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RESEARCH ARTICLE

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Error performance of zero-forcing beamforming with signal space diversity over correlated and uncorrelated Rician fading channels

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Abstract

In this manuscript, the error performance evaluation of multiple-input multiple-output (MIMO) method implementing zero-forcing beamforming (ZFBF) technique along with signal space diversity (SSD) is investigated under Rician fading channels. MIMO systems can be defined as systems created by using multiple antennas on both the transmitter and receiver sides. MIMO systems can enhance spectral efficiency and improve communication reliability. They hold significant importance for current 5G systems and ongoing research of 6G systems. Rician fading channels are commonly encountered in wireless communication systems and include line of sight (LOS) along with non-LOS propagation paths. In this research, examinations are conducted on a $t \times r$ MIMO scenario (with t transmitter antennas and r receiver antennas), assuming that the channel is known only at the receiver side. The ZFBF technique is used at the receiver to boost the spectral efficiency, while SSD is utilized to enhance the system reliability. The performance of SSD over Rician fading channels is evaluated and it is observed that there is a noteworthy improvement in the bit error rate performance a gain of up to 5.5 dB is obtained. Even under transmitter antenna correlation, it is shown that this technique can provide an important performance enhancement given by approximately 2 dB gain without any additional computational complexity.

Keywords: Multiple-input multiple-output systems; zero-forcing beamforming; signal space diversity; Rician fading channels

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1. Introduction

In current wireless technology standards such as 5G, the secure and uncorrupted transmission of information is of great importance and the significance of studies on this subject is increasing day by day. Recently, researchers have been showing interest in enhancing the bit error rate (BER) performance of wireless transmission systems, as there is a rapidly rising request for high data rate (large spectral efficiency) transmission. As one of the fundamental technologies of contemporary networks, multiple-input multiple-output (MIMO) is pivotal in improving spectral efficiency and BER performance through beamforming and spatial multiplexing [1].

Zero-Forcing Beamforming (ZFBF) is a spatial multiplexing approach with reduced-complexity. Under certain conditions, it can decompose a MIMO matrix channel into min(t, r) parallel and independent sublayers for a system with t transmitter antennas and r receiver antennas. Therefore, ZFBF allows for complete spatial multiplexing benefit [2-7]. With ZFBF, the signal is amplified in a specific direction, while interference from other directions/users is blocked. Thus, the signal is delivered to the desired directions/users without being directed to unwanted directions/users. In ZFBF systems, each min(t,r) substream is encoded separately. A linear operation (spatial processing) is applied at the transmitter as a ZFBF precoder or at the receive side as zero forcing receive beamforming (ZFRBF) on the substreams. This way, interference between different substreams is avoided by utilizing the spatial dimensions provided by the matrix channel. In addition, the beamforming vectors used for space-time filtering in ZFBF are determined based on the columns of the Moore-Penrose inverse (or pseudoinverse) of the matrix corresponding to the MIMO channel [8]. ZFBF has two main pros. One of them is the ease of implementation and the other one is the absence of error propagation possibility between subchannels. However, ZFBF is often negatively affected from its cons, most importantly from its power inefficiency (or noise enhancement) problem due to the selection of beamforming weight vectors [9].

The performance of MIMO systems is remarkably degraded as a result of the multipath fading, which causes destructive effects on the transmitted signal [10]. Signal space diversity (SSD), which is also referred as modulation diversity in the literature, is resorted to reduce the BER of ZFBF under fading without causing additional complexity. With SSD, the signal constellation is first rotated, and then component interleaving (CI) is employed according to a certain interleaving strategy that optimizes the performance [11-13]. In [14-17], the uncorrelatedness between fading coefficients corresponding to the coordinates of the transmitted signals is ensured by transmitting the real and imaginary coordinates of every modulated signal point over distinct transmit antennas. SSD has been broadly studied for single-input single-output systems [8-13]. Particularly, assuming binary phase shift keying (BPSK), it has been shown that SSD does not increase decoding complexity in [11]. Also, MIMO systems are combined with SSD in [18].

With Rician fading, in addition to the line of sight (LOS) propagation path, there are multiple non-line of sight (NLOS) propagation paths between the transmitter and receiver. The Rician fading is typically described by the Rician K-factor. The K-factor represents the ratio between the powers of the direct path and multiple reflections and plays a significant role in determining the BER in Rician fading channels. In this study, we demonstrate the reduction in BER using SSD with ZFBF technique over Rician fading channels for MIMO systems. We assume that the channel state information (CSI) is accessible at the receive side only. It is demonstrated that without the need for any additional complexity, the performance is significantly improved with ZFBF and SSD. With multiple antennas the signals are accurately transmitted by mitigating potential losses through multiple paths as the real and imaginary components of the symbols are transmitted through distinct paths. The rotation angle for BPSK signal set is chosen as described in the literature [10]. It is demonstrated that the proposed scheme is able to highly improve the BER performance even when the transmitter antennas are correlated. We aim to show that integrating SSD to MIMO systems can result in a performance gain up to 5.5 dB with this study.

The remainder of the article is arranged as follows. The ZFBF-SSD system model is defined for 2×2 MIMO systems and $t \times r$ MIMO systems in Section 2. The scenario that involves correlated transmitter antennas is also provided. Then, Section 3 presents the numerical results. Finally, the conclusion of the study is presented in Section 4. The notation used in the article can be expressed as follows: $(.)^T$, $\|.\|$, j, $(.)^H$, $\Im\{.\}$, and $\Re\{.\}$ operators represent the transpose, absolute value, √-1, Hermitian transpose, the imaginary and real parts of a complex number, respectively. Capital bold letters are adopted to represent matrices, and small bold letters are utilized to refer column vectors.

2. System model

In this paper, we focus on a MIMO point-to-point transmission scenario, assuming more than one substreams are transmitted simultaneously. The numbers of transmit and receive antennas are respectively denoted by t and r. Hence, the discrete-time baseband received signal at the receive side can be expressed as:

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{n} \tag{1}$$

where, $\mathbf{H}$ stands for the matrix channel (of dimension $r \times t$). The (i,k)th entry of the $\mathbf{H}$ matrix represents the fading coefficient from the k-th transmitter antenna and the i-th receiver antenna. The vector $\mathbf{x}$ ($\mathbf{x} \in \mathbb{C}^{t\times 1}$) denotes a column vector whose entries are simultaneously transmitted from transmitter antennas. The objective is to send m symbols s_i (which are modulated independently) in a synchronous fashion. Here, $s_i = s_{iI} + j\, s_{iQ}$ (i belongs to the set $\{1, 2, ..., m\}$), where s_{iI} and s_{iQ} respectively stand for the in-phase (I) and quadrature (Q) coordinates of the symbol s_i. Also, m represents the number of substreams transmitted simultaneously. The modulated symbols come from a rotated BPSK signal constellation. The rotation of the BPSK signal constellation causes each symbol to have non-zero I and Q components. There is a power constraint per symbol, expressed as $|s_i|^2 = E_b$ (E_b denotes the energy per bit). The noise vector $\mathbf{n}$ stands for the additive white Gaussian noise (AWGN) at the receive side ($\mathbf{n} \in \mathbb{C}^{r\times 1}$). The components of the $\mathbf{H}$ matrix are assumed to be independent and identically distributed (IID) Gaussian (complex) random variates with a variance of one and with a non-zero mean. The CSI is assumed to be accessible only at the receive terminal. Additionally, it is accepted that the fading components stay constant during the whole of a frame transmission and change from one transmission to another in an uncorrelated fashion.

2.1. ZFBF with SSD for 2×2 MIMO systems

First, we consider the scenario with t=r=2 (2 × 2 MIMO) as shown in Fig. 1 [9]. The bits are modulated according to a 45 degree counter-clockwise rotated BPSK scheme. Following that, CI is applied onto the modulated symbols regarding a certain CI strategy that optimizes the performance. Here, with the CI strategy, the real portions of two modulated signals s_1 and s_2 are merged as β_1 and can be written as $\beta_1 = s_{1I} + j\, s_{2I}$. Also, the imaginary parts of s_1 and s_2 are combined as β_2 and can be written as $\beta_2 = s_{1Q} + j\, s_{2Q}$. The interleaved symbols, β_1 and β_2 are passed through a digital-to-analog converter (DAC) to form the transmitted signal. The baseband transmitted vector signal can be represented as $\mathbf{x} = [\beta_1\, \beta_2]^T$.

The received discrete-time baseband signal can be expressed as given in (1). Additionally, considering that the columns of the $\mathbf{H}$ channel matrix are respectively denoted by $\mathbf{h}_1, \mathbf{h}_2 \in \mathbb{C}^{2\times 2}$, we can write:

$$\boldsymbol{y} = \boldsymbol{h}_1\beta_1 + \boldsymbol{h}_2\beta_2 + \boldsymbol{n} \tag{2}$$

The receive side uses ZFRBF to force inter-symbol interference to zero. Thus, the projection of the received discrete-time baseband signal $\mathbf{y}$ ($\mathbf{y} \in \mathbb{C}^{2\times 1}$) onto the vector $\mathbf{h}_1^H \mathbf{P}_{\mathbf{h}_2}^{\perp} \in \mathbb{C}^{1\times 2}$ is initially computed. Here, $\mathbf{P}_{\mathbf{h}_i}^{\perp}$ stands for the projection operation onto the orthogonal space of the vector $\mathbf{h}_i$.

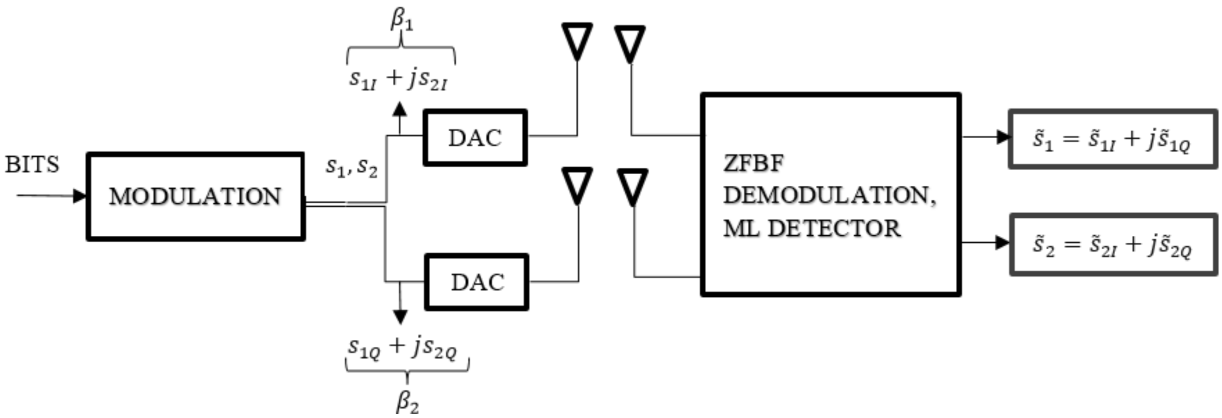

Fig.1. The block diagram of ZFBF with SSD in a 2×2 MIMO scenario.

The resulting equation can be stated as follows:

$$\begin{aligned}\frac{\mathbf{h}_1^H \mathbf{P}_{\mathbf{h}_2}^{\perp}}{\|\mathbf{h}_1^H \mathbf{P}_{\mathbf{h}_2}^{\perp}\|}\mathbf{y} &= \|\mathbf{h}_1^H \mathbf{P}_{\mathbf{h}_2}^{\perp}\| s_1 + \frac{\mathbf{h}_1^H \mathbf{P}_{\mathbf{h}_2}^{\perp}}{\|\mathbf{h}_1^H \mathbf{P}_{\mathbf{h}_2}^{\perp}\|}\mathbf{n} \\ &= \|\mathbf{h}_1^H \mathbf{P}_{\mathbf{h}_2}^{\perp}\| s_{1I} + \Re\left\{\frac{\mathbf{h}_1^H \mathbf{P}_{\mathbf{h}_2}^{\perp}}{\|\mathbf{h}_1^H \mathbf{P}_{\mathbf{h}_2}^{\perp}\|}\mathbf{n}\right\} + j\left(\|\mathbf{h}_1^H \mathbf{P}_{\mathbf{h}_2}^{\perp}\| s_{2I} + \Im\left\{\frac{\mathbf{h}_1^H \mathbf{P}_{\mathbf{h}_2}^{\perp}}{\|\mathbf{h}_1^H \mathbf{P}_{\mathbf{h}_2}^{\perp}\|}\mathbf{n}\right\}\right).\end{aligned} \tag{3}$$

Subsequently, $\mathbf{y}$ is projected onto the vector $\mathbf{h}_2^H \mathbf{P}_{\mathbf{h}_1}^{\perp} \in \mathbb{C}^{1\times 2}$. In this case, the following equality is obtained:

$$\frac{\mathbf{h}_2^H \mathbf{P}_{\mathbf{h}_1}^{\perp}}{\|\mathbf{h}_2^H \mathbf{P}_{\mathbf{h}_1}^{\perp}\|}\mathbf{y} = \|\mathbf{h}_2^H \mathbf{P}_{\mathbf{h}_1}^{\perp}\| s_{1Q} + \Re\left\{\frac{\mathbf{h}_2^H \mathbf{P}_{\mathbf{h}_1}^{\perp}}{\|\mathbf{h}_2^H \mathbf{P}_{\mathbf{h}_1}^{\perp}\|}\mathbf{n}\right\} + j\left(\|\mathbf{h}_2^H \mathbf{P}_{\mathbf{h}_1}^{\perp}\| s_{2Q} + \Im\left\{\frac{\mathbf{h}_2^H \mathbf{P}_{\mathbf{h}_1}^{\perp}}{\|\mathbf{h}_2^H \mathbf{P}_{\mathbf{h}_1}^{\perp}\|}\mathbf{n}\right\}\right). \tag{4}$$

The maximum likelihood (ML) detector relies on the following decision variables:

$$d_1 = \Re\left\{\frac{\mathbf{h}_1^H \mathbf{P}_{\mathbf{h}_2}^{\perp}}{\|\mathbf{h}_1^H \mathbf{P}_{\mathbf{h}_2}^{\perp}\|}\mathbf{y}\right\} + j\Re\left\{\frac{\mathbf{h}_2^H \mathbf{P}_{\mathbf{h}_1}^{\perp}}{\|\mathbf{h}_2^H \mathbf{P}_{\mathbf{h}_1}^{\perp}\|}\mathbf{y}\right\}. \tag{5}$$

and [10]

$$d_2 = \Im\left\{\frac{\mathbf{h}_1^H \mathbf{P}_{\mathbf{h}_2}^{\perp}}{\|\mathbf{h}_1^H \mathbf{P}_{\mathbf{h}_2}^{\perp}\|}\mathbf{y}\right\} + j\Im\left\{\frac{\mathbf{h}_2^H \mathbf{P}_{\mathbf{h}_1}^{\perp}}{\|\mathbf{h}_2^H \mathbf{P}_{\mathbf{h}_1}^{\perp}\|}\mathbf{y}\right\}. \tag{6}$$

2.2. ZFBF with SSD for $t \times r$ MIMO systems

In this section, we explore the MIMO scenario with t transmitter antennas and r receiver antennas ($2 \leq t \leq r$). The i-th entry of the **x** vector represents the signal conveyed from the i-th transmitter antenna before the DAC process, where i belongs to the set $\{1, 2, \ldots, t\}$. Here, the two parts of every modulated signal can be transmitted through distinct channels by exploiting the following transmitted baseband signal structure:

$$\mathbf{x} = \begin{bmatrix} s_{1I} + js_{2I} \\ s_{1Q} + js_{2Q} \\ s_{3I} + js_{4I} \\ s_{3Q} + js_{4Q} \\ \vdots \\ s_{t-1,I} + js_{tI} \\ s_{t-1,Q} + js_{tQ} \end{bmatrix}, \text{ for } t \text{ even} \tag{7}$$

$$\mathbf{x} = \begin{bmatrix} s_{tI} + js_{2I} \\ s_{1Q} + js_{2Q} \\ s_{3I} + js_{4I} \\ s_{3Q} + js_{4Q} \\ \vdots \\ s_{t-2,I} + js_{t-1,I} \\ s_{t-2,Q} + js_{t-1,Q} \\ s_{1I} + js_{tQ} \end{bmatrix}, \text{ for } t \text{ odd.} \tag{8}$$

The ML detector uses the following decision variables, if the number of transmitter antennas t is even:

$$d_i = \begin{cases} \Re\left\{\dfrac{\mathbf{h}_i^H \mathbf{P}_i^\perp}{\| \mathbf{h}_i^H \mathbf{P}_i^\perp \|}\mathbf{y}\right\} + j\Re\left\{\dfrac{\mathbf{h}_{i+1}^H \mathbf{P}_{i+1}^\perp}{\| \mathbf{h}_{i+1}^H \mathbf{P}_{i+1}^\perp \|}\mathbf{y}\right\} & \text{, for } i \text{ odd,} \\ \Im\left\{\dfrac{\mathbf{h}_{i-1}^H \mathbf{P}_{i-1}^\perp}{\| \mathbf{h}_{i-1}^H \mathbf{P}_{i-1}^\perp \|}\mathbf{y}\right\} + j\Im\left\{\dfrac{\mathbf{h}_i^H \mathbf{P}_i^\perp}{\| \mathbf{h}_i^H \mathbf{P}_i^\perp \|}\mathbf{y}\right\} & \text{, for } i \text{ even} \end{cases} \tag{9}$$

and as follows, if the number of transmitter antennas t is odd:

$$d_i = \begin{cases} \Re\left\{\dfrac{\mathbf{h}_t^H \mathbf{P}_t^\perp}{\| \mathbf{h}_t^H \mathbf{P}_t^\perp \|}\mathbf{y}\right\} + j\Re\left\{\dfrac{\mathbf{h}_2^H \mathbf{P}_2^\perp}{\| \mathbf{h}_2^H \mathbf{P}_2^\perp \|}\mathbf{y}\right\} & \text{, for } i=1, \\ \Re\left\{\dfrac{\mathbf{h}_i^H \mathbf{P}_i^\perp}{\| \mathbf{h}_i^H \mathbf{P}_i^\perp \|}\mathbf{y}\right\} + j\Re\left\{\dfrac{\mathbf{h}_{i+1}^H \mathbf{P}_{i+1}^\perp}{\| \mathbf{h}_{i+1}^H \mathbf{P}_{i+1}^\perp \|}\mathbf{y}\right\} & \text{, for odd } i \text{ together with } 1<i<t, \\ \Im\left\{\dfrac{\mathbf{h}_{i-1}^H \mathbf{P}_{i-1}^\perp}{\| \mathbf{h}_{i-1}^H \mathbf{P}_{i-1}^\perp \|}\mathbf{y}\right\} + j\Im\left\{\dfrac{\mathbf{h}_i^H \mathbf{P}_i^\perp}{\| \mathbf{h}_i^H \mathbf{P}_i^\perp \|}\mathbf{y}\right\} & \text{, for even } i \text{ together with } 1<i<t, \\ \Re\left\{\dfrac{\mathbf{h}_1^H \mathbf{P}_1^\perp}{\| \mathbf{h}_1^H \mathbf{P}_1^\perp \|}\mathbf{y}\right\} + j\Im\left\{\dfrac{\mathbf{h}_t^H \mathbf{P}_t^\perp}{\| \mathbf{h}_t^H \mathbf{P}_t^\perp \|}\mathbf{y}\right\} & \text{, for } i=t \end{cases} \quad (10)$$

As in Section 2.1., the same notation can be embraced for the definitions of the $\mathbf{h}_i$ vectors where i comes from the set $\{1, 2, \ldots, t\}$ (here, $\mathbf{h}_i \in \mathbb{C}^{r\times 1}$, is the i-th column of the $\mathbf{H}$ channel matrix). Additionally, the projection matrix onto the orthogonal space of the set of the vectors $\mathbf{h}_1, \mathbf{h}_2, \ldots \mathbf{h}_{i-1}, \mathbf{h}_{i+1}, \mathbf{h}_{i+2}, \ldots \mathbf{h}_t$ is reprensented by the expression $\mathbf{P}_i^\perp$.

2.3. Correlated transmitter antennas scenario

In this subsection, we consider a scenario in which we have correlated transmit antennas. In this case:

$$\mathbf{H} = \tilde{\mathbf{H}}\mathbf{R}^{1/2} \quad (11)$$

where, $\mathbf{R}$ is a square matrix with dimensions t-by-t and encapsulates the correlation effect within the transmitter antennas. The components of the matrix $\tilde{\mathbf{H}}$ in (11) are IID zero-mean Gaussian (complex) random variates with a variance that is equal to one. In our work, a type of correlation matrix structure, namely uniform correlation model is used. The R matrix for uniform correlation model is given as:

$$\mathbf{R} = \begin{bmatrix} 1 & \rho & \rho & \rho & \cdots & \rho & \rho \\ \rho & 1 & \rho & \rho & \cdots & \rho & \rho \\ \vdots & \vdots & \vdots & \vdots & \ddots & \vdots & \vdots \\ \rho & \rho & \rho & \rho & \cdots & \rho & 1 \end{bmatrix} \quad (12)$$

where, ρ is the correlation coefficient between any two distinct transmitter antennas [19].

3. Numerical results

The BER performance of the introduced scheme, which has a BPSK signal constellation rotated counterclockwise by 45 degrees, is illustrated in this section. A number of comparisons to the standard ZFBF scheme without SSD are shown for 3×3, 4×6, and 5×9 MIMO systems and the results are depicted in the figures Fig. 2, Fig. 3, and Fig. 4, respectively. In all the scenarios the significant impact of SSD can be observed.

In Fig. 2 we can see that, a gain of approximately 5.44 dB is obtained at a BER of 10^{-4} when K=5 for 3 × 3 MIMO scenario (For an SNR of 20 dB, the BER reduces to 0.0045 with the adoption of the proposed scheme as compared to the BER of the standard approach which equals 0.01). Also, a gain of approximately 4.32 dB is acquired at a BER of 10^{-4} when K=10 (For an SNR of 20 dB, the BER reduces to 0.008 with the adoption of the proposed scheme as compared to the BER of the standard approach which equals 0.019).

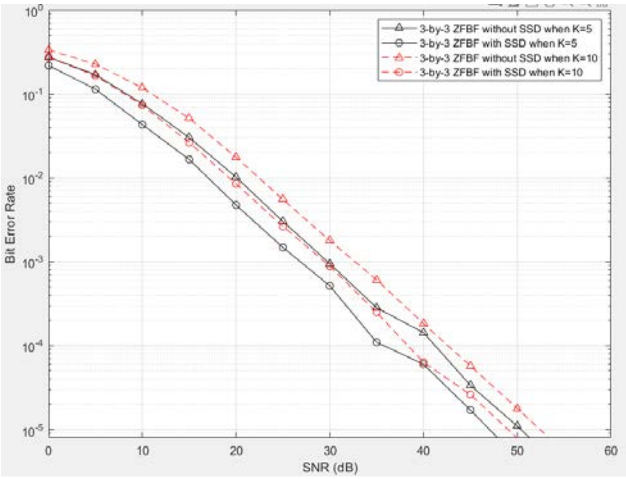

Fig.2. 3-by-3 ZFBF scenario with and without SSD, when K=5 and K=10.

Fig. 3 indicates that a gain of approximately 1.6 dB is achieved at a BER of 10^{-4} when K=5 for 4 × 6 MIMO scenario. In addition, at a BER of 10^{-4}, a gain around 1.67 dB is obtained when K=10 for 4 × 6 MIMO scenario.

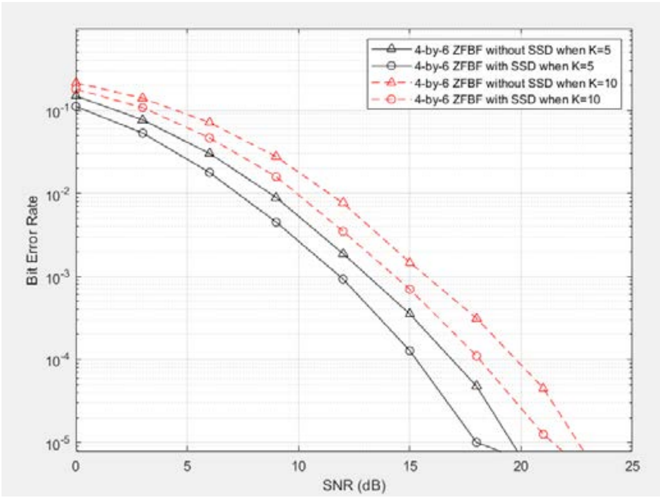

Fig.3. 4-by-6 ZFBF scenario with and without SSD, when K=5 and K=10.

Fig. 4 indicates that a gain of about 1.11 dB is attained at a BER of 10^{-4} when K=5 for 5×9 MIMO scenario. Besides, a gain of approximately 1.27 dB can be accomplished at a BER of 10^{-4} when K=10 for 5×9 MIMO scenario.

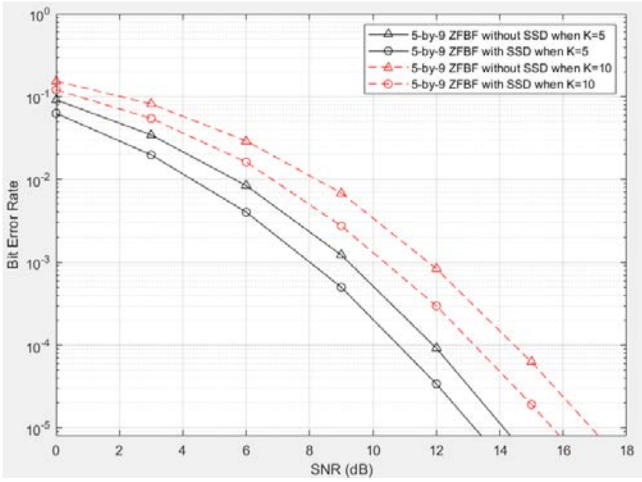

Fig.4. 5-by-9 ZFBF scenario with and without SSD, when K=5 and K=10.

In Fig. 5, the suggested technique with SSD and the original model with no SSD are compared under uniform transmitter antenna correlation. The correlation coefficient value (ρ) is set to be equal to 0.4. It can be seen from Fig. 5 that even under transmitter antenna correlation, a gain of approximately 1.86 dB can be obtained at a BER of 10^{-4} when $K = 5$ for 5×9 MIMO scenario. Furthermore, under transmitter antenna correlation, a gain around 1.41 dB can be attained at a BER of 10^{-4} when $K = 10$ for 5×9 MIMO scenario.

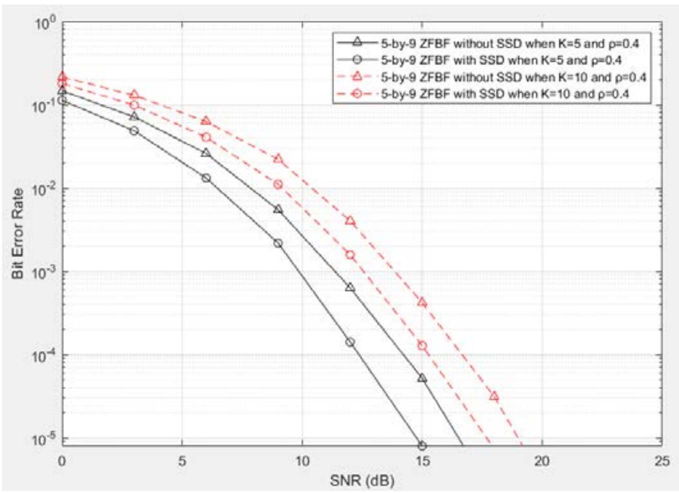

Fig.5. For ρ =0.4, 5-by-9 ZFBF scenario with and without SSD, when K=5 and K=10.

4. Conclusion

In this study, it has been shown that significant BER improvement over Rician fading channels can be achieved by combining ZFBF with SSD. What is more, this advancement occurs without any need for increased spectrum or time slot usage, resulting in a negligible growth in complexity. The BPSK signal constellation is rotated 45 degrees counterclockwise and the I and Q coordinates of the modulated signals are interleaved. In our work, both correlated and uncorrelated transmitter antennas have been examined. Uniform correlation matrix structure is adopted in correlated antennas scenario. The proposed scheme can also be used for multi-user scenarios with multiple receivers and transmitters, such as in a cellular network. In addition, the presented technique can be applied to quadrature amplitude and phase-shift keying modulations with a chosen modulation level that is greater than two.

Author Contribution

All authors contributed to the study conception and design. Material preparation, data collection, and analysis were performed by Hilal Uslu, M. Anıl Reşat and Serdar Özyurt. All authors read and approved the final manuscript.

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