HYBRID MIMO-OFDM FOR DOWNLINK MULTI-USER COMMUNICATIONS OVER MILLIMETER CHANNELS WITH NO INSTANTANEOUS FEEDBACK

Po-Hsuan Lee and Yuan-Pei Lin

Department of Electrical Engineering,
National Chiao Tung University, Hsinchu, Taiwan 300, ROC.

ABSTRACT

In this paper, we consider hybrid precoding for multiuser MIMO-OFDM systems in downlink communications with no instantaneous channel information. In a hybrid MIMO-OFDM, precoding consists of two parts: a RF (Radio Frequency) precoder that is subcarrier independent and subcarrier precoders that are subcarrier dependent. We design the precoders of both parts based on the statistics of the user channels, which requires infrequent feedback. To address interference among the users, the signals of different users are transmitted using spatially orthogonal beams. Due to the sparse nature of mmWave channels, the use of orthogonal beams leads to spatial separation among the users and effective prevention of interference. Simulations are given to show that the performance of the proposed statistical hybrid precoding system is close to that of a hybrid system that enjoys full-CSI (Channel State Information).

1. INTRODUCTION

The performance of a MIMO system is known to improve with the number of antennas at the transmitter and receiver. Recently mmWave systems that employ a large number of antennas have been of great interest [1]. However cost and power constraints often prohibit having one dedicated RF chain for each antenna [1, 2]. A promising technique to overcome the RF limitation is the so called hybrid scheme [3], in which analog processing of RF signals is combined with digital processing in the baseband to improve the performance [3]-[5]. In a hybrid MIMO-OFDM, a common RF precoder is shared by all subcarrier beamformers and thus the subcarrier beamformers can not be designed independently. Hybrid beamforming for MIMO-OFDM systems is considered in [4] and a joint design of RF precoders and subcarrier baseband precoders is proposed in [5] for MIMO-OFDM systems. The design of hybrid precoders for multiuser MIMO-OFDM is formulated in [6] and solved using an alternating minimization algorithm. Angular information is exploited in [7] and an iterative method using manifold optimization is given therein.

The use of a large number of antennas escalates the difficulty of feeding back channel information in applications where the transmitter does not have the full CSI, e.g., frequency division duplex systems. A statistical design requires only infrequent update of channel statistics but not instantaneous feedback [8]-[10]. Recently channel covariance is exploited to design beamformers for multiuser narrowband transmission [11]-[13] and statistical construction of RF precoder is proposed in [15] for MIMO-OFDM multiuser systems. However in these systems [11]-[15], the transmitter are zero-forcing and the basesband precoders in these systems requires the information of the current user channels. Solutions of statistical beamforming on the Grassmann manifold for two-user broadcast channel are given in [16]. Statistical beamformers for single-user MIMO-OFDM are derived in [17] based on mmWave channel statistics.

In this paper we consider downlink multiuser transmission. We design hybrid precoding for multiuser MIMO-OFDM systems when there is no instantaneous feedback. The transmitter is designed based on statistics, which can be acquired in the initialization phase, but requiring no frequent feedback of the current channel. By exploiting the statistics of user channels, the RF precoder and subcarrier precoders are so designed that different user signals are transmitted through spatially orthogonal beams. Thanks to the sparse nature of the mmWave channels, the users are spatially separated and thus interference reduced when orthogonality is incorporated in the design. We assume each user knows its own channel as well as statistics of all user channels, which are infrequently updated like the transmitter. The knowledge of the statistics allows interference to be removed completely at the users side via zero-forcing receivers. Simulations are given to show that the proposed statistical precoding yields results close to that of a full-CSI system.

Notation. The variance of a random variable x is denoted as σ_x^2 and the expectation of x by $E[x]$. The 2-norm of a vector $\mathbf{f}$ is denoted as $||\mathbf{f}||$. The notation $\mathbf{A}^\dagger$ denotes the transpose and conjugate of a matrix $\mathbf{A}$.

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2. SYSTEM MODEL

Consider a downlink MIMO system with U users. The transmitter is equipped with N_t antennas and each user with N_r antennas. We adopt the clustered geometric channel representation that is useful for modeling mm-Wave propagation [18] [19]. Suppose the sampling period is T_s and the channel has N_{cl} clusters with N_{ray} paths in each cluster. The uth user's channel is of the form [20]

$$\begin{aligned}\mathbf{H}_u^t(n) = \sqrt{L}\textstyle\sum_{\ell=1}^{N_{cl}}\sum_{i=1}^{N_{ray}} \alpha_{u,\ell,i}p(nT_s-\tau_{u,\ell})\\ \mathbf{a}_r(\phi^r_{u,\ell,i},\theta^r_{u,\ell,i})\mathbf{a}_t(\phi^t_{u,\ell,i},\theta^t_{u,\ell,i})^\dagger,\end{aligned} \tag{1}$$

where the scalar $\alpha_{u,\ell,i}$ is the complex gain of the ith path in ℓth cluster of the uth user and L is $\frac{N_tN_r}{N_{cl}N_{ray}}$. $\theta^t_{u,\ell,i}(\phi^t_{u,\ell,i})$ and $\theta^r_{u,\ell,i}(\phi^r_{u,\ell,i})$ represent, respectively, the angles of departure (AoD) and the angles of arrival (AoA) in azimuth (elevation) angles. The AoD $\theta^t_{u,\ell,i}(\phi^t_{u,\ell,i})$ in the same cluster are of the same mean and variance $\sigma^2_{\theta^t_{u,\ell}}(\sigma^2_{\phi^t_{u,\ell}})$. The vectors $\mathbf{a}_t(\phi^t_{u,\ell,i},\theta^t_{u,\ell,i})$ and $\mathbf{a}_r(\phi^r_{u,\ell,i},\theta^r_{u,\ell,i})$ are, respectively, the transmit and receive antenna array response vectors. The array response vector for an UPA (Uniformed Planar Arrays) arranged on the yz-plane with size $N_z\times N_y$ (N_z in the z-direction and N_y in the y-direction) is given by

$$[\mathbf{a}(\phi,\theta)]_{m+nN_z} = \frac{1}{\sqrt{N_yN_z}}e^{-j\zeta(m\cos(\theta)+n\sin(\theta)\cos(\phi))},$$

for $0\le m<N_z$ and $0\le n<N_y$, where $\zeta=2\pi d$ and d is the antenna spacing normalized by the wavelength. The rays in the ℓth cluster for the uth user are assumed to have the same delay $\tau_{u,\ell}$. The function $p(t)$ represents the lumped pulse shaping functions. The inter-symbol interference due to multipath can be removed through the insertion of proper cyclic prefix and applying DFT and IDFT in MIMO-OFDM. The channel on the kth subcarrier for the uth user is given by

$$\mathbf{H}_u[k] = \sum_{\ell=1}^{N_{cl}}\sum_{i=1}^{N_{ray}}\beta_{u,\ell,i,k}\mathbf{a}_r(\phi^r_{u,\ell,i},\theta^r_{u,\ell,i})\mathbf{a}_t(\phi^t_{u,\ell,i},\theta^t_{u,\ell,i})^\dagger, \tag{2}$$

where $\beta_{u,\ell,i,k}=\sqrt{L}\alpha_{u,\ell,i}p_{u,\ell,k}$ and $p_{u,\ell,k}=\sum_{n=0}^{N_{cp}}p(nT_s-\tau_{u,\ell})e^{-j2\pi kn/M}$, $k=0,\dots,M-1$. Let the transmitter transmits one data stream to each user and the transmitted matrix on the kth subcarrier be $\mathbf{s}[k]=[s_1[k],s_2[k],\dots,s_U[k]]^T$, $s_u[k]$ for the uth user, and $E\{\mathbf{s}[k]\mathbf{s}^\dagger[k]\}=\frac{P_t}{U}\mathbf{I}_U$, where P_t is the transmit power.

The number of RF chains at the transmitter N^t_{rf} is usually much smaller than the number of antennas N_t. In a hybrid structure, the kth subcarrier beamformer of the uth user is of the form,

$$\mathbf{f}_u[k] = \mathbf{F}_{rf}\mathbf{f}_{bb,u}[k], \tag{3}$$

where $\mathbf{F}_{rf}$ is an $N_t\times N^t_{rf}$ matrix that represents the analog processing, and $\mathbf{f}_{bb,u}[k]$ is the $N^t_{rf}\times 1$ baseband beamformer of the kth subcarrier for the uth user. At the receiver, let the $N_r\times 1$ combiner used for the kth subcarrier be $\mathbf{g}_u[k]$. Then the output of the kth subcarrier for the uth receiver is

$$\mathbf{r}_u[k] = \mathbf{g}_u^\dagger[k]\mathbf{H}_u[k]\mathbf{F}[k]\mathbf{s}[k] + \mathbf{g}_u^\dagger[k]\mathbf{n}_u[k], \tag{4}$$

where $\mathbf{F}[k]=[\mathbf{F}_1[k],\dots,\mathbf{F}_U[k]]$ and $\mathbf{N}[k]$ is the $N_r\times 1$ additive Gaussian noise matrix with zero mean and variance N_0. When a hybrid structure is used, $\mathbf{g}_u[k]$ is of the form $\mathbf{g}_u[k]=\mathbf{G}_{rf,u}\mathbf{g}_{bb,u}[k]$, where $\mathbf{G}_{rf,u}$ is the $N_r\times N^r_{rf}$ analog combiner and $\mathbf{g}_{bb,u}[k]$ is the $N^r_{rf}\times 1$ baseband combiner.

3. STATISTICAL PRECODING

Assume that the transmitter knows the statistics of the channel but has no instantaneous feedback of the current channel information. When we consider only the uth user, it is known that the SNR maximizing combiner is $\mathbf{g}_u[k]=\mathbf{H}_u[k]\mathbf{f}_u[k]$ and the resulting average SNR is given by $\frac{P_t}{UN_0}\mathbf{f}_u^\dagger[k]\mathbf{B}_u[k]\mathbf{f}_u[k]$, where $\mathbf{B}_u[k]=E[\mathbf{H}_u^\dagger[k]\mathbf{H}_u[k]]$ [17]. The optimal statistical $\mathbf{f}_u[k]$ is the dominant eigenvector of $\mathbf{B}_u[k]$. If we repeat this process for each user separately, the resulting beamformers may overlap in spatial frequency, which gives rise to interference among users. To alleviate inter-user interference, we propose to use orthogonal beams, i.e. orthogonal $\mathbf{f}_u[k]$. When there is little inter-user interference, the SNR is approximately $\frac{P_t}{UN_0}\mathbf{f}_u^\dagger[k]\mathbf{B}_u[k]\mathbf{f}_u[k]$. Thus, we maximize the sum of SNR as follows:

$$\max\sum_{u=1}^{U}\sum_{k=1}^{M}\mathbf{f}_u^\dagger[k]\mathbf{B}_u[k]\mathbf{f}_u[k],\ \text{s.t. } \mathbf{f}_i^\dagger[k]\mathbf{f}_j[k]=\delta(i-j), \tag{5}$$

where $\delta(n)$ is the discrete dirac delta function, $\delta(0)=1$ and $\delta(n)=0$ for $n\neq 0$. The task now is to design orthogonal $\mathbf{f}_i[k]$ for maximum sum SNR. Let us first consider the case $N^t_{rf}=U$ and each user is allocated one RF chain. The case $N^t_{rf}>U$ will be discussed at the end of this section. It is shown in [17] that $\mathbf{B}_u[k]$ can be expressed as

$$\mathbf{B}_u[k] = \sum_{\ell=1}^{N_{cl}}\sum_{i=1}^{N_{ray}}|\beta_{u,\ell,i,k}|^2E[\mathbf{a}_t(\phi^t_{u,\ell,i},\theta^t_{u,\ell,i})\mathbf{a}_t(\phi^t_{u,\ell,i},\theta^t_{u,\ell,i})^\dagger].$$

When the pulse shaping filter is low-pass with good roll-off, $|p_{u,\ell,k}|$ is approximately constant except for the subcarriers close to the Nyquist frequency and $|\beta_{u,\ell,i,k}|^2$ is independent of k for most subcarriers. Let

$$\mathbf{B}_u = \sum_{\ell=1}^{N_{cl}}\sum_{i=1}^{N_{ray}}|\alpha_{u,\ell,i}|^2E[\mathbf{a}_t(\phi^t_{u,\ell,i},\theta^t_{u,\ell,i})\mathbf{a}_t(\phi^t_{u,\ell,i},\theta^t_{u,\ell,i})^\dagger].$$

Observe from (3) that each subcarrier beamformer is a linear combination of the column vectors of $\mathbf{F}_{rf}$. That is, the columns of $\mathbf{F}_{rf}$ set the main beamforming direction, so we choose the column vectors of $\mathbf{F}_{rf}$ to be orthogonal as well.

When there is no baseband processing, $\mathbf{f}_u[k] = \mathbf{f}_{rf,u}$. The problem in (5) becomes the following.

$$\max \quad \sum_{u=1}^{U} \mathbf{f}_{rf,u}^{\dagger} \mathbf{B}_u \mathbf{f}_{rf,u}, \text{ s.t. } \mathbf{F}_{rf}^{\dagger} \mathbf{F}_{rf} = \mathbf{I}_{N_{rf}^t}. \quad (6)$$

The constraint in (6) is nonlinear. To solve (6), we use a greedy approach to design the RF beamformers one by one. For the first column vector $\mathbf{f}_{rf,1}$, the objective is simply the maximization of $\mathbf{f}_{rf,1}^{\dagger}\mathbf{B}_1\mathbf{f}_{rf,1}$, and the optimal solution for $\mathbf{f}_{rf,1}$ is the dominant eigenvector of $\mathbf{B}_1$. The rest of the RF column vectors are designed sequentially with additional orthogonality constraint imposed. In particular, for the uth column vector, $u = 2, \ldots, U$, the problem is

$$\max \quad \mathbf{f}_{rf,u}^{\dagger} \mathbf{B}_u \mathbf{f}_{rf,u}, \text{ s.t. } \mathbf{f}_{rf,u}^{\dagger} \mathbf{f}_{rf,j} = \delta(u-j), \quad (7)$$

where $j = 1, \ldots, u-1$. This implies that $\mathbf{f}_{rf,u}$ is in the orthogonal complement of $S_u = \{\mathbf{f}_{rf,1}, \mathbf{f}_{rf,2} \ldots, \mathbf{f}_{rf,u-1}\}$. With no loss of generality, let $\mathbf{f}_{rf,u}$ be $\mathbf{f}_{rf,u} = \mathbf{U}_u\mathbf{c}_u$, where $\mathbf{U}_u$ is the $N_t \times (N_t - (u-1))$ orthogonal matrix whose column vectors form a basis for the orthogonal complement of S_u and $\mathbf{c}_u$ is an $(N_t - (u-1)) \times 1$ unit vector. The orthogonality constraint in (7) can be removed and it becomes the following design problem,

$$\max \quad \mathbf{c}_u^{\dagger} \mathbf{U}_u^{\dagger} \mathbf{B}_u \mathbf{U}_u \mathbf{c}_u, \quad \text{s.t.} \quad \|\mathbf{c}_u\|^2 = 1. \quad (8)$$

The optimal $\mathbf{c}_u$ is the dominant eigenvector of $\mathbf{U}_u^{\dagger}\mathbf{B}_u\mathbf{U}_u$ and $\mathbf{f}_{rf,u} = \mathbf{U}_u\mathbf{c}_u$. Having designed an RF precoder with orthogonal column vectors, the problem of maximizing the sum of average SNR in (5) becomes

$$\max \quad \sum_{u=1}^{U} \mathbf{f}_{bb,u}^{\dagger}[k] \mathbf{F}_{rf}^{\dagger} \mathbf{B}_u[k] \mathbf{F}_{rf} \mathbf{f}_{bb,u}[k], \quad (9)$$

$$\text{s.t.} \quad \mathbf{F}_{bb}^{\dagger}[k]\mathbf{F}_{bb}[k] = \mathbf{I}_U,$$

where $\mathbf{F}_{bb}[k] = [\mathbf{f}_{bb,1}[k], \mathbf{f}_{bb,2}[k], \ldots, \mathbf{f}_{bb,U}[k]]$. The problem in [9] is similar to the one in (5) and we can follow a similar greedy approach to design the column vectors of $\mathbf{F}_{bb}[k]$ one by one for each subcarrier.

Notice that in the above discussion, we start from the design of the beamformer for the first user and allocate one RF chain in each step. The users whose beamformers are designed first enjoy fewer constraints and thus higher rates for the same channel condition. We may also choose an ordering that takes the quality of service into consideration. For example, in each step we can choose the user that has the smallest transmission rate, i.e. maximizing the worst rate. On the other hand, note that the RF precoder $\mathbf{F}_{rf}$ that is designed as described above does not satisfy the unit modulus constraint that is imposed on the RF precoder for phase-shifter implementation in [1, 3]. It can be implemented using the two-phase-shifter-per-coefficient (THIC) method [21, 22].

Remark. When $N_{rf}^t > U$, we have more RF chains than users. We can allocate the extra RF chains by designing the remaining column vectors of $\mathbf{F}_{rf}$ one by one as before. For fairness, we may use a reverse ordering by starting from the last user in allocating the extra RF chains. As long as $\mathbf{f}_{rf,j}$ in each step is orthogonal to the column vectors that have already been designed, the column vectors of $\mathbf{F}_{rf}$ are orthogonal.

4. HYBRID COMBINER DESIGN

We assume that each mobile user knows its own channel as well as the channel statistics, which can be used to derive the precoder used at the transmitter. The assumption is reasonable when the channel statistics are changing slowly and can be sent to the users. The user can derive the precoder based on the statistics. We also assume that $N_{rf}^r \geq U$, i.e. more RF chains than users. Let $\mathbf{H}_{e,u}[k] = \mathbf{H}_u[k]\mathbf{F}[k]$ be the equivalent channel for the kth subcarrier of the uth user. To remove interference, the kth subcarrier combiner of the uth user $\mathbf{g}_u[k]$ needs to satisfy $\mathbf{g}_u^{\dagger}[k]\mathbf{H}_{e,u}[k] = \mathbf{e}_u^{\dagger}$, where $\mathbf{e}_u$ is an $U \times 1$ standard vector $[\mathbf{e}_u]_u = 1, [\mathbf{e}_u]_j = 0, \forall j \neq u$. Then the optimal zero-forcing combiner for the kth subcarrier is given by

$$\bar{\mathbf{g}}_u[k] = \mathbf{H}_{e,u}[k](\mathbf{H}_{e,u}^{\dagger}[k]\mathbf{H}_{e,u}[k])^{-1}\mathbf{e}_u, \quad (10)$$

for $k = 0, \ldots, M-1$. The implementation of the subcarrier combiners given in (10) generally requires a fully digital system, in which the number of RF chains is the same as the number of receive antennas.

For a hybrid implementation, the combiners are of the form $\mathbf{g}_u[k] = \mathbf{G}_{rf,u}\mathbf{g}_{bb,u}[k]$, where $\mathbf{G}_{rf,u}$ is the RF combiner that is common to all the subcarriers while $\mathbf{g}_{bb,u}[k]$, the baseband combiner at the kth subcarrier, can be subcarrier dependent. To obtain $\mathbf{G}_{rf,u}$ and $\mathbf{g}_{bb,u}[k]$ from the ideal $\bar{\mathbf{g}}_u[k]$, in (10) the vector quantization approach in [23] can be used. We can also formulate it as an alternating optimization problem. In particular, let $\bar{\mathbf{G}}_u = [\bar{\mathbf{g}}_{\mathbf{u}}[0], \ldots, \bar{\mathbf{g}}_{\mathbf{u}}[M-1]]$. We would like to find $\mathbf{G}_{rf,u}$ and $\mathbf{G}_{bb,u} = [\mathbf{g}_{bb,u}[0], \ldots, \mathbf{g}_{bb,u}[M-1]]$ to solve

$$\min_{\mathbf{G}_{rf,u}, \mathbf{G}_{bb,u}} \quad \|\bar{\mathbf{G}}_u - \mathbf{G}_{rf,u}\mathbf{G}_{bb,u}\|_F, \quad (11)$$

where $\|\mathbf{A}\|_F$ denotes the Frobenius norm of a matrix $\mathbf{A}$. The problem can be solved using the alternating method given in [6]. For a given initial $\mathbf{G}_{bb,u}$, we can solve for the best $\mathbf{G}_{rf,u}$, which can then be used to find a better $\mathbf{G}_{bb,u}$. The steps can be repeated until further iterations do not improve the objective in (10). Once we have designed $\mathbf{G}_{rf,u}$, the equivalent subcarrier channel becomes $\mathbf{H}_{e,u}^{'}[k] = \mathbf{G}_{rf,u}^{\dagger}\mathbf{H}_u[k]\mathbf{F}[k]$. The zero-forcing condition can be achieved if the baseband combiner $\mathbf{g}_{bb,u}[k]$ satisfies $\mathbf{g}_{bb,u}^{\dagger}[k]\mathbf{H}_{e,u}^{'}[k] = \mathbf{e}_u^{\dagger}$. When $N_{rf}^r \geq U$, we can choose

$$\mathbf{g}_{bb,u}[k] = \mathbf{H}_{e,u}^{'}[k](\mathbf{H}_{e,u}^{'\dagger}[k]\mathbf{H}_{e,u}^{'}[k])^{-1}\mathbf{e}_u, \quad (12)$$

then inter-user interference can be removed completely.

5. SIMULATION RESULTS

Consider a multi-user MIMO-OFDM system with (N_t, N_r) = (64,16), DFT size $M = 128$, the length of cyclic prefix $N_{cp} = 16$ and three users. Each user channel is modeled by (1) with 6 clusters and 10 paths in each cluster. The complex gains for all paths are Gaussian distributed with zero mean. The variances of the complex gains are uniformly distributed over $[0, 1]$. The path delay of each cluster is uniformly distributed over $[0, N_{cp}T_s]$, where $T_s = 1/(30.72 \times 10^6)$ seconds, the same as in [15]. The pulse shaping function $p(t)$ is the raised-cosine filter with roll off factor 0.5. UPA arrays with omnidirectional antennas are assumed. The antenna spacing is half wavelength, thus $d = 1/2$. The means of AoD and AoA are uniformly distributed over $[0, 2\pi]$. We have used 3×10^4 channels in the simulations.

Fig. 1. shows the average transmission rate of all users as a function of angular spread ($\sigma_{\theta^t_{u,\ell}} = \sigma_{\phi^t_{u,\ell}}$) for P_t/N_0=0 dB when the proposed hybrid statistical beamforming is used for $(N^t_{rf}, N^r_{rf}) = (3, 3)$ and $(N^t_{rf}, N^r_{rf}) = (3, 5)$. The proposed orthogonal hybrid precoding system with $N^t_{rf} = 3$ and $N^r_{rf} = 3$ is within 0.2 bit/Hz of a fully-digital statistical system, in which each antenna is endowed with an RF chain. The degradation due to hybrid structure is a small one. Also, we see that by increasing N^r_{rf} to 5, the rate is improved. This is because the statistical system relies on the receiver to achieve zero-forcing; having more RF chains allows more design freedom and thus better performance. Notice that statistical precoders are designed based on statistics of AoD. When angular spread increases, there is a larger variation on the instantaneous AoD and thus a larger rate loss, as shown in Fig 1. We have also shown the performance of the statistical system when the orthogonality condition is not incorporated in the design of subcarrier precoders (labeled as 'stat, hybrid, non-ortho'). We can see that sending the signals in orthogonal beams alleviates the interference among different users and a higher rate can be achieved.

Fig 2 shows the transmission rate of the statistical precoding system as a function of SNR P_t/N_0. For comparison, we also show the rates of three MIMO-OFDM systems with full CSI, the fully digital system in [24] and the hybrid system in [6] and [15]. In all hybrid systems considered, $(N^t_{rf}, N^r_{rf}) = 3$. In [6], an alternating approach is used to design hybrid precoders, for which ideal subcarrier precoders are computed first and alternating method applied. In [15], the RF precoder is designed using this statistics of subcarrier covariance matrices while the baseband precoders are zero-forcing. Compared to the two hybrid systems with full CSI, the proposed statistical precoder achieves a performance close to [15] and it is around 0.7 bits/Hz away from [6]. The knowledge of channel statistics at both the transmitter and receivers greatly reduce the interference and satisfactory performance can be obtained even though there is no instantaneous feedback.

6. CONCLUSION

In this paper, we consider statistical hybrid precoding for multiuser MIMO-OFDM systems over mmWave channels. There is no instantaneous feedback and the transmitter cannot be zero-forcing, as in most earlier works on multiuser communications. Exploiting the statistics of the user channels, we design the precoders so that different user signals are sent on orthogonal beams. The use of orthogonal beam allows the user signals to be spatially separated, which greatly alleviates the inter-user interference due to the sparse nature of mmWave channels. Simulations show that the performance of the proposed statistical hybrid precoding system yields a rate close to that of a full CSI systems.

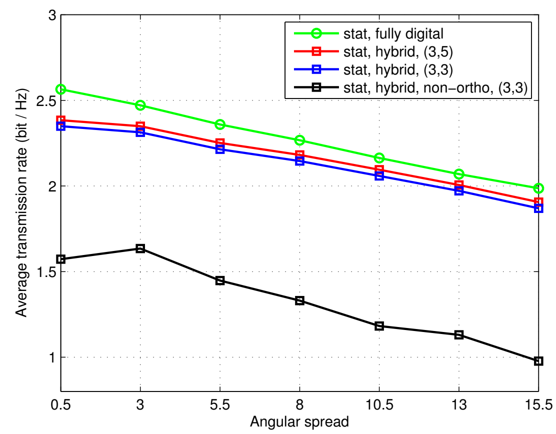

Fig. 1. Average transmission rate performance of statistical hybrid MIMO-OFDM systems v.s angular spread for $P_t/N_0 = 0$ dB.

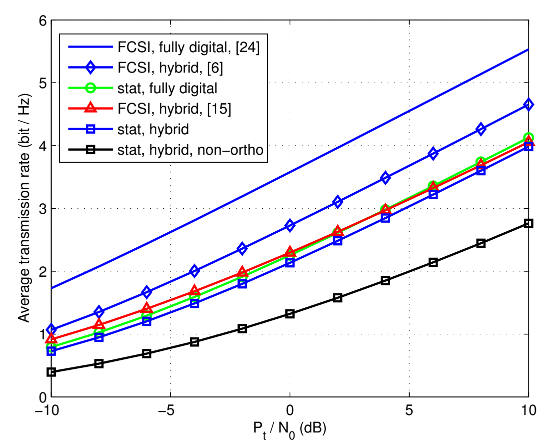

Fig. 2. Performance comparison of hybrid precoding MIMO-OFDM systems.

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