Analog- Digital Precoder for Downlink System on the Impact of Antenna Impairments

Wasan Al- Masoody* and Naz Islam*

Abstract- In this paper, the BER performance of the downlink multi-user multiple-input-multiple-output (MU- MISO) system for the perfect channel state information (CSI) is investigated. Rather than compensating or eliminating the effect of the mutual coupling among antenna elements, an optimization process is used by the aid of an analog-digital (AD) precoder. In this precoder, a conventional linear scheme is used in the digital domain, while a standard optimization technique is applied in the analog domain to manipulate the values of the load impedance and adjust the source of the antenna impairment which is the mutual coupling (MC) leading to the optimum values. The results are compared with the ideal case, where there is no mutual coupling, from one hand and with the conventional case, where there is mutual coupling. Although the results from the ideal case are much better than the results from the conventional one, the superior results can be achieved using the AD precoder.

Index Terms— MIMO, BER, precoding, mutual coupling, analog digital precoding, perfect CSI, optimization.

1 Introduction

IN wireless communication systems, the applications of multiple-input multiple-output (MIMO) have resulted in performance gains over the conventional single-input single-output (SISO) systems. The precoding techniques associated with MIMO can transfer the computational complexity from the user side to the base station is more suitable for such studies and have been

extensively studied [1]. For example, the capacity for achieving dirty paper coding (DPC) has been proposed to pre-subtract the interference before transmission [2]. However, due to the DPC's impractical assumption and high computational complexity, this technique is difficult to implement [2]. On the other hand, suboptimal non-linear technique such as Tomlinson-Harshima precoding (THP) and vector perturbation (VP) have also been proposed [3]- [5]. However, the most common precoding approaches are the linear precoding which has been receiving increasing research attention due to its low computational complexity. Besides, the zero-forcing (ZF) precoding is a simple and effective technique with tolerable sub-optimal performance with a significant computational complexity reduction [6], [7]. C. B. Peel et al. in [8] developed a regularized form of ZF (RZF), by introducing a regularization factor, which improved performance, especially at low signal-to-noise ratios (SNRs). On the other hand, a correlation rotation linear scheme (also known as phase alignment) was proposed in [9], [10], where the interference has been exploited to further benefit the system performance. It was also shown that as the number of antennas increases, the computational complexity and cost for the non-linear approaches could be very high although they can provide rate benefits for the MIMO system [11-13]. Therefore, due to the complexity benefits of linear approaches over non- linear ones, the focus of this work will be on linear precoding schemes.

Existing studies on the precoding scheme of MIMO systems usually assume an uncorrelated Rayleigh flat fading channel, which means there is no spatial correlation or mutual coupling impact among antenna elements. However, in practice when the antenna spacing is small, the spatial correlation and mutual coupling effects cannot be neglected [12], [13], [18- 21]. Therefore, many experimental studies have been conducted to investigate the correlation and mutual coupling. The effect of spatial correlation and mutual coupling is examined when a large number of antenna elements are fitted within a fixed physical space [18], [19].

Spatial correlation can be interpreted as a correlation between the received average signal gain and the spatial direction of the signal. Many experimental studies have been conducted on the spatial correlation effect [20], [22], [23]. The impact of the spatial correlation on the system performance of MIMO has also been investigated earlier [24-28]. The designs of the robust precoding scheme in spatially correlated channel were also studied [29-31]. The effect of mutual coupling induced by two real-world antennas has been analyzed. Matrix

ISSN 2229-5518

derivation of the mutual coupling based on a 2×2 MIMO system and its effect on the MIMO capacity is also investigated [32]. Many studies have investigated the effect of the mutual coupling on the system performance and shown that the existence of mutual coupling can degrade the detection performance [33-36]. The impact of mutual coupling between antenna elements on the outage capacity of a 2×2 MIMO system in flat fading channels is studied. The effect of mutual coupling between antenna elements of adaptive array and its performance is studied in [37]. The performance of the adaptive arrays is affected by the mutual coupling even for large interelement spacing and this effect increases when the inter-spacing between antenna elements decreases. In [36], both the positive and negative effects of mutual coupling is studied for MIMO systems at high SNR. In order to alleviate this performance loss, many compensation techniques for the effects of mutual coupling have been proposed [38-40], which are essentially based on the derivation of the compensation matrix. A compensation technique for mutual coupling in a small antenna array was developed and verified experimentally by Steyskal and Herd [38]. In this work, the compensation is realized by forming a matrix, which is multiplied by the received signal vector and the effectiveness of this scheme is also validated [ibid]. Corcoles and et. al. [39] further introduced a flexible method to compensate the mutual coupling effects that degrades the field pattern of a real antenna array is proposed. Here, the mutual coupling compensation matrix is derived and calculated from the generalized scattering matrix of the antenna array and the spherical mode expansion of its radiate field. The introduction of a general method to obtain the compensation matrix of mutual coupling effects in transmitting arrays for the total field in all direction is discussed by Rubio et. al [40]. It was also found that there is a simple relationship between the compensation matrices of the transmitting and the receiving arrays. In literature, many novel structures have been proposed to eliminate the mutual coupling effects [41-43]. The concept of the mantle cloaking method to reduce the mutual coupling effect between strip dipole antennas at low- terahertz (THz) frequencies is proposed in [41]. Here, it was shown that the electromagnetic interaction between the antennas can be suppressed by covering each antenna with an elliptically shaped graphene monolayer. A novel structure suppressing the mutual coupling is also proposed by Farsi et.al. in [42]. To reduce the effect of mutual coupling between antenna elements, a simple U- shaped microstrip is composed [ibid]. Li et. al. in [43] further introduces the use of parasitic elements to reduce the mutual coupling effect is studied. By adding parasitic elements, a double- coupling path is introduced and it can create a reverse coupling to reduce the mutual coupling effect of MIMO antennas. Other techniques that target at the mutual coupling compensation are found in [44- 46]. Most of the above cited researches, however, are not from the signal processing perspective. In this work, we construct an analog-digital (AD) precoding scheme that can exploit the mutual coupling effect in order to further improve the system performance. The problem is formulated into convex optimization to obtain the optimal beamforming vectors and load impedance value for each antenna array. In the proposed scheme, by equipping each antenna element with tunable load impedance, the mutual coupling effect can be controlled by tuning the value for each load impedance. Here, the conventional linear precoding is applied in the digital domain, while convex optimization is applied in the analog domain. In addition, the optimization problem is formulated as a standard least square problem with convex constraints which can be solved by existing methods and algorithms such as primal-dual interior-point method [47]. Moreover, simulation results for both systems, the one with the existing mutual coupling and the other with the proposed scheme, are compared to the ideal case, that is, a system with no mutual coupling. As a result of our approach, an improved detection performance can be expected by using various optimization methods and algorithms.

Notations: Throughout this paper, a, a, and A denote scaler, vector, and matrix respectively. $\mathrm{E}\{\cdot\}$, $(\cdot)^T$, $(\cdot)^H$, $(\cdot)^{-1}$, and $tr(\cdot)$ denote expectation, transpose, conjugate transpose, inverse, and trace of a matrix respectively. $\|\cdot\|$ denotes the Frobenius norm and I is the identity matrix. The conversion of a vector into a diagonal matrix, the vector values on its main diagonal is represented by $diag(\cdot)$. $C^{n \times n}$ represents $n \times n$ matrix in the complex set. Finally, $\Re(\cdot)$ and $\Im(\cdot)$ denote the real part and imaginary part of a complex number, respectively.

Naz E. Islam is a professor with Electrical and Computer Engineering Department, University of Missouri- Columbia, Columbia, MO 65201, USA. E- mail: (islamn@missouri.edu)
Wasan Al- Masoody is currently pursuing PhD program in Electrical and Computer Engineering at University of Missouri- Columbia, Columbia, MO 65201, USA. E- mail: (whahg5@mail.missouri.edu) University of Babylon, Babylon, Iraq

2. SYSTEM MODEL

We consider the model of a MU- MISO downlink system where a base station (BS) equipped with N_t antennas are designed to communicate with K users simultaneously, as shown in Fig. 1 below, where $\leq N_t$.

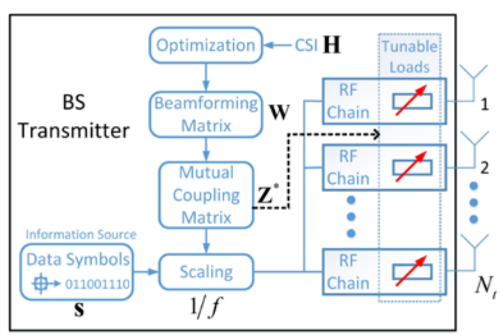

Fig. 1 BS Transmit Structure

The transmit symbol vector is processed with the precoding matrix F at the BS, and the received signal vector can be obtained from equation (1) below

$$\mathbf{y} = \mathbf{Hx} + \mathbf{n} = \mathbf{HFs} + \mathbf{n} \qquad (1)$$

where $\boldsymbol{x} \in C^{K\times 1}$ and $\boldsymbol{y} \in C^{N_t\times 1}$ are the transmit and receive signal vectors respectively, $\mathbf{H} \in C^{K\times N_t}$ is the channel matrix, and $\boldsymbol{n} \in C^{K\times 1}$ is the noise vector whose elements are assumed to be the additive white gaussian noise (AWGN) with zero mean and variance σ^2. Indeed, by linearly precoding the data symbol vector $\boldsymbol{s} \in C^{K\times 1}$, and with the assumption that $\mathrm{E}\{\mathbf{ss}^H\} = \mathbf{I}$, with the precoding matrix $\mathbf{F} \in C^{N_t\times K}$ results in the precoded transmit signal x. The precoded signal is then given by $\boldsymbol{x} = \frac{1}{f}.\mathbf{Fs}$ and it includes the noise amplification factor f, and thereby guaranteeing that the average transmit power is not change after transmission. Here, $\mathbf{Z}$ is the mutual coupling matrix at the transmit side whose details will be given in the section that follow.

2.1 Spatially Correlated Channel Model

We shall next consider the model of a MU- MISO downlink system in designing the precoder. As stated earlier, ideal antenna arrays are assumed in most existing works, which suggest that no spatial correlation among antenna elements is considered [48], [49]. However, in practice when the antenna spacing is small, the effect of spatial correlation must be taken into account in the channel model. Therefore, a semi-correlated geometric non- line of sight (NLOS) Rayleigh flat fading channel is applied, where the spatial correlation is considered at the transmitter side (BS) [11], [21]. We shall model the channel as

$$\mathbf{H} = [\mathbf{h}_1^T, \mathbf{h}_2^T, \ldots\ldots, \mathbf{h}_K^T]^T \qquad (2)$$

Where, $\mathbf{h}_K \in \mathcal{C}^{1\times N_t}$ is the channel vector for user k, and can be expressed through equation (3) below, based on [47], [11], [21].

$$\mathbf{h}_K = \mathrm{g}_K \mathbf{A}_K \mathbf{Z} \qquad (3)$$

Here each element in $g_K \sim \mathcal{CN}(0,1)$ forms the Rayleigh component and follows the standard complex Gaussian distribution. The transmit- side steering matrix $A_K \in \mathcal{C}^{M\times N_t}$ contains M steering vectors of transmit antenna arrays, where M represents the number of direction of departure (DoDs). Assuming uniform linear arrays (ULAs) in this work, A_K can then be expressed as

$$\mathbf{A}_K = \frac{1}{\sqrt{M}} \left[\mathbf{a}^T(\phi_{k,1}), \mathbf{a}^T(\phi_{k,2}), \ldots\ldots, \mathbf{a}^T(\phi_{k,M})\right]^T \qquad (4)$$

where $\mathbf{a}(\phi_{k,i}) \in \mathcal{C}^{1\times N_t}$ is given by

$$\mathbf{a}(\phi_{k,i}) = [1, e^{j2\pi d sin\phi_{k,i}}, \ldots\ldots, e^{j2\pi(N_t-1)d sin\phi_{k,i}}] \qquad (5)$$

(1)

where d in equation (5) is the spacing between each antenna element normalized by the carrier wavelength, and $\emptyset_{k,i}$ denotes the angles of departure (AoDs) which is assumed to be randomly and independently distributed in $[-\Phi, \Phi]$ with a uniform distribution.

2.2 Modeling the Mutual Coupling Effect

Based on [11], [32], the mutual coupling matrix with tunable load impedance can be derived as in [48], [49]: Where Z, can be represented as

$$\mathbf{Z} = [z_A.\mathbf{I} + diag(\mathbf{z}_L)][\mathbf{\Gamma} + diag(\mathbf{z}_L)]^{-1} \qquad (6)$$

where z_A is the antenna impedance that is assumed constant and $\mathbf{z}_L = \left[z_{L_1}, z_{L_2}, \ldots.. z_{L_{N_t}},\right]^T$, is the load impedance vector we are going to optimize. $\mathbf{\Gamma}$, is the mutual impedance matrix that is defined as follows

$$\mathbf{\Gamma} = \begin{bmatrix} z_A & z_{m_1} & z_{m_2} & \cdots & z_{m_{N_t-1}} \\ z_{m_1} & z_A & z_{m_1} & \ddots & \vdots \\ z_{m_2} & z_{m_1} & \ddots & \ddots & z_{m_2} \\ \vdots & \vdots & \ddots & \ddots & z_{m_1} \\ z_{m_{N_t-1}} & \cdots & z_{m_1} & z_{m_1} & z_A \end{bmatrix} \qquad (7)$$

where z_{m_k} is the mutual impedance of two antenna elements with a distance of k.d. Based on Chapter 8 of [34], the value of z_A and z_{m_k} can be found by the electromagnetic- field (EMF) method based on a transmit antenna spacing of d.

3 Proposed Analog- Digital Precoding Scheme

We introduce the proposed scheme based on the linear ZF precoder. In this scheme the mutual coupling effect can be controlled by equipping each antenna element with a varactor as load impedance. Indeed, by optimizing the value of each load impedance will lead to minimizing the noise amplification factor of the proposed precoder [49]. To further

ISSN 2229-5518

exploit the mutual coupling effect, we first rewrite the channel matrix as $= \mathbf{WZ}$, where W is given by

$$\mathbf{W} = [\mathbf{w}_1^T, \mathbf{w}_2^T, \ldots\ldots, \mathbf{w}_K^T]^T \tag{8}$$

and $\mathbf{w}_K = \mathbf{g}_K \mathbf{A}_K$. Then (1) can be rewritten as

$$\mathbf{y} = \mathbf{WZFs} + \mathbf{n} \tag{9}$$

Based on (9) we construct the proposed precoder F as

$$\mathbf{F} = \frac{1}{f} \,.\, \mathbf{Z}^{-1}\mathbf{G} \tag{10}$$

where following the formulation of closed-form beamformers $f = \|\mathbf{Z}^{-1}\mathbf{G}\|$ is the scaling factor that ensures the signal power is not changing during the beamforming, and (9) is further transformed into

$$\mathbf{y} = \frac{1}{f} \,.\, \mathbf{WGs} + \mathbf{n} \tag{11}$$

As can be noticed from (11), with this precoding structure the mutual coupling impact can be fully eliminated in the channel, while it still has an impact on the system performance, which is characterized by the resulting scaling factor f .

Based on the concept of the ZF and to fully eliminate the multi-user interference we reconstruct the beamformer $\mathbf{G} = \mathbf{W}^H(\mathbf{W}\mathbf{W}^H)^{-1}$. Then the precoded signal vector x can be obtained as

$$\mathbf{x} = \frac{1}{f} \,.\, \mathbf{Fs} = \frac{1}{f}\ \mathbf{Z}^{-1}\mathbf{Gs} \tag{12}$$

The received signal vector can be obtained at the user side, by substituting (12) into (1), is

$$\mathbf{y} = \frac{1}{f}\, \mathbf{WZ}\,\mathbf{Z}^{-1}\mathbf{G}\mathrm{s} + \boldsymbol{n} = \frac{1}{f} \cdot \mathbf{s} + \boldsymbol{n} \tag{13}$$

Then, the received signal vector needs to be scaled back before demodulation to eliminate the scaling factor f to be $\mathbf{r} = f.\mathbf{y} = \mathbf{s} + f.\boldsymbol{n}$ as a new rescaled vector.

Although the mutual coupling effect is fully eliminated with this precoder [49], it still has an effect on the precoding matrix and consequently on the noise amplification factor, f. therefore, we shall pursue the following optimization problem and a better system performance can be achieved by minimizing the noise amplification factor. This can be fulfilled by optimizing the value of each load impedance. This optimization problem can be expressed as

$$\text{P\_0}:\ \min_{z_L} \|\mathbf{Z}^{-1}\mathbf{G}\|^2 \tag{14}$$

Before proceeding our work, we shall study the inverse of the mutual coupling matrix. Based on (6), $\mathbf{Z}^{-1}$ can be derived as follows [48], [49]:

$$\begin{aligned}\mathbf{Z}^{-1} &= \{[z_A.\mathbf{I} + diag(\mathbf{z}_L)][\boldsymbol{\Gamma} + diag(\mathbf{z}_L)]^{-1}\}^{-1} \\ &= [\boldsymbol{\Gamma} + diag(\mathbf{z}_L)]\,.\,diag(\mathbf{z}_T)\end{aligned} \tag{15}$$

where

$$\mathbf{z}_T = \left[\frac{1}{z_1}, \frac{1}{z_2}, \ldots, \frac{1}{z_{N_t}}\right]^T, \text{ and}$$

$$z_i = z_A + z_{L_i} \tag{16}$$

Through expanding (15) further, $\mathbf{Z}^{-1}$ can be expressed as

$$\mathbf{Z}^{-1} = \begin{bmatrix} 1 & \frac{z_{m_1}}{z_2} & \frac{z_{m_2}}{z_3} & \cdots & \frac{z_{m_{N_t-1}}}{z_{N_t}} \\ \frac{z_{m_1}}{z_1} & 1 & \frac{z_{m_1}}{z_3} & \ddots & \vdots \\ \frac{z_{m_2}}{z_1} & \frac{z_{m_1}}{z_2} & \ddots & \ddots & \frac{z_{m_2}}{z_{N_t}} \\ \vdots & \vdots & \ddots & \ddots & Z_{m_1} \\ \frac{z_{m_{N_t-1}}}{z_1} & \cdots & \frac{z_{m_2}}{z_{N_{t-2}}} & \frac{z_{m_1}}{z_{N_{t-1}}} & 1 \end{bmatrix} \tag{17}$$

And by denoting $\boldsymbol{\delta}$ as

$$\begin{aligned}\boldsymbol{\delta} &= diag\left(\frac{z_{m_1}}{z_1}, \frac{z_{m_1}}{z_2}, \ldots, \frac{z_{m_1}}{z_{N_t}}\right) \\ &= diag\left(\alpha_1, \alpha_2, \ldots, \alpha_{N_t}\right)\end{aligned} \tag{18}$$

Resulting in the matrix can be written as

$$\mathbf{D} = \begin{bmatrix} 0 & 1 & \frac{z_{m_2}}{z_3} & \cdots & \frac{z_{m_{N_t-1}}}{z_{N_t}} \\ 1 & 0 & 1 & \ddots & \vdots \\ \frac{z_{m_2}}{z_1} & 1 & \ddots & \ddots & \frac{z_{m_2}}{z_{N_t}} \\ \vdots & \vdots & \ddots & \ddots & Z_{m_1} \\ \frac{z_{m_{N_t-1}}}{z_1} & \cdots & \frac{z_{m_2}}{z_{N_{t-2}}} & 1 & 0 \end{bmatrix} \tag{19}$$

and $\mathbf{Z}^{-1}$ can be expressed as

$$\mathbf{Z}^{-1} = \mathbf{D}\boldsymbol{\delta} + \mathbf{I} \tag{20}$$

By substituting (20) into (14), this optimization problem can be transformed into

$$\text{P\_I}:\quad \min_{\Theta} \|\mathbf{D}\boldsymbol{\delta}\mathbf{G} + \mathbf{G}\|_F^2 \tag{21}$$

P_I is considered a least square problem and our task now is how to solve it over the optimization variables. Indeed, this problem can be seen as a quadratic program (QP) and has a standard analytical solution [47]. In practice, when we implement the varactors as load impedance, the real part of the varactor should be positive such that the antenna array can radiate power [50], [51]. This feature adds to the constraint of the optimization problem P-I. Based on (16) and (18) each load impedance can be represented as a function of the variable α_i, which is given by

$$z_{Li} = \frac{z_{m1}}{\alpha_i} - z_A \tag{22}$$

and the real part of each one of the load impedances should be positive, as can be represented as

$$\mathcal{R}\,(z_{Li}) \geq 0\,, \quad \forall_i \in \{1,2,\dots,\mathrm{N}_t\} \tag{23}$$

which is also convex, where we notice that a practical antenna has $\mathcal{R}\,(z_A) > 0$ [34]. Therefore, by combining (21) and (23) the optimization problem can be formulated as

$$\begin{aligned} &\text{P\_IV} \;: \quad \min_{\boldsymbol{\Theta}} \|(\mathbf{D}\boldsymbol{\delta}\mathbf{G} + \mathbf{G})\|_F^2 \\ &\text{s.t.} \\ &\mathcal{R}\,(z_{Li}) \geq 0\,, \quad \forall_i \in \{1,2,\dots,\mathrm{N}_t\} \end{aligned} \tag{24}$$

Which is again a least square problem with convex constraints and can be considered as a QP with convex constraints. Following [47], this optimization problem is an inequality constrained minimization problem and can be efficiently solved by the interior-point methods such as the barrier method or the primal-dual interior-point method. indeed, the latter method is often more efficient than the former method, especially when high accuracy is required.

In practice, the optimization problem (24) is a standard least square problem with convex constraints and can be efficiently solved by using convex optimization tools such as CVX. Indeed, these software packages include all the techniques and algorithms that are applicable to solve the optimization problem analytically. in addition, these software tools contain numerical methods such as primal-dual interior-pint methods that can provide both the optimal primal variable and the optimal dual variables, i.e., Lagrange multiplier [47]. One only needs to reformulate the optimization problems to be similar to the existing standard forms. Then, each load impedance can be obtained by (24) and consequently, the resulting optimal mutual coupling matrix is obtained based on (20) as

$$\mathbf{Z}^* = (\mathbf{D}\boldsymbol{\delta}^* + \mathbf{I})^{-1} \tag{25}$$

More performance gain can be achieved when the regularized zero forcing (RZF) scheme is employed and it is simply done by expanding $\mathbf{G}$ in the ZF approach to be $\mathbf{G} = \mathbf{W}^H\left(\mathbf{W}\mathbf{W}^H + \frac{K}{\rho}\cdot\mathbf{I}\right)^{-1}$. Moreover, the proposed technique can be applied to other precoding approaches by simply substituting $\mathbf{G}$ with other precoders, and consequently the performance gain can still be achieved.

So far, we have shown that the optimal precoder can be achieved by solving (24) not only theoretically but also practically by CVX simulation. However, it should be pointed out that

the adaptation of each load impedance z_{Li}, depending on the channel variation, is required for the proposed scheme. This could be done through adaptive impedance approaches. There are many existing varactor technologies which support adaptive impedance tuning. Semiconductor-based varactor diodes, ferroelectric- based varactors, and microelectromechanical system (MEMS) varactors are the main categories of the varactor technologies [52]. In addition, semiconductor- based and ferroelectric- based varactors can support up to 1-100 ns tuning speed. Furthermore, adaptive matching networks are employed in [52] based on an automated impedance tuning unit with ferroelectric varactor diodes which are applicable to facilitate the proposed approach. Moreover, the electronically steerable parasitic array radiators (ESPARs) are applied which can further support the implementation of the proposed approach. In fact, the radiation pattern of (ESPARs) are formed by changing the values of each load impedance on a symbol -by-symbol basis [53-57]. Thus, based on practical aspects, the proposed precoder can be applied and is mostly suitable for slow or quasi-static fading channels as the it they change slowly.

4 Numerical Results

The performance of the proposed AD precoder with the aid of Monte Carlo simulation is evaluated. In this work, a MU_MISO system operating on a frequency of $f = 2.6$ GHz is assumed. The transmitter and receiver are designed with $N_t = 4, 6$ and $K = 4$ respectively. The distance between each antenna element is assumed to be $d = 0.25$ and the dipole length of each element is $l = 0.3$. The simulation of the channel as a Raleigh flat fading channel with a number of steering vectors of $M = 50$ and angle spread of $\Phi = \pi / 8$ is considered. Although the performance benefits are extending to other receives, our focus in this work will be on the linear ZF and RZF receivers.

TABLE I

ABBREVIATION OF DIFFERENT SCHEMES

Abbreviation	Scheme
ZF with no	ZF receiver with no mutual coupling
ZF with MC	ZF receiver with fixed mutual
RZF with no	RZF receiver with no mutual
RZF with MC	ZF receiver with fixed mutual
ZF with AD	ZF receiver with the proposed AD
RZF with AD	RZF receiver with the proposed AD

In Fig. 2, the bit error rate (BER) performance for the proposed precoder using QPSK

modulation is simulated with respect to the transmit SNR. In this figure, we compare the BER performance with conventional ZF technique, where there is no mutual coupling. As can be, the ideal ZF case, with no MC, is superior to the conventional ZF, with MC, and that the later achieves the worst performance. This is indeed due to the fixed mutual coupling among antenna elements. Compared to the previous two schemes, the proposed AD precoder, outperforms the previous schemes with an SNR gain over 4 dB. To elevate the system performance RZF is used instead of ZF. In this case much better improvements in the BER performance is achieved for all the previous schemes.

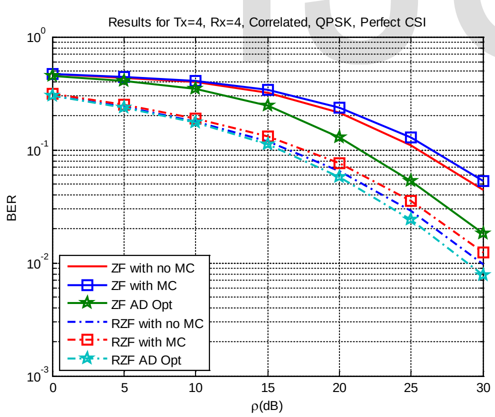

Fig.2, BER performance for the AD precoder with $N_t = 4, K = 4$,

QPSK modulation

In Fig. 3, The performance gain for all the above precoding schemes becomes larger when we increase the number of transmit antennas, but still this improvement at the expense of larger and more complex transmitter (BS). It can also be observed that the precoding schemes with fixed mutual coupling are the most effective ones, i.e., larger performance gain compared the previous scheme, with $N_t = 4$. All in all, the performance gain is expected to be larger when $N_t < K$.

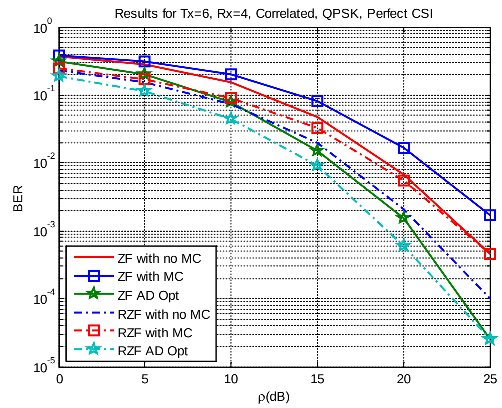

Fig.3, BER performance for the AD precoder with $N_t = 6, K = 4$,

QPSK modulation

5. CONCLUSION

In this paper, a joint analog- digital precoding scheme for the downlink multi-user multiple-input single-output MU-MISO system for the perfect CSI is proposed. Different from the ideal scheme, where a linear precoding approach is used, we tend to manipulate the case where there is a mutual coupling effect (the source of the antenna impairment) among the antenna elements. However, this effect cannot be solely eliminated. Thus, we tend to optimize this effect rather than compensating or eliminating it as have been done before. By reformulating the problem into a least square problem with a practical linear constraint, we can get a standard convex optimization problem, which can used to get the optimal values of the load impedances that can lead to the optimum results. This of course has been done with the aid of the QPSK modulation. Simulation results show that the proposed AD precoder, a better BER performance can be achieved compared to the ideal or conventional cases. The future research focus will be on investigating the performance for the robust precoder with imperfect CSI.

ACKNOWLEDGMENT

This work was funded by Ministry of Higher Education and Scientific Research (MOHESR)in Iraq. The authors, therefor, acknowledge with thanks (MOHESR) for technical and financial support.

REFERENCES

[1] C. Windpassinger, R. Fischer, T. Vencel, and J. Huber, "Precoding in Multiantenna and Multiuser Communications," IEEE Trans. Wireless. Commun., vol. 3, no. 4, pp. 1305–1316, July 2004.

[2] M. Costa, "Writing on Dirty Paper," IEEE Trans. Inf. Theory, vol. IT-29, no. 3, pp. 439–441, May 1983.

[3] A. Garcia-Rodriguez and C. Masouros, “Power-Efficient Tomlinson-Harashima Precoding for the Downlink of Multi-User MISO Systems,” IEEE Trans. Commun., vol. 62, no. 6, pp. 1884–1896, April 2014.

[4] B. M. Hochwald, C. B. Peel, and A. L. Swindlehurst, “A Vector-Perturbation Technique for Near-Capacity Multiantenna Multiuser Communication-Part ii: Perturbation,” IEEE Trans. Commun., vol.53, no.3, pp.537–544, Mar.2005.

[5] J. Maurer, J. Jalden, D. Seethaler, and G. Matz, “Vector Perturbation Precoding Revisited,” IEEE Trans. Sig. Process., vol. 59, no. 1, pp. 315–328, Jan. 2011.

[6] T. Haustein, C. von Helmolt, E. Jorswieck, V. Jungnickel, and V.Pohl, “Performance of MIMO Systems with Channel Inversion,” Proc. 55th IEEE veh. Technol. Conf. (VTC), vol. 1, pp. 35–39, May 2002.

[7] C. Wang, E. K. S. Au, R. D. Murch, W. H. Mow, R. S. Cheng, and V. Lau, “On the performance of the MIMO zero-forcing receiver in the presence of channel estimation error,” IEEE Trans. Wireless Commun., vol. 6, no. 3, pp. 805–810, Mar. 2007.

[8] C. B. Peel, B. M. Hochwald, and A. L. Swindlehurst, “A Vector-Perturbation Technique for Near-Capacity Multiantenna Multiuser Communication-Part i: Channel Inversion and Regularization,” IEEE Trans. Commun., vol.53, no.1, pp. 195–202, Jan. 2005.

[9] C. Masouros, “Correlation Rotation Linear Precoding for MIMO Broadcast Communications,” IEEE Trans. Sig. Process., vol. 59, no. 1, pp. 252–262, Jan. 2011.

[10] S. M. Razavi, T. Ratnarajah, and C. Masouros, “Transmit-power efficient linear precoding utilizing known interference for the multiantenna downlink,” IEEE Trans. Veh. Technol., vol. 63, no. 9, pp. 4383–4394, Nov. 2014.

[11] C. Masouros, M. Sellathurai, and T. Ratnarajah, “Large-Scale MIMO Transmitters in Fixed Physical Spaces: The Effect of Transmit Correlation and Mutual Coupling,” IEEE Trans. Commun., vol. 61, no. 7, pp. 2794–2804, July 2013.

[12] E. Larsson, O. Edfors, F. Tufvesson, and T. Marzetta, “Massive MIMO for Next Generation Wireless Systems,” IEEE Commun. Mag., vol. 52, no. 2, pp. 186–195, Feb. 2014.

[13] L. Lu, G. Y. Li, A. L. Swindlehurst, A. Ashikhmin, and R. Zhang, “An Overview of Massive MIMO: Benefits and Challenges,” IEEE J. Sel. Topics Sig. Process., vol. 8, no. 5, pp. 742–758, April 2014.

[14] J. Wang, M. Bengtsson, B. Ottersten, and D. P. Palomar, “Robust MIMO Precoding for Several Classes of Channel Uncertainty,” IEEE Tran. on Sig. Process., vol. 61, no. 12, pp. 3056- 3070, Jun. 2013.

[15] J. Wang, and D. P. Palomar, “Worst- Case Robust MIMO Transmission with Imperfect Channel Knowledge” IEEE Tran. on Sig. Process., vol. 57, no. 8, August 2009.

[16] M. H. Al- Ali, and K. C. HO, “Transmit Precoding in Underlay MIMO Cognitive Radio with Unavailable or Imperfect Knowledge of Primary Interference Channel” IEEE Trans. on Wireless Commu., vol. 15, no. 8, August 2016.

[17] M. H. Al- Ali, and D. K. C. HO, “Precoding for MIMO Channels in Cognitive Radio Networks with CSI Uncertainty and for MIMO Compound Capacity” IEEE Tran. on Sig. Process., vol. 65, no. 15, August 2017.

[18] S. Biswas, C. Masouros, and T. Ratnarajah, “Performance analysis of large multiuser MIMO systems with space-constrained 2-D antenna arrays,” IEEE Trans. Wireless Commun., vol. 15, no. 5, pp. 3492–3505, May 2016.

[19] C. Masouros, M. Sellathurai, and T. Ratnarajah, “Large-scale MIMO transmitters in fixed physical spaces: The effect of transmit correlation and mutual coupling,” IEEE Trans. Commun., vol. 61, no. 7, pp. 2794–2804, Jul. 2013.

[20] M. T. Ivrlac, W. Utschick, and J. A. Nossek, “Fading correlations in wireless MIMO communication systems,” IEEE J. Sel. Areas Commun., vol. 21, no. 5, pp. 819–828, Jun. 2003.

[21] C. Wang and R. D. Murch, “Adaptive Downlink Multi-user MIMO Wireless Systems for Correlated Channels with Imperfect CSI,” IEEE Wireless Commun., vol. 5, no. 9, pp. 2435– 2446, Sept. 2006.

[22] D. Piazza, N. J. Kirsch, A. Forenza, R. W. Heath, and K. R. Dandekar, “Design and evaluation of a reconfigurable antenna array for MIMO systems,” IEEE Trans. Antennas Propag., vol. 56, no. 3, pp. 869–881, Mar. 2008.

[23] P. L. Kafle, A. Intarapanich, A. B. Sesay, J. Mcrory, and R. J. Davies, “Spatial correlation and capacity measurements for wideband MIMO channels in indoor office environment,” IEEE Trans. Wireless Commun., vol. 7, no. 5, pp. 1560–1571, May 2008.

[24] A. M. Tulino, A. Lozano, and S. Verdu, “Impact of antenna correlation on the capacity of multiantenna channels,” IEEE Trans. Inf. Theory, vol. 51, no. 7, pp. 2491–2509, Jul. 2005.

[25] G. Alfano, A. M. Tulino, A. Lozano, and S. Verdu, “Capacity of MIMO channels with one-sided correlation,” in Proc. IEEE Conf. Spread Spectr. Techn. Appl., Aug./Sep. 2004, pp. 515–519.

[26] M. Kiessling and J. Speidel, “Analytical performance of MIMO zero forcing receivers in correlated Rayleigh fading environments,” in Proc. 4th IEEE Workshop Signal Process. Adv. Wireless Commun. (SPAWC), Jun. 2003, pp. 383–387.

[27] H. Liu, Y. Song, and R. C. Qiu, “The impact of fading correlation on the error performance of MIMO systems over Rayleigh fading channels,” IEEE Trans. Wireless Commun., vol. 4, no. 5, pp. 2014–2019, Sep. 2005.

[28] C. K. Wen, Y. N. Lee, J.T. Chen, and P. Ting, “Asymptotic Spectral efficiency of MIMO multiple-access wireless systems exploring only channel spatial correlations,” IEEE Trans. Signal Process., vol. 53, no. 6, pp. 2059–2073, Jun. 2005.

[29] A. Alexiou and M. Qaddi, “Robust linear precoding to compensate for antenna correlation in orthogonal space-time block coded systems,” in Proc. Sensor Array Multichannel Signal Process. Workshop, Jul. 2004, pp. 701–705.

[30] H. R. Bahrami and T. Le-Ngoc, “Precoder design based on the channel correlation matrices,” IEEE Trans. Wireless Commun., vol. 5, no. 12, pp. 3579–3587, Dec. 2006.

[31] J. Akhtar and D. Gesbert, “Spatial multiplexing over correlated MIMO channels with a closed-form precoder,” IEEE Trans. Wireless Commun., vol. 4, no. 5, pp. 2400–2409, Sep. 2005.

[32] B. Clerckx, C. Craeye, D. Vanhoenacker-Janvier, and C. Oestges, “Impact of antenna coupling on 2×2 MIMO communications,” IEEE Trans. Veh. Technol., vol. 56, no. 3, pp. 1009–1018, May 2007.

[33] S. Lu, H. T. Hui, M. E. Bialkowski, X. Liu, H. S. Hui, and N. V. Shuley, “Effects of antenna mutual coupling on the performance of MIMO systems,” in Proc. 29th Symp. Inf. Theory, Benelux, Belgium, May 2008, pp. 2945–2948.

[34] C. A. Balanis, Antenna Theory: Analysis and Design, 4th ed. Hoboken, NJ, USA: Wiley, Mar. 2016.

[35] I. J. Gupta and A. A. Ksienski, “Effect of mutual coupling on the performance of adaptive arrays,” IEEE Trans. Antennas Propag., vol. 31, no. 5, pp. 785–791, Sep. 1983.

[36] A. A. Abouda and S. G. Haggman, “Effect of mutual coupling on capacity of MIMO wireless channels in high SNR,” Prog. Electromagn. Res., vol. 65, pp. 27–40, Jun. 2005.

[37] H. Steyskal and J. S. Herd, “Mutual Coupling Compensation in Small Array Antennas,” IEEE Trans. Ant. Propag., vol. 38, no. 2, pp. 1971–1975, Dec. 1990.

[38] H. Steyskal and J. S. Herd, “Mutual coupling compensation in small array antennas,” IEEE Trans. Antennas Propag., vol. AP-38, no. 12, pp. 1971–1975, Dec. 1990.

[39] J. Corcoles, M. A. Gonzalez, and J. Rubio, “Mutual coupling compensation in arrays using a spherical wave expansion of the radiated field,” IEEE Antennas Wireless Propag. Lett., vol. 8, pp. 108–111, 2009.

[40] J. Rubio, J. F. Izquierdo, and J. Córcoles, “Mutual coupling compensation matrices for transmitting and receiving arrays,” IEEE Trans. Antennas

Propag., vol. 63, no. 2, pp. 839–843, Feb. 2015.

[41] G. Moreno, H. M. Bernety, and A. B. Yakovlev, “Reduction of mutual coupling between strip dipole antennas at terahertz frequencies with an elliptically shaped graphene monolayer,” IEEE Ant. Wireless Propag. Lett., vol. 15, pp. 1533–1536, Dec. 2015.

[42] S. Farsi, H. Aliakbarian, D. Schreurs, B. Nauwelaers, and G. A. E. Vandenbosch, “Mutual coupling reduction between planar antennas by using a simple microstrip U-section,” IEEE Antennas Wireless Propag. Letter, vol. 11, pp. 1501–1503, 2012.

[43] Z. Li, Z. Du, M. Takahashi, K. Saito, and K. Ito, “Reducing mutual coupling of MIMO antennas with parasitic elements for mobile terminals,” IEEE Trans. Ant. Propag., vol. 60, no. 2, pp. 473–481, Feb. 2012.

[44] H. T. Hui, “A practical approach to compensate for the mutual coupling effect in an adaptive dipole array,” IEEE Trans. Antennas Propag., vol. 52, no. 5, pp. 1262–1269, May 2004.

[45] I. Salonen, A. Toropainen, and P. Vainikainen, “Linear pattern correction in a small microstrip antenna array,” IEEE Trans. Antennas Propag., vol. 52, no. 2, pp. 578–586, Feb. 2004.

[46] R. S. Adve and T. K. Sarkar, “Compensation for the effects of mutual coupling on direct data domain adaptive algorithm,” IEEE Trans. Antennas Propag., vol. 48, no. 1, pp. 86–94, Jan.

[47] S. Boyd and L. Vandenberghe, Convex Optimization. Cambridge, U.K.: Cambridge Univ. Press, 2004.

[48] A. Li, and C. Masouros, “Exploiting Constructive Mutual Coupling in P2P MIMO by Analog- Digital Phase Alignment” IEEE Trans. on Wireless Commun., vol. 16, no. 3, March 2017.

[49] A. Li, and C. Masouros, “Exploiting Mutual Coupling by Means of Analog-Digital Zero Forcing,” IEEE Conf. …….

[50] J. Choma and W. K. Chen, Feedback Networks: Theory and Circuit Applications, U.S.A.: World Scientific Publishing, May 2007.

[51] H. A. Haus, Electromagnetic Noise and Quantum Optical Measurements, Springer, Nov. 2000.

[52] J.-S. Fu, “Adaptive Impedance Matching Circuits based on Ferroelectric and Semiconductor Varactors,” Ph.D. dissertation, Electrical Engineering, University of Michigan, 2009.

[53] A. Li, C. Masouros, and C. B. Papadias, “MIMO Transmission for Single-fed ESPAR with Quantized Loads,” IEEE Trans. Commun., vol. 65, no. 7, pp. 2863–2876, July 2017.

[54] A. Li, C. Masouros, M. Sellathurai, and C. B. Papadias, “Tunable Load MIMO with Quantized Loads,” in 2017 25th European Signal Processing Conference (EUSIPCO), Kos, 2017, pp. 1699–1703.

[55] L. Zhou, F. A. Khan, T. Ratnarajah, and C. B. Papadias, “Achieving Arbitrary Signals Transmission Using a Single Radio Frequency Chain,” IEEE Trans. Commun., vol. 63, no. 12, pp. 4865–4878, Oct. 2015.

[56] G. C. Alexandropoulos, V. I. Barousis, and C. B. Papadias, “Precoding for Multiuser MIMO Systems with Single-Fed Parasitic Antenna Arrays,” in 2014 IEEE Global Communications Conference (GLOBECOM), Dec. 2014, pp. 3897–3902.

[57] O. N. Alrabadi, C. Divarathne, P. Tragas, A. Kalis, N. Marchetti, C. B. Papadias, and R. Prasad, “Spatial Multiplexing with a Single Radio: Proof-of-Concept Experiments in an Indoor Environments with a 2.6GHz Prototype,” IEEE Commun. Lett., vol. 15, no. 2, pp. 178–180, Feb. 2011.