mathbf{x}) & \Im(\mathbf{x}) \end{bmatrix}^T$, $\mathbf{v} = \begin{bmatrix} \Re(\mathbf{u}) & \Im(\mathbf{u}) \end{bmatrix}^T$, và $\mathbf{w} = \begin{bmatrix} \Re(\mathbf{n}) & \Im(\mathbf{n}) \end{bmatrix}^T$. Doing QR decomposition of $\mathbf{M}_k$, $\mathbf{M}_k = \mathbf{Q}_k \mathbf{R}_k$.

Multiplying (3.13) by $\mathbf{Q}_k^H$, we have

$$\mathbf{t}_k = \mathbf{R}_k \mathbf{s} + \bar{\mathbf{w}}_k, \tag{3.14}$$

Converting the equivalent equation, we obtain

$$\|\mathbf{t}_k - \mathbf{R}_k\mathbf{s}\|^2 = \|\mathbf{v} - \mathbf{M}_k\mathbf{s}\|^2 - \mathbf{v}^H\mathbf{v} + \mathbf{t}_k^H\mathbf{t}_k. \tag{3.15}$$

Thus, the ML decoding rule for $\mathbf{s}$ conditioned on $\mathbf{S}_k$ can be written as

$$\begin{aligned}(\hat{\mathbf{s}})_k &= \arg\min_{\mathbf{s}\in\Omega_N} \|\mathbf{v} - \mathbf{M}_k\mathbf{s}\|^2 \\ &= \arg\min_{\mathbf{s}\in\Omega_N} \|\mathbf{t}_k - \mathbf{R}_k\mathbf{s}\|^2 + \mathbf{v}^H\mathbf{v} - \mathbf{t}_k^H\mathbf{t}_k,\end{aligned} \tag{3.16}$$

where Ω_N is the set of integers corresponding to the M-QAM constellation.

Using all possible $(\hat{\mathbf{s}})_k$, for $k = 1, 2, \cdots, K$, the index $\hat{k}$ associated with the transmitted SC codeword $\mathbf{S}_k$ is determined as given

$$\hat{k} = \arg\min_{k=1,2,\cdots,K} \left\|\mathbf{t}_k - \mathbf{R}_k(\hat{\mathbf{s}})_k\right\|^2 + \mathbf{v}^H\mathbf{v} - \mathbf{t}_k^H\mathbf{t}_k. \tag{3.17}$$

After that, the signal vector $\hat{\mathbf{s}}$ corresponding to the transmitted signal vector $\mathbf{x}$ can be recovered as $\hat{\mathbf{s}} = (\hat{\mathbf{s}})_{\hat{k}}$. The modified SD algorithm is detailed in Table 3.2.

3.6 Complexity analysis

The complexity of the pre-processing stage ρ_{Pre}, include those of computing $\tilde{\mathbf{H}}_k$ in (3.11), decomposition of $\mathbf{M}_k$ in (3.13), and signal vector $\mathbf{t}_k$, $k = 1, 2, ..., K$, in (3.14) is calculated as given

$$\rho_{\text{Pre}} = \left[\frac{2}{T}(32n_Rn_T + 536n_R - 36) + (64n_R - 8)\right]K. \tag{3.18}$$

Therefore, the SD complexity is presented as given

$$\rho = \frac{\rho_{\text{Pre}} + \rho_s}{l + 4m}, \tag{3.19}$$

where ρ_s of average complexity on the DT-SM codewords in the searching state (from Step 3 to Step 11).

Fig. 3.2 compares the detection complexity of the DT-SM with that of related MIMO schemes such as SM-OSTBC, DSTTD, Perfect Codes, SM-DC using . All schemes are compared in a $(4, 4)$ MIMO configuration and are implemented the SD detector at the same spectral efficiency. Fig. 3.2 shows that the proposed SD enables the DT-SM scheme to substantially reduce detection complexity compared with the ML decoder. With the aid of the proposed

Bảng 3.2: The modified SE-SD detection algorithm

Step	**Performance**
(1)	Set $k = 1$, $D_{\min} = C_0$, $\hat{k} = 0$, and $\hat{\mathbf{s}} = \emptyset$.
(2)	Compute $\tilde{\mathbf{H}}_k = \sqrt{\frac{\gamma}{4E_S}}\mathbf{H}\mathbf{S}_k$, generate $\mathbf{M}_k$, compute QR decomposition of $\mathbf{M}_k$ ($\mathbf{M}_k = \mathbf{QR}$) and $\mathbf{t} = \mathbf{Q}^H\mathbf{v}$.
(3)	Compute the sphere radius $R = D_{\min} - \mathbf{v}^H\mathbf{v} + \mathbf{t}^H\mathbf{t}$, and set $D_{\text{temp}} = \infty$.
(4)	If $R \le 0$, go to Step 11, else set $i = 8, T_8 = 0, \xi_8 = 0$.
(5)	(DFE[1] trên s_i) Set $s_i = \lfloor (t_i - \xi_i)/r_{i,i} \rceil$, and $\Delta = \text{sign}(t_i - \xi_i - r_{i,i}s_i)$.
(6)	(Main step) If $R < T_i + (t_i - \xi_i - r_{i,i}s_i)^2$, then go to Step 7. Else if $s_i \notin \Omega_N$ go to Step 9. Else if $i > 1$, then let $\xi_{i-1} = \sum_{j=i}^{8} r_{i-1,j}s_j$, $T_{i-1} = T_i + (t_i - \xi_i - r_{i,i}s_i)^2$, $i = i - 1$, go to Step 5. Else go to Step 8.
(7)	If $i = 8$, go to Step 10, else set $i = i + 1$ and go to Step 9.
(8)	Set $R = T_1 + (t_1 - \xi_1 - r_{1,1}s_1)^2$, $D_{\text{temp}} = R$, save $\check{\mathbf{s}} = \mathbf{s}$, and set $i = i + 1$.
(9)	Let $s_i = s_i + \Delta_i$, $\Delta_i = -\Delta_i - \text{sign}(\Delta_i)$, and go to Step 6.
(10)	Let $D_{\text{temp}} = D_{\text{temp}} + (\mathbf{v}^H\mathbf{v} - \mathbf{t}^H\mathbf{t})$. If $D_{\text{temp}} < D_{\min}$, then save $\hat{\mathbf{s}} = \check{\mathbf{s}}$ and $\hat{k} = k$, set $D_{\min} = D_{\text{temp}}$.
(11)	Set $k = k + 1$, if $k \le K$, goto Step 2.
(12)	If $\hat{\mathbf{s}} = \emptyset$ (i.e., no solution found), then increase C_0 then go to Step 1, else terminate.