Research Article

Diversity and Multiplexing Technologies by 3D Beams in Polarized Massive MIMO Systems

Xin Su¹ and KyungHi Chang²

¹College of Internet of Things (IOT) Engineering, Hohai University, Changzhou 213022, China

²Electronic Engineering Department, Inha University, Incheon 402751, Republic of Korea

Correspondence should be addressed to KyungHi Chang; khchang@inha.ac.kr

Received 13 July 2015; Revised 14 November 2015; Accepted 29 November 2015

Academic Editor: Jing Liang

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Massive multiple input, multiple output (M-MIMO) technologies have been proposed to scale up data rates reaching gigabits per second in the forthcoming 5G mobile communications systems. However, one of crucial constraints is a dimension in space to implement the M-MIMO. To cope with the space constraint and to utilize more flexibility in 3D beamforming (3D-BF), we propose antenna polarization in M-MIMO systems. In this paper, we design a polarized M-MIMO (PM-MIMO) system associated with 3D-BF applications, where the system architectures for diversity and multiplexing technologies achieved by polarized 3D beams are provided. Different from the conventional 3D-BF achieved by planar M-MIMO technology to control the downtilted beam in a vertical domain, the proposed PM-MIMO realizes 3D-BF via the linear combination of polarized beams. In addition, an effective array selection scheme is proposed to optimize the beam-width and to enhance system performance by the exploration of diversity and multiplexing gains; and a blind channel estimation (BCE) approach is also proposed to avoid pilot contamination in PM-MIMO. Based on the Long Term Evolution-Advanced (LTE-A) specification, the simulation results finally confirm the validity of our proposals.

1. Introduction

Massive multiple input, multiple output (M-MIMO) is being developed as a promising technology for several attractive features [1–3]; for example, the system capacity can be theoretically increased by installing sufficient antennae, and the transmit power can potentially be reduced in inverse proportional to the square root of the number of applied antennae [4, 5]. This not only is relevant from a commercial standpoint but also provides green transmission to address health concerns in wireless communications [6]. When channel reciprocity is exploited in time division duplex (TDD) M-MIMO systems, the overhead related to channel compensation scales linearly only with the number of mobile user-equipment (UE) per cell rather than the number of antennae per base station (BS) [7]. Matched filtering is suggested as being optimal for linear precoders and detectors because thermal noise, interference, and channel estimation errors can theoretically vanish in M-MIMO systems [1]. The remaining performance limitation is

the pilot contamination [8], which is the residual interference caused by the reuse of pilot patterns. The above claims are validated based on several crucial but optimistic assumptions of perfect channel estimation, hardware implementations, and the number of antennae applied in practice. Recently, a lot of the literature has studied M-MIMO with more realistic assumptions [9–13].

On the other hand, there has been a gradual demand for the use of polarized antenna systems, especially for 5G mobile communications systems [14–16]. This is mainly because the antenna polarization is a pivotal resource to be exploited for the design of space-limited wireless devices. Techniques such as space-time diversity, multiplexing, and array processing can be applied to polarized antenna systems to boost system throughput.

In this paper, we propose a polarized M-MIMO (PM-MIMO) array system, where three orthogonally colocated antenna branches are applied at each array element (AE) of an M-MIMO system. Three-dimensional beamforming (3D-BF)

can be realized by the proposed PM-MIMO system, and the generated beams can be steered and varied at X-Y, X-Z, and Y-Z planes, respectively. The system architectures of diversity and multiplexing schemes achieved by polarized 3D beams are provided based on the proposed PM-MIMO array system. A robust array selection scheme for 3D-BF applications is additionally proposed to efficiently optimize the beam-width and to enhance system performance by the exploration of diversity and multiplexing gains.

Normally in M-MIMO systems, when the number of BS antennae grows large, the size of M-MIMO channel matrix grows large. The vector of M-MIMO channel matrix becomes very long and any two of them are pairwise orthogonal. However, in a space-limited system, pairwise orthogonality cannot be maintained because the adjacent AE space is usually set equal to or to less than a half signal wavelength. Therefore, for this paper, we additionally modified a conventional blind channel estimation (BCE) approach to exploit the pairwise orthogonality according to the particular characteristics of PM-MIMO systems. That is, the polarized cross-branch links in the system are usually uncorrelated [17]. The proposed BCE approach is presented for PM-MIMO to avoid the pilot contamination and to enhance the system spectrum efficiency. By applying our proposals under the polarized MIMO channel model, Monte Carlo simulations finally confirm the validity of our proposals.

The remaining parts of this paper are organized as follows: Section 2 describes the proposed PM-MIMO array system, and the proposed AE selection scheme is provided in Section 3. In Section 4, a BCE approach for PM-MIMO is introduced, and the simulation results are demonstrated and discussed in Section 5. Finally, our conclusions are drawn in Section 6.

2. PM-MIMO Array System

Figure 1 provides an example of a uniform linear array (ULA) with antenna polarization in the array and branch (A&B) multiple antennae configuration, where three orthogonally colocated antenna branches are fixed at each AE (i.e., antenna port). The beam-width is proved to be relevant to array configuration, where it is inversely proportional to the number of AEs and array element spacing. Because the spacing of three colocated branches at each AE is set to zero, as shown in Figure 1, which makes beam-width scale up to 360°, the beams should be generated via corresponding cross-array branches rather than the colocated branches at each AE [16]. Therefore, we can obtain three orthogonal beams generated by a polarized ULA as follows:

- (i) Beam steered in X-Y plane is generated by the branches set of $A_p B_1$, where p is the index of AE.
- (ii) Beam steered in X-Z plane is generated by the branches set of $A_p B_2$.
- (iii) Beam steered in Y-Z plane is generated by the branches set of $A_p B_3$.

Let $P_{B_1}(\theta, \phi, \theta_{0,1})$, $P_{B_2}(\theta, \phi, \theta_{0,2})$, and $P_{B_3}(\theta, \phi, \theta_{0,3})$ represent the beam radiation patterns generated by the

FIGURE 1: Uniform linear array with antenna polarization.

above three sets $A_p B_1$, $A_p B_2$, and $A_p B_3$, respectively. We have

$$P_{B_1}(\theta, \phi, \theta_{0,1}) = \sum_{n=1}^N e^{j(n-1)(2\pi/\epsilon)(\sin \theta - \sin \theta_{0,1})} \quad (1)$$

$$= \frac{\sin((\pi N/\epsilon)(\sin \theta - \sin \theta_{0,1}))}{N \cdot \sin((\pi/\epsilon)(\sin \theta - \sin \theta_{0,1}))},$$

$$P_{B_2}(\theta, \phi, \theta_{0,2}) = \sum_{n=1}^N e^{j(n-1)(2\pi/\epsilon)(\sin \phi - \sin \theta_{0,2})} \quad (2)$$

$$= \frac{\sin((\pi N/\epsilon)(\sin \phi - \sin \theta_{0,2}))}{N \cdot \sin((\pi/\epsilon)(\sin \phi - \sin \theta_{0,2}))},$$

$$P_{B_3}(\theta, \phi, \theta_{0,3}) = \sum_{n=1}^N e^{j(n-1)(2\pi/\epsilon)(\sin(\pi/2 - \phi) - \sin \theta_{0,3})} \quad (3)$$

$$= \frac{\sin((\pi N/\epsilon)(\cos \phi - \sin \theta_{0,3}))}{N \cdot \sin((\pi/\epsilon)(\cos \phi - \sin \theta_{0,3}))},$$

where θ and ϕ denote the azimuth and elevation angle of the radiation pattern; ϵ is the array element spacing factor defined by [16], and $\theta_{0,1}$, $\theta_{0,2}$, and $\theta_{0,3}$ are the off bore-sight angles corresponding to $A_p B_1$, $A_p B_2$, and $A_p B_3$, respectively [16]. Figure 2(a) with $N = 8$ then depicts the 3D beams generated by a triple polarized ULA (TPULA) system via (1), (2), and (3), where three orthogonal beams can be steered and varied separately on the X-Y, Z-X, and Y-Z planes. For the second subfigure of Figure 2(a), $\theta_{0,1}$, $\theta_{0,2}$, and $\theta_{0,3}$ are set as 30°. According to Figure 2(b), a very narrow beam-width is obtained by applying a large number of AEs, for example, $N = 64$ AEs that performs quite well in terms of BF interferences mitigation.

By employing the TPULA at each multiplexer (MUX), the structure of our proposed PM-MIMO array system is illustrated in Figure 3 with the received signal being given as

$$y = \sum_{q=1}^Q \sum_{b=1}^3 (w_{p,b,q} h_{p,b,q} x_{p,b,q} + n_{p,b,q}), \quad (4)$$

where $h_{p,b,q}$ is the subpolarized channel corresponding to Q th $A_p B_b$ array element, $w_{p,b,q}$ denotes the 3D-BF weights multiplied at each antenna branch, and p (1, 2, ..., P), b (1, 2, 3), and q (1, 2, ..., Q) represent the AE index, branch

(a) 3D beams generated by a triple-polarized ULA system

(b) Beam-width varied with different number of AEs being employed, for example, 8 and 64 AEs

FIGURE 2: 3D beams generated by a polarized ULA.

index, and MUX index, respectively. The 3D-BF, which can be applied for multipath reception or combination of relay signals [18], is introduced as a promising technique in M-MIMO systems to enhance the cellular performance by deploying antenna elements in both horizontal and vertical (H&V) dimensions [19]. However, different from the conventional 3D-BF achieved by planar M-MIMO system to control the downtilted beam in a vertical domain, the proposed PM-MIMO realizes the 3D-BF via the linear combination of polarized beams. It highlights the fact that the 3D-BF not only is addressed by the H&V exploration but also can be achieved by antenna polarization; meanwhile, the antenna polarization also serves as a pivotal solution dedicated to space constraint in M-MIMO systems.

3. Diversity and Multiplexing Achieved by Polarized 3D Beams with an Array Selection Scheme

3.1. System Architecture. Figure 4 describes the architecture of the proposed diversity system via polarized 3D beams, where the multiple users with each equipping three colocated antenna branches are considered. As studied by Dao et al. [17], the cross-array links (e.g., from BS A_1B_1 and A_2B_1 to UE B_1) can be highly correlated by setting the AE space equal to, or less than, a half-wavelength. The cross-branch links (e.g., from BS A_1B_1 and A_1B_2 to UE B_1), on the other hand, are usually uncorrelated due to the space polarization. This inspired us to incorporate space-time block

FIGURE 3: Structure of the proposed PM-MIMO array system.

coding (STBC) and BF techniques simultaneously in the TPULA system to boost performance. Moreover, in order to achieve full-rate coding with an odd number of transmit antennae, quasi-orthogonal STBC (QO-STBC) has emerged in the literature [20, 21], which fully explores diversity gain but increases the complexity of decoding due to nonorthogonal interference. In this paper, the proposed diversity scheme combines QO-STBC for three transmit antennae and BF techniques via the TPULA system as illustrated in Figure 4.

In Figure 4, the BF weights ($w_{p_b}^{\text{Tx}}$) are multiplied before the inverse fast Fourier transform (IFFT) block of the BS. $\mathbf{X}_p$ in (5) is the transmitted QO-STBC symbol matrix [20]:

$$\mathbf{X}_p = \begin{pmatrix} x_{\text{II}} & x_{\text{III}} & x_{\text{IV}} \\ x_{\text{I}}^* & -x_{\text{IV}}^* & x_{\text{III}}^* \\ x_{\text{IV}} & x_{\text{I}} & x_{\text{II}} \\ x_{\text{III}}^* & -x_{\text{II}}^* & x_{\text{I}}^* \end{pmatrix}_{I\text{-by-}J}, \quad (5)$$

where the Roman numeral is the symbol index, i is the time index modulo of 4 ($I = 4$), and j is the antenna index modulo of 3 ($J = 3$). At each UE, the QO-STBC decoding is applied after the FFT block at each antenna branch [20], and then we

can have the received signal for each branch of a specific UE before QO-STBC decoding as

$$\begin{aligned} \mathbf{R}_{p \sim \text{ithUE}_b} &= (\mathbf{h}_{p \sim \text{ithUE}_b} (\mathbf{W}_{p \sim \text{ithUE}_b}^{\text{Tx}} \odot \mathbf{X}_{p \sim \text{ithUE}_b})^T)^T \\ &\quad + \mathbf{n}_{p \sim \text{ithUE}_b} \\ &= (\mathbf{W}_{p \sim \text{ithUE}_b}^{\text{Tx}} \odot \mathbf{X}_{p \sim \text{ithUE}_b}) \mathbf{h}_{p \sim \text{ithUE}_b}^T + \mathbf{n}_{p \sim \text{ithUE}_b} \\ &= (r_{p \sim \text{ithUE}_b, \text{I}} \ r_{p \sim \text{ithUE}_b, \text{II}} \ r_{p \sim \text{ithUE}_b, \text{III}} \ r_{p \sim \text{ithUE}_b, \text{IV}})^T, \end{aligned} \quad (6)$$

where “ $\odot$ ” denotes the Hadamard product and $\mathbf{W}_p^{\text{Tx}}$ is the TxBF weighting matrix given by

$$\mathbf{W}_{p \sim \text{ithUE}_b}^{\text{Tx}} = \begin{pmatrix} w_{p_1}^{\text{Tx}} & w_{p_2}^{\text{Tx}} & w_{p_3}^{\text{Tx}} \\ w_{p_1}^{\text{Tx}} & w_{p_2}^{\text{Tx}} & w_{p_3}^{\text{Tx}} \\ w_{p_1}^{\text{Tx}} & w_{p_2}^{\text{Tx}} & w_{p_3}^{\text{Tx}} \\ w_{p_1}^{\text{Tx}} & w_{p_2}^{\text{Tx}} & w_{p_3}^{\text{Tx}} \end{pmatrix}. \quad (7)$$

$\mathbf{h}_{p \sim \text{ithUE}_b}$ is the vector of the polarized cross-branch links, given by

$$\mathbf{h}_{p \sim \text{ithUE}_b} = (h_{p_1 \sim \text{ithUE}_b} \ h_{p_2 \sim \text{ithUE}_b} \ h_{p_3 \sim \text{ithUE}_b}), \quad (8)$$

FIGURE 4: Diversity system architecture by polarized 3D beams.

and $\mathbf{n}_{p \sim \text{ithUE}_b}$ is the noise vector given by

$$\begin{aligned} \mathbf{n}_{p \sim \text{ithUE}_b} &= (\sigma_{p \sim \text{ithUE}_b, \text{I}} \quad \sigma_{p \sim \text{ithUE}_b, \text{II}} \quad \sigma_{p \sim \text{ithUE}_b, \text{III}} \quad \sigma_{p \sim \text{ithUE}_b, \text{IV}})^T. \end{aligned} \quad (9)$$

The STBC decoding applied thereafter gives

$$\begin{aligned} \tilde{\mathbf{X}}_{p \sim \text{ithUE}_b} &= \mathbf{D}_{p \sim \text{ithUE}_b} \mathbf{T}_{p \sim \text{ithUE}_b} \\ &= (\tilde{x}_{p \sim \text{ithUE}_b, \text{I}} \quad \tilde{x}_{p \sim \text{ithUE}_b, \text{II}} \quad \tilde{x}_{p \sim \text{ithUE}_b, \text{III}} \quad \tilde{x}_{p \sim \text{ithUE}_b, \text{IV}})^T. \end{aligned} \quad (10)$$

Here, $\mathbf{D}_{p \sim \text{ithUE}_b}$ is the STBC decoding matrix, which is modified based on equation (12) in [20] by considering TxBF weights:

$$\mathbf{D}_{p \sim \text{ithUE}_b} = \begin{pmatrix} 0 & w_{p_1}^{\text{Tx}} h_{p_1 \sim \text{ithUE}_b} & (w_{p_2}^{\text{Tx}})^* h_{p_2 \sim \text{ithUE}_b}^* & w_{p_3}^{\text{Tx}} h_{p_3 \sim \text{ithUE}_b} \\ (w_{p_1}^{\text{Tx}})^* h_{p_1 \sim \text{ithUE}_b}^* & 0 & (w_{p_3}^{\text{Tx}})^* h_{p_3 \sim \text{ithUE}_b}^* & -w_{p_2}^{\text{Tx}} h_{p_2 \sim \text{ithUE}_b} \\ (w_{p_2}^{\text{Tx}})^* h_{p_2 \sim \text{ithUE}_b}^* & w_{p_3}^{\text{Tx}} h_{p_3 \sim \text{ithUE}_b} & 0 & w_{p_1}^{\text{Tx}} h_{p_1 \sim \text{ithUE}_b} \\ (w_{p_3}^{\text{Tx}})^* h_{p_3 \sim \text{ithUE}_b}^* & -w_{p_2}^{\text{Tx}} h_{p_2 \sim \text{ithUE}_b} & (w_{p_1}^{\text{Tx}})^* h_{p_1 \sim \text{ithUE}_b}^* & 0 \end{pmatrix}, \quad (11)$$

where $\mathbf{T}_{p \sim q_b}$ is defined as

$$\mathbf{T}_{p \sim \text{ithUE}_b} = \begin{pmatrix} \tilde{\mathbf{r}}_{p \sim \text{ithUE}_b, \text{I}} & \tilde{\mathbf{r}}_{p \sim \text{ithUE}_b, \text{II}}^* & \tilde{\mathbf{r}}_{p \sim \text{ithUE}_b, \text{III}} & \tilde{\mathbf{r}}_{p \sim \text{ithUE}_b, \text{IV}}^* \end{pmatrix}^T \quad (12)$$

by taking the conjugate of the second and the fourth elements of $\mathbf{R}_{p \sim \text{ithUE}_b}$. The final output is obtained by the functional block of equal gain combining (EGC) as

$$\bar{\mathbf{X}} = \sum_{b=1}^3 \begin{pmatrix} \bar{\mathbf{x}}_{p \sim \text{ithUE}_b, \text{I}} & \bar{\mathbf{x}}_{p \sim \text{ithUE}_b, \text{II}} & \bar{\mathbf{x}}_{p \sim \text{ithUE}_b, \text{III}} & \bar{\mathbf{x}}_{p \sim \text{ithUE}_b, \text{IV}} \end{pmatrix}^T. \quad (13)$$

Figure 5 describes the architecture of multiplexing system by polarized 3D beams, where each generated beam is dedicated to a piece of UE. In this case, the data rate is three times higher than the 3D-BF diversity system. Here, we also consider multiple users, with each of them equipping three colocated antenna branches, and note that, in these two proposed schemes, the zero-forcing BF is assumed and applied at transmitter Tx (i.e., the BS). Compared with diversity case, the proposed multiplexing scheme is relatively simple that the received signal for a specific UE after EGC process can be given as

$$\mathbf{R}_{p \sim \text{ithUE}_b} = \sum_{b=1}^3 \left(\mathbf{h}_{p \sim \text{ithUE}_b} \left(\mathbf{W}_{p \sim \text{ithUE}_b}^{\text{Tx}} \odot \mathbf{S}_{p \sim \text{ithUE}_b} \right) + \mathbf{n}_{p \sim \text{ithUE}_b} \right). \quad (14)$$

3.2. Array Selection Scheme for PM-MIMO Array System. The sharp beam as illustrated in Figure 2(b) performs quite well in terms of BF interferences mitigation. However, it is not efficient to assign all AEs of a MUX to form one beam, especially for cell-edge UEs. Overall, AEs in a MUX can be divided into several groups to form beams for separate UEs, and this part analyses the minimum number of AEs required to mitigate BF interferences.

Let us have two simultaneous beams point two adjacent pieces of UEs, as demonstrated in Figure 6, where L gives the beam coverage and D and d denote the distance from UE to BS and the distance between the adjacent UEs, respectively. To make the analysis meaningful, we assume that there is a sufficient amount of UE deployed in a cell, and therewith, the UEs are close each other, resulting in $d_1 \approx v_1$ and $d_2 \approx v_2$. According to Figure 6, in the case of UE (UE-1 and UE-2) located at a 3 dB beam area, that is, a half power beam-width (HPBW) area, the HPBW needs to be controlled at less than $2d_1$ to avoid BF interferences. On the other hand, when UEs (UE-3 and UE-4) are located at a beam peak area, the HPBW can be set larger than $2d_2$. In addition to assuming that UE is distributed in a square cell, which can be treated as a city block for horizontal BF or a building for vertical BF, we provide Figure 7 for calculating average D and d . Suppose that UEs follow a spatial Poisson process with an intensity of ρ , so the number of UEs in a cell is given as

$$K = \rho L^2. \quad (15)$$

The average distance between the BS and UE is calculated by

$$\begin{aligned} D &= \iint_{L^2} \frac{1}{L^2} r \, dr \, d\theta \\ &= \frac{2}{L^2} \left(\int_0^\omega \int_0^{a/\cos\theta} r^2 \, dr \, d\theta + \int_\omega^{\pi/2} \int_0^{2a/\sin\theta} r^2 \, dr \, d\theta \right) \\ &= \frac{\xi L}{2}, \end{aligned} \quad (16)$$

where $\xi = \ln(2 + \sqrt{5})/12 + \sqrt{5} - 2 \ln((\sqrt{5} - 1)/2)/3 \approx 1.187$. The average distance between two adjacent pieces of UEs is given as

$$d = \iint_A \frac{1}{A} r \, dr \, d\theta = \int_0^{2\pi} \int_0^{r_{\text{UE}}} \frac{r^2}{\pi r_{\text{UE}}^2} \, dr \, d\theta = \frac{2}{3} r_{\text{UE}}, \quad (17)$$

where A denotes the local coverage area of UE with a radius of r_{UE} depending on ρ .

Because D is calculated close to $L/2$, the HPBW $< 2d$ needs to be maintained to avoid BF interferences. According to Su and Chang [16], we have

$$\text{HPBW} \approx 2 \left| \arcsin \left(1.391 \frac{\varepsilon}{\pi P} + \sin \theta_0 \right) \right| < \frac{4}{3} r_{\text{UE}}, \quad (18)$$

where ε denotes the array spacing factor; that is, $\varepsilon = \lambda/s$ with s and λ representing the array spacing and signal wavelength, respectively. θ_0 is the signal incidence angle shifted from the bore-sight direction, and the antenna bore-sight is the axis vertical to the orientation of the array alignment. For example, if the radius of a user's local area (r_{UE}) is 15 meters, at least 8 AEs are required to avoid BF interferences when $\varepsilon = 2$ and $\theta_0 = \pi/4$. Please note that the above analysis is derived based on the isotropic array antenna system. The required number of AEs may decrease by using the dipole antenna because it does not radiate in the longitudinal direction of an antenna structure that maintains a higher radiation gain compared with the isotropic antenna.

As discussed in Su and Chang [16] and Liu [22], the beam-width is increased significantly when the beam steers to an angle far off the bore-sight direction, such as the case of θ_0 in (18) reaching 90° . In order to avoid beam-width extension, we propose a scheme to cope with the large off bore-sight angle by dynamically selecting the set of polarized branches for 3D-BF that can effectively work without increasing the dimension of the array system. As depicted in Figure 8, let α_k , β_k , and γ_k denote the acute angles corresponding to the X , Y , and Z axes of the k th incident signal; we have

$$\begin{aligned} \alpha_k &= \arcsin \frac{\sqrt{b_k^2 + c_k^2}}{D_k}, \\ \beta_k &= \arcsin \frac{\sqrt{a_k^2 + c_k^2}}{D_k}, \\ \gamma_k &= \arcsin \frac{\sqrt{a_k^2 + b_k^2}}{D_k}, \end{aligned} \quad (19)$$

FIGURE 5: Multiplexing system architecture by polarized 3D beams.

FIGURE 6: Two beams steer toward two adjacent pieces of UE.

FIGURE 7: Illustration of a square cell used for calculating D and d .

FIGURE 8: Selection on the set of branches for 3D-BF via incident signal.

where α_k , β_k , and $\gamma_k \in (0, \pi/2)$. In order to avoid beam-width extension by considering the off bore-sight angle, the following criteria need to be carried out for 3D-BF applications:

- (i) When $\alpha_k = \max(\alpha_k, \beta_k, \gamma_k)$, use the set of branches of $A_p B_3$ to form the beam for the k th incident signal.
- (ii) When $\beta_k = \max(\alpha_k, \beta_k, \gamma_k)$, use the set of branches of $A_p B_2$ to form the beam for the k th incident signal.
- (iii) When $\gamma_k = \max(\alpha_k, \beta_k, \gamma_k)$, use the set of branches of $A_p B_1$ to form the beam for the k th incident signal.

Let $y_1(t), y_2(t), \dots, y_k(t)$ denote the incident signal sequences that come from random directions; Figure 9 then provides the flowchart of the proposed AE selection scheme for the PM-MIMO system. According to Figure 9, the proposed scheme at first detects the incident signal sequences and determines the minimum number of AEs used for 3D-BF via the criterion provided by (18). And then, the proposed scheme categorizes the incident signal sequences into three categories according to $\max(\alpha_k, \beta_k, \gamma_k)$. For each category, the proposed schemes prepare the set of branches used for 3D-BF via the criteria provided before. If the 3D-BF application

is dedicated to the cell-edge users, the diversity by using polarized 3D beams is employed to maintain the cell-edge users' performances. Otherwise, multiplexing via polarized 3D beams is suggested in order to increase overall system throughput.

4. Blind Channel Estimation to Avoid Pilot Contamination in PM-MIMO Array System

Theoretically, the M-MIMO system is proved to have many attractive features in wireless communications. However, these features are obtained mainly based on the perfect channel estimation. Practically, the BS does not have perfect channel state information that limits the exploration of M-MIMO systems. Moreover, conventional channel estimation by using training sequences may not be applicable to M-MIMO systems because usually there are tens or hundreds of antennae applied in M-MIMO systems. Spectrum efficiency could not only be decreased dramatically by reserving many pilots for channel estimation, but pilot contamination also limits performance because the pilot positions in a resource block have to be reused due to massive antenna employment [23–26].

FIGURE 9: Flowchart of the proposed AE selection scheme for PM-MIMO system.

The BCE approach, which requires no, or a minimal number of, pilots needs to be applied in M-MIMO systems. One of the BCE strategies is based on eigenvalue decomposition (EVD) for the covariance matrix of a received signal that needs to preserve pairwise orthogonality among channel vectors [27–29]. Let us define an M-MIMO system model as

$$\mathbf{y}_k = \mathbf{H}\mathbf{x}_k + \mathbf{n}_k, \quad (20)$$

where $\mathbf{x}_k$ is the transmitted symbols, $\mathbf{H}$ is an M -by- N channel matrix between the BS and the k th UE, and $\mathbf{n}_k$ denotes

the additive white noise. The covariance matrix of the received signal is then defined by

$$\mathbf{R}_{\mathbf{y}_k} \triangleq \mathbf{y}_k \mathbf{y}_k^H = \mathbf{H} \mathbf{R}_{\mathbf{x}_k} \mathbf{H}^H + \mathbf{R}_{\mathbf{n}_k}, \quad (21)$$

where H represents the Hermitian Transpose operation. Additionally multiplying $\mathbf{H}$ at both sides of (21), we have

$$\mathbf{R}_{\mathbf{y}_k} \mathbf{H} = \mathbf{H} \mathbf{R}_{\mathbf{x}_k} \mathbf{H}^H \mathbf{H} + \mathbf{H} \mathbf{R}_{\mathbf{n}_k}. \quad (22)$$

From the law of large number, the channel vectors between the BS and the deployed UEs become very long, random, and pairwise orthogonal, which satisfies the condition

$$\frac{1}{M} \mathbf{H}^H \mathbf{H} \rightarrow \mathbf{I}_N, \quad \text{as } M \rightarrow \infty. \quad (23)$$

Thereafter, the estimated channel can be obtained as the eigenvectors of $\mathbf{R}_{y_k}$ via EVD processing [29].

Note that since we focus on the link-level performance, the element of channel matrix $\mathbf{H}$ should be normalized, and only the additive noise and fast fading effects are taken into consideration to generate $\mathbf{H}$. If the array spacing is larger than a half wave-length, the elements of $\mathbf{H}$ would be lowly correlated, and then the condition of (23) is easy to be satisfied as reported elsewhere [30], where the BS AEs are usually separated by several wave-lengths resulting in an uncorrelated Tx radiation pattern to preserve the pairwise orthogonality among channel vectors. However, in space-limited M-MIMO systems, the pairwise orthogonality cannot be maintained because the adjacent AE space is fixed, normally, at equal

or less than a half signal wave-length. Consequently, in this paper, we modify the EVD-based blind channel estimation scheme studied in Ngo and Larsson [29]. The modified BCE approach can exploit the pairwise orthogonality via the particular characteristics of PM-MIMO systems; that is, the polarized cross-branch links in the system usually are uncorrelated, even though the adjacent AE spacing is set equal to, or less than, a half signal wave-length [17].

Based on other researches [14, 16, 31], an extension of the channel matrix for PM-MIMO systems can be represented as

$$\mathbf{H}_{\text{PM-MIMO}}(t) = \begin{pmatrix} \mathbf{h}_{\text{Tx}A_1-\text{Rx}A_1}(t) & \cdots & \mathbf{h}_{\text{Tx}A_1-\text{Rx}A_{\hat{P}}}(t) \\ \vdots & \ddots & \vdots \\ \mathbf{h}_{\text{Tx}A_P-\text{Rx}A_1}(t) & \cdots & \mathbf{h}_{\text{Tx}A_P-\text{Rx}A_{\hat{P}}}(t) \end{pmatrix}, \quad (24)$$

where

$$\mathbf{h}_{\text{Tx}A_P-\text{Rx}A_{\hat{P}}}(t) = \sqrt{\frac{\eta}{I}} \sum_{i=1}^I \begin{pmatrix} h_{\text{Tx}A_P B_1-\text{Rx}A_{\hat{P}} B_1}(t) & \sqrt{\mu} \cdot h_{\text{Tx}A_P B_2-\text{Rx}A_{\hat{P}} B_1}(t) & \sqrt{\mu} \cdot h_{\text{Tx}A_P B_3-\text{Rx}A_{\hat{P}} B_1}(t) \\ \sqrt{\mu} \cdot h_{\text{Tx}A_P B_1-\text{Rx}A_{\hat{P}} B_2}(t) & h_{\text{Tx}A_P B_2-\text{Rx}A_{\hat{P}} B_2}(t) & \sqrt{\mu} \cdot h_{\text{Tx}A_P B_3-\text{Rx}A_{\hat{P}} B_2}(t) \\ \sqrt{\mu} \cdot h_{\text{Tx}A_P B_1-\text{Rx}A_{\hat{P}} B_3}(t) & \sqrt{\mu} \cdot h_{\text{Tx}A_P B_2-\text{Rx}A_{\hat{P}} B_3}(t) & h_{\text{Tx}A_P B_3-\text{Rx}A_{\hat{P}} B_3}(t) \end{pmatrix}. \quad (25)$$

Here, η and I denote the transmit power and the number of scatterers, respectively. Channel matrix (24) has a size of $3P$ -by- $3\hat{P}$, which is composed of 3-by-3 submatrices. Each submatrix holds the polarized cross-branch links between the P th AE of Tx (the BS) and the $\hat{P}$ th AE (UE) of receiver (Rx). Please note that throughout the paper the UE is assumed with one AE; that is, $\hat{P} = 1$. The channel vectors of submatrix (25) are random and pairwise orthogonal because the polarized links from a Tx AE with three orthogonal branches to a single Rx branch are usually highly orthogonal [17]. Consequently, we can apply the EVD-based BCE via submatrix of (25). By redefining the covariance matrix of the received signal as

$$\mathbf{R}_{y_k} \triangleq \mathbf{y}_k \mathbf{y}_k^H = \mathbf{h}_{\text{Tx}A_P-\text{Rx}A_{\hat{q}}} \mathbf{R}_{x_k} \mathbf{h}_{\text{Tx}A_P-\text{Rx}A_{\hat{q}}}^H + \mathbf{R}_{n_k}, \quad (26)$$

the estimated channel can be obtained as the eigenvectors of $\mathbf{R}_{y_k}$ via EVD processing because the condition of $(1/3) \mathbf{h}_{\text{Tx}A_P-\text{Rx}A_{\hat{q}}}^H \mathbf{h}_{\text{Tx}A_P-\text{Rx}A_{\hat{q}}} \rightarrow \mathbf{I}_3$ is satisfied.

5. Performance Verification

Table 1 lists the parameters settings of the performance verification based on the LTE-A specification [32]. The downlink data mapping for LTE-A resource blocks used in simulations is demonstrated in Figure 10, which is based on the QO-STBC

symbol matrix of (5). We use the training sequences provided by [32], which are defined as

$$r_{l,n_s}(m) = \frac{1}{\sqrt{2}} (1 - 2 \cdot c(2m)) + j \frac{1}{\sqrt{2}} (1 - 2 \cdot c(2m+1)), \quad (27)$$

where $m = 0, 1, \dots, 2N_{RB}^{\text{max},DL} - 1$. The pilot density is 14.3%, and the initialization of c is defined as

$$c_{\text{init}} = 2^{10} \cdot (7 \cdot (n_s + 1) + l + 1) \cdot (2 \cdot N_{\text{ID}}^{\text{cell}} + 1) + 2 \cdot N_{\text{ID}}^{\text{cell}} + N_{\text{CP}}, \quad (28)$$

where n_s is the slot number within a frame, and

$$N_{\text{CP}} = \begin{cases} 1 & \text{for normal CP} \\ 0 & \text{for extended CP.} \end{cases} \quad (29)$$

Figure 11 depicts the probability densities of the HPBW generated in simulations, where the HPBW can be effectively kept as 50° on average by using the proposed AE selection scheme of Figure 9. However, the average HPBW extends by about 15° when not considering the off bore-sight angle effect. This demonstrates that the proposed AE selection scheme is robust that can optimize the generated beamwidth to avoid BF interference.

FIGURE 10: Downlink data mapping of the LTE-A resource blocks used in the simulation.

FIGURE 11: Probability density of HPBW.

FIGURE 12: Users average BLER performances based on the proposed diversity system architecture.

Figure 12 compares the users' average block error rate (BLER) performances obtained, based on the proposed diversity system architecture by using different transmission schemes, including Case 1 without STBC and TxBF, Case 2 with STBC only, and Case 3 with STBC and TxBF via the LTE pilot-assistant practical channel estimation (PACE) approach. According to Figure 12, Case 1 has the worst performance because there is no space-time coding and BF gain achieved. About 1.7 dB of signal-to-noise ratio (SNR) gain at the target BLER can be achieved by Case 2, compared with Case 1 under the non-line-of-sight NLOS scenario. There is about a 1 dB SNR gain that can be further achieved by employing the TxBF of Case 3. In addition, the simulation results under the LOS scenario are also provided as a comparison, where almost 7 dB gain can be obtained when the LOS exists for those three cases.

Next, we simulate the BCE approaches and compare its efficiency with the PACE based on the proposed diversity system architecture. Figure 13 demonstrates the users' average BLER via PACE and BCE approaches for three different

TABLE 1: Simulation parameters.

Parameters	Value
Carrier frequency	1.8 GHz
System bandwidth	20 MHz
FFT size	2048
Number of data carriers	1200
Number of samples in CP	144
Subcarrier spacing	15 kHz
BS antenna configuration (PM-MIMO)	1 MUX $\times$ 32 AE $\times$ 3 branches
BS AE spacing	Half wavelength
User's antenna configuration	1 AE $\times$ 3 branches
Number of users	4 for 3D-BF diversity 12 for 3D-BF multiplexing
Radius of user's local area (r_{UE})	15 m
Antenna type	Dipole
Modulation	QPSK
Number of scatterers	Outdoor: 4 clusters with 16 scatterers per cluster (64 in total)
Scattering sphere radius	10 m
Velocity of UEs	3, 60, and 120 km/h
Fading	Flat
XPD value	5.8 dB and 9.7 dB for NLOS and LOS
Correlation	0.32 and 0.34 for NLOS and LOS
Rician K -factor	9 dB

user's velocities, including 3, 60, and 120 km/h, under the NLOS scenario. An extension of other researches [14, 31] is applied to simulations of the polarized MIMO channel, for which channel characteristics are also listed in Table 1. We see that the PACE performance decreases a lot due to high mobility at 60 and 120 km/h, which indicates that the number of pilots is not enough to compensate the channel correctly in an environment with fast time-varying phase response. The pilot density is 14.3% for PACE scheme, and by considering the trade-off regarding spectrum efficiency, BCE that requires no, or a minimal number of, pilots might be better employed. Figure 13 additionally shows the performances of BCE schemes where performance is found that is not relevant to the user's velocity. Compared with the BCE reported by Ngo and Larsson [29], our proposed BCE performance is better because the EVD is based on low correlated submatrix and has less complexity for doing the EVD based on a submatrix with a smaller size. Due to no pilot contamination by BCE approach, the proposed BCE outperforms PACE for a velocity of 3 km/h when E_b/N_0 is less than 25 dB. However, BCE has a higher complexity than PACE, and it is condition constraint. Again, according to Figure 13, BCE performs worse than PACE with a velocity of 3 km/h when E_b/N_0 is larger than 25 dB, and the trend of error floor for

FIGURE 13: Users' average BLER performance by PACE and BCE under an NLOS environment based on the proposed diversity system architecture.

FIGURE 14: Users' average BLER performance by PACE and BCE under a LOS environment based on the proposed diversity system architecture.

the proposed BCE occurs at 30 dB. Figure 14 is given by considering the signal LOS component, where performance characteristics observed are the same as Figure 13.

At last, we simulate the proposed multiplexing system architecture via 3D beams, where the channel matrix dedicated to a user is then given as

$$\mathbf{h}_{TA_p,k}(t) = \sqrt{\frac{\eta}{S}} \sum_{s=1}^S \begin{pmatrix} h_{TA_1B_b,k_{B_1}}(t) & \sqrt{\mu} \cdot h_{TA_1B_b,k_{B_2}}(t) & \sqrt{\mu} \cdot h_{TA_1B_b,k_{B_3}}(t) \\ h_{TA_2B_b,k_{B_1}}(t) & \sqrt{\mu} \cdot h_{TA_2B_b,k_{B_2}}(t) & \sqrt{\mu} \cdot h_{TA_2B_b,k_{B_3}}(t) \\ \vdots & \vdots & \vdots \\ h_{TA_pB_b,k_{B_1}}(t) & \sqrt{\mu} \cdot h_{TA_pB_b,k_{B_2}}(t) & \sqrt{\mu} \cdot h_{TA_pB_b,k_{B_3}}(t) \end{pmatrix}. \quad (30)$$

According to channel matrix (30), the performance of the EVD-based BCE approach may decrease because the constraint conditions of BCE are not preserved. However, the pilot contamination effect is reduced because there are fewer antennae using the same pilot positions. For example, there are 24 Tx antennae used in the proposed diversity scheme dedicated to a user, where four antennae share the same pilot positions. In the proposed multiplexing scheme, eight Tx antennae are dedicated to a user, where two of them need to share the same pilot position with other antennae. Figures 15 and 16 illustrate the user's average BLER comparing the multiplexing and diversity system architectures by 3D beams. Note that the data rate of multiplexing is three times higher than the diversity scheme. For PACE performances, there is about 3 dB SNR gain at target BLER of 10^{-2} (at 3 km/h case) achieved by the diversity scheme, compared with the multiplexing scheme, mainly due to the gain of QO-STBC. However, the gain is not significant, because the pilot contamination effect is reduced when employing the multiplexing scheme with fewer antennae using the same pilot positions. The BCE performance, as discussed earlier, cannot be maintained with multiplexing since the constraint condition of BCE is not fulfilled.

6. Conclusions

M-MIMO has been developed as a promising technology due to several attractive features. However, there is less research on M-MIMO systems with antenna polarization, where the antenna polarization can cope with one of crucial constraints: a dimension in space to implement the M-MIMO. In this paper, we propose a PM-MIMO array system with three orthogonally colocated antenna branches equipped at each AE of an M-MIMO system. System architectures of diversity and multiplexing schemes realized by polarized 3D beams are then proposed based on the proposed PM-MIMO array system. An array selection scheme for 3D-BF applications is additionally provided in this paper to efficiently optimize the beam-width and to enhance system performance by the exploration of diversity and multiplexing gains. In order to avoid pilot contamination in PM-MIMO, we also propose a BCE approach to exploit pairwise orthogonality according to the particular characteristics of PM-MIMO systems. With the proposed BCE approach, 14.3% of spectral efficiency can be increased, while the gain in BLER performance is dependent on mobility, compared with PACE. Finally, the simulation results, including the performances comparison

FIGURE 15: Users' average BLER performance by the proposed diversity and multiplexing schemes via 3D beams under an NLOS environment.

FIGURE 16: Users average BLER performance by the proposed diversity and multiplexing schemes via 3D beams under a LOS environment.

between PACE and BCE, as well as diversity and multiplexing schemes, confirmed the validity of our proposals.

Conflict of Interests

The authors declare that there is no conflict of interests regarding the publication of this paper.

Acknowledgments

This research was supported by the MSIP (Ministry of Science, ICT and Future Planning), Korea, under the ITRC (Information Technology Research Center) Support Program (IITP-2015-H8501-15-1019) supervised by the IITP (Institute

for Information & Communications Technology Promotion) and also was supported by "the Fundamental Research Funds for the Central Universities (2015B30614)."

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