Transmit Beamforming Strategies with Iterative Equalization for Hybrid mmW Systems

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Abstract– **The aim of this manuscript is to compare the performance of two transmit beamforming approaches with an iterative equalizer for millimeter wave (mmW) systems. At the transmitter, we assume two hybrid beamforming: a sparse beamforming approach recently proposed and fixed random beamforming which the coefficients are computed independently of the instantaneous channel realization. At the receiver, we consider a hybrid iterative block space-time structure to efficiently separate the spatial streams. We consider that both the transmitter and the received are equipped with a large antenna array and the number of radio frequency (RF) chains is lower that the number of antennas. The hardware limitations impose several constraints in the analog domain that are considered when the transmit beamforming/precoder is generated. Our performance results have shown that the performance of the sparse precoder is better for low signal-to-noise ratio (SNR) regime while random precoder outperforms sparse based one for medium to high SNR regime.**

Keywords—massive MIMO, mmWave communications, iterative block equalization, hybrid analog/digital architectures.

I. Introduction

The use of a large number of antennas makes achievable higher data rates for future wireless networks [1]. Additionally, the global bandwidth shortage facing wireless carriers has motivated the exploration of the underutilized millimeter wave (mmW) frequency spectrum for future broadband cellular communication networks [2] and that allows the access to more bandwidth. Due to the small wavelength, the use of mmW with massive MIMO (mMIMO) is very attractive, since the terminals can be equipped with large number of antennas very compacted [3].

In mmW, the channel is very correlated that is critical for a mMIMO implementation, where propagation tends to be line of sight (LOS) or near-LOS. Thus, the use of beamforming to minimize de interference is very important for these systems [3]. MmW massive MIMO systems may exploit new and efficient spatial processing techniques [4], but the design at these techniques should follow different approaches than the ones adopted for lower frequencies counterparts, mainly due to the hardware limitations [5]. The high cost and power consumption of some mmW mixed-signal components, make it difficult to have a fully dedicated radio frequency (RF) chain for each antenna [6] as in conventional MIMO systems [7] and to overcome these limitations, hybrid analog/digital architectures were proposed. At these architectures, some signal processing is done at the digital level and some left to the analog domain, as discussed in [8].

Some beamforming and/or combining/equalization schemes have been proposed for hybrid architectures [9]-[13]. A precoding scheme based on the knowledge of partial channel information at both terminals, in the form of angles of arrival (AoA) and departure (AoD), was proposed in [9]. The authors in [10] designed a joint digital and analog beamforming at the transmitter side, where first a set of fixed analog beamforming coefficients is selected and then a digital eigenmode based precoder is computed, but they consider a fully digital receiver. In [11], a hybrid spatially sparse precoding and combining approach was proposed for mmW massive MIMO systems. The spatial structure of mmW channels was exploited to formulate the single-user multi-stream precoding/combining scheme as a sparse reconstruction problem. A digitally assisted analog beamforming technique for mmW systems was considered in [12], where a digital beamsteering system using coarsely quantized signals assists the analog beamformer. In [13], a turbo-like beamforming was proposed to jointly compute the transmit and receive analog beamforming coefficients, but the digital processing part was not taken into account.

Nonlinear equalizers are considered to efficiently separate the spatial streams in the current MIMO based networks [14]. Iterative block decision feedback equalization (IB-DFE) approach was originally proposed in [15] and it is one of the most promising nonlinear equalization schemes [14]. IB-DFE can be regarded as a low complexity turbo equalizer implemented in the frequency-domain that does not require the channel decoder output in the feedback loop. The IB-DFE principles can be used in mmW massive MIMO context to efficiently separate the spatial streams. However, as discussed, mmW massive MIMO brings new major challenges that prevent a direct plug and play of the iterative equalization based solutions developed for conventional fully digital MIMO systems.

In this paper, we evaluate the performance of two transmit beamforming strategies combined with an efficient iterative block space-time equalizer for hybrid mmW massive MIMO systems. In the first strategy, we consider fixed random precoders computed without the knowledge of the channel state information (CSI) keeping the transmitter with very low com-

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plexity. In the second one, we assume a sparse transmit beamforming recently proposed in [11]. At the transmitter side, a space-time encoder structure is employed, before the digital and analog precoders, to 1) ensure that the transmit signal and consequently the noise plus interference, at the receiver side, are Gaussian distributed (which simplifies the receiver optimization), 2) warrant that the signal to interference plus noise ratio (SINR) is independent of each spatial stream and time slot and 3) increase the inherent diversity of the mmW massive MIMO system. At the receiver, we design a hybrid iterative block space-time structure to efficiently separate the spatial streams. We assume a fully connected hybrid architecture where each RF chains are connected to all antennas. The analog and digital parts of the hybrid equalizer are jointly optimized using as a metric the mean square error (MSE) between the transmitted data vector and its estimate after the digital equalizer. The specificities of the analog domain impose several constraints in the joint optimization. To efficiently deal with the constraints the analog part is selected from a dictionary based on the array response vectors.

The remainder of the paper is organized as follows: Section II describes the hybrid mmW massive MIMO systems model. Section III, presents the random precoder. Section IV, starts by briefly describing the iterative space-time receiver structure. Then, the fully digital equalizer is presented and finally the proposed hybrid space-time equalizer is derived in detail. Section V presents the main performance results and the conclusions will be drawn in section VI.

Notations: Boldface capital letters denote matrices and boldface lowercase letters denote column vectors. The operation $(.)^H$ represents the Hermitian transpose of a matrix. Consider a vector $\mathbf{a}$ and a matrix $\mathbf{A}$, then diag($\mathbf{a}$) and diag($\mathbf{A}$) correspond to a diagonal matrix with diagonal entries equal to vector $\mathbf{a}$ and a diagonal matrix with entries equal to the diagonal entries of the matrix $\mathbf{A}$, respectively. $\mathbf{A}(j,l)$ denotes the element at row j and column l of the matrix $\mathbf{A}$. $\mathbf{I}_N$ is the identity matrix with size $N\times N$.

II. System Characterization

In this section, we present the mmW massive MIMO signal definition, the transmitter and receiver characterization. We consider a hybrid based architecture, as shown in Fig 1. Furthermore, we assume a single-user mmW system with N_t transmit antennas and N_r receive antennas, where the transmitter sends N_s data streams to the receiver, per time-slot.

We considered a clustered channel, $\mathbf{H}\in\mathbb{C}^{N_r\times N_t}$, that is the sum of the contribution of N_{cl} clusters, each of which contribute N_{ray} propagation paths which follows the clustered sparse mmW channel model discussed in [11]. It may be expressed as

$$\mathbf{H}=\mathbf{A}_r\mathbf{\Lambda}\mathbf{A}_t^H \tag{1}$$

where $\mathbf{\Lambda}$ is a diagonal matrix, with entries (j,l) corresponding to the paths gains of the lth ray in the ith scattering cluster. $\mathbf{A}_t=[\mathbf{a}_t(\theta_{1,1}^t),\ldots,\mathbf{a}_t(\theta_{N_{cl},N_{ray}}^t))]$, $\mathbf{A}_r=[\mathbf{a}_r(\theta_{1,1}^r),\ldots,\mathbf{a}_r(\theta_{N_{cl},N_{ray}}^r))]$ are the matrix of array response vectors at the transmitter and receiver, whereas $\theta_{j,l}^r$ and $\theta_{j,l}^t$ are the azimuth angles of arrival and departure, respectively. The channel path gains and the angles are generated according to the random distributions discussed in [11]. We consider a block fading channel, i.e., the channel remains constant during a block, with size T, but it varies independently between blocks.

III. Random Precoder

In this section, we present a low complexity transmitter. We assume that the transmitter have no access to CSI simplifying the overall system design. The transmitter processing is decomposed into two parts, the digital baseband and the analog circuitry that are modeled mathematically by precoder matrices $\mathbf{F}_a\in\mathbb{C}^{N_t\times N_t^{RF}}$ and $\mathbf{F}_d\in\mathbb{C}^{N_t^{RF}\times N_s}$, respectively. The digital part has N_t^{RF} transmit chains, with $N_s\le N_t^{RF}\le N_t$. Due to hardware constraints, the analog part is implemented using a matrix of analog phase shifters, which force all elements of matrix $\mathbf{F}_a$ to have equal norm $(|\mathbf{F}_a(i,l)|^2=N_t^{-1})$. As such the analog precoder matrix is generated randomly accordingly to

$$\mathbf{F}_a=[\mathrm{e}^{j2\pi\phi_{n,p}}]_{1\le n\le N_t,1\le p\le N_t^{RF}}, \tag{2}$$

where $\phi_{n,p}$, $n\in\{1,\ldots,N_t\}$, $p\in\{1,\ldots,N_t^{RF}\}$ are i.i.d uniform random variables with support $\phi_{n,p}\in[0,1]$.

We assume that all RF resources are used by transmitter, i.e., $N_s=N_t^{RF}$, and then we can assume that digital precoder is equal to identity matrix, $\mathbf{F}_d=\mathbf{I}_{N_s}$. The transmitter total power constraint is $\|\mathbf{X}\|_F^2=N_sT$. The transmit signal is given by

$$\mathbf{X}=\mathbf{F}_a\mathbf{F}_d\mathbf{C}, \tag{3}$$

where $\mathbf{C}=[\mathbf{c}_1,\ldots,\mathbf{c}_T]\in\mathbb{C}^{N_s\times T}$ denotes a codeword constructed by using a space-time block code (STBC) that can be mathematically described by

$$\mathbf{z}_t=\mathbf{S}\mathbf{f}_t, \tag{4}$$

$$\mathbf{c}_t=\mathbf{\Pi}_t\mathbf{z}_t, \tag{5}$$

where $t=1,\ldots,T$ denotes the time index, $\mathbf{f}_t\in\mathbb{C}^T$ denotes column t of a T-point DFT matrix $(\mathbf{F}_T=[\mathbf{f}_1,\ldots,\mathbf{f}_T])$, $\mathbf{\Pi}_t\in\mathbb{C}^{N_s\times N_s}$, $t=1,\ldots,T$ is a random permutation matrix, known both at the transmitter and receiver sides and $\mathbf{S}=[s_{s,t}]_{1\le s\le N_s,1\le t\le T}\in\mathbb{C}^{N_s\times T}$, with $s_{t,s}$, $t\in\{1,\ldots,T\}$, $s\in\{1,\ldots,N_s\}$ denoting a complex data symbol chosen from a QAM constellation with $\mathbb{E}[|s_{s,t}|^2]=\sigma_s^2$, where $\sum_{s=1}^{N_s}\sigma_s^2=N_s$. For the sake of simplicity and, without loss of generality, in this work we consider only QPSK constellations. To compute codeword $\mathbf{C}$ we need to apply an FFT transform to the rows of the symbol matrix $\mathbf{S}$ (see (4)) and then permute each of the resulting T columns with a random permutation $\mathbf{\Pi}_t, t=1,\ldots,T$ (see (5)).

IV. Hybrid Iterative Space-Time Receiver Design

In this section, we derive the hybrid iterative block space-time feedback equalizer shown in Fig. 2. We start by designing the fully digital receiver, that can serve as lower bound for the hybrid one and then a detailed formulation of the iterative approach is presented. The received signal is given by

$$\mathbf{Y}=\mathbf{HX}+\mathbf{N}, \tag{6}$$

where $\mathbf{Y}=[\mathbf{y}_1,\ldots,\mathbf{y}_T]\in\mathbb{C}^{N_r\times T}$ denotes the received signal matrix, $\mathbf{X}=[\mathbf{x}_1,\ldots,\mathbf{x}_T]\in\mathbb{C}^{N_t\times T}$ is the transmitted signal and $\mathbf{N}=[\mathbf{n}_1,\ldots,\mathbf{n}_T]\in\mathbb{C}^{N_r\times T}$ a zero mean Gaussian noise with variance σ_n^2. The received signal is firstly processed through the analog phase shifters, modeled by the matrix $\mathbf{W}_a\in\mathbb{C}^{N_r^{RF}\times N_r}$, then follows the baseband processing, composed of N_r^{RF} processing chains. All elements of the matrix $\mathbf{W}_a$ must have equal norm $(|\mathbf{W}_a(j,l)|^2=N_r^{-1})$. Specifically, the baseband processing includes a digital feedback closed-loop comprising a forward and a feedback path. For the forward path the signal first passes through a linear filter $\mathbf{W}_d\in\mathbb{C}^{N_s\times N_r^{RF}}$, then follows the decoding of the STBC (demodulation included). In the feedback path, the data recovered in the forward path is first modulated and encoded using the STBC, then it passes through the feedback matrix $\mathbf{B}_d\in\mathbb{C}^{N_s\times N_s}$. The encoding of the STBC follows (4) and (5), and its decoding obeys

$$\tilde{\mathbf{Z}}=[\mathbf{\Pi}_1^H\tilde{\mathbf{c}}_1,\ldots,\mathbf{\Pi}_T^H\tilde{\mathbf{c}}_T], \tag{7}$$

$$\tilde{\mathbf{S}}=\tilde{\mathbf{Z}}\mathbf{F}_T^H. \tag{8}$$

The feedback and feedforward paths are combined by subtracting the signal output of the feedback path from the filtered received signal $\mathbf{W}_d\mathbf{W}_a\mathbf{Y}$. At the *i*th iteration the received signal at the *t*th time slot, after the de-interleaver, is given by

$$\tilde{\mathbf{z}}_t^{(i)}=\mathbf{\Pi}_t^H(\mathbf{W}_{d,t}^{(i)}\mathbf{W}_{a,t}^{(i)}\mathbf{y}_t-\mathbf{B}_{d,t}^{(i)}\mathbf{\Pi}_t\hat{\mathbf{z}}_t^{(i-1)}), \tag{9}$$

$$\hat{\mathbf{Z}}^{(i-1)}=\hat{\mathbf{S}}^{(i-1)}\mathbf{F}_T, \tag{10}$$

where $\mathbf{\Pi}_t^H\in\mathbb{C}^{N_s\times N_s}$ is the de-interleaver matrix and $\hat{\mathbf{Z}}^{(i-1)}=[\hat{\mathbf{z}}_1^{(i-1)},\ldots,\hat{\mathbf{z}}_T^{(i-1)}]\in\mathbb{C}^{N_s\times T}$ is the DFT of the detector output $\hat{\mathbf{S}}^{(i-1)}$. The matrix $\hat{\mathbf{C}}^{(i)}=[\mathbf{\Pi}_1\hat{\mathbf{z}}_1^{(i)},\ldots,\mathbf{\Pi}_T\hat{\mathbf{z}}_T^{(i)}]$ is the hard estimate of the transmitted codeword $\mathbf{C}$ and $\hat{\mathbf{S}}^{(i)}=\mathrm{sign}\left(\tilde{\mathbf{S}}^{(i)}\right)$ the hard decision associated to QPSK data symbols $\mathbf{S}$, at iteration *i*.

From the central limit theorem the entries of vector $\mathbf{z}_t$, $t\in\{1,\ldots,T\}$ are Gaussian distributed, then as the input-output relationship between variables $\mathbf{z}_t$ and $\hat{\mathbf{z}}_t^{(i)}$, $t\in\{1,\ldots,T\}$ is memoryless, follows

$$\hat{\mathbf{z}}_t^{(i)}=\mathbf{\Psi}^{(i)}\mathbf{z}_t+\hat{\epsilon}_t^{(i)},\quad t\in\{1,\ldots,T\}, \tag{11}$$

where $\mathbf{\Psi}^{(i)}$ is a diagonal matrix given by

$$\mathbf{\Psi}^{(i)}=\mathrm{diag}\left(\psi_1^{(i)},\ldots,\psi_s^{(i)},\ldots,\psi_{N_s}^{(i)}\right), \tag{12}$$

$$\psi_s^{(i)}=\frac{\mathbb{E}\left[\hat{\mathbf{z}}_t^{(i)}(s)\mathbf{z}_t^*(s)\right]}{\mathbb{E}\left[|\mathbf{z}_t(s)|^2\right]},\ s\in\{1,\ldots,N_s\}, \tag{13}$$

and $\hat{\epsilon}_t^{(i)}$ is a zero mean error vector uncorrelated with $\mathbf{z}_t$, $t\in\{1,\ldots,T\}$, with $\mathbb{E}\left[\hat{\epsilon}_t^{(i)}\hat{\epsilon}_t^{(i)^H}\right]=\left(\mathbf{I}_{N_s}-\left|\mathbf{\Psi}^{(i)}\right|^2\right)\sigma_s^2$, and then it can be proven that the average error power is given by

$$\begin{aligned}\mathrm{MSE}_t^{(i)}&=\mathbb{E}[\|\tilde{\mathbf{z}}_t^{(i)}-\mathbf{z}_t\|^2]\\&=\left\|(\mathbf{W}_{ad,t}^{(i)})^{\Pi}\mathbf{H}_t^{\Pi}-\mathbf{I}_{N_s}-(\mathbf{B}_{d,t}^{(i)})^{\Pi}\mathbf{\Psi}^{(i-1)}\right\|_F^2\sigma_s^2\\&+\left\|(\mathbf{B}_{d,t}^{(i)})^{\Pi}(\mathbf{I}_{N_s}-|\mathbf{\Psi}^{(i-1)}|^2)^{1/2}\right\|_F^2\sigma_s^2+\left\|(\mathbf{W}_{ad,t}^{(i)})^{\Pi}\right\|_F^2\sigma_n^2,\end{aligned} \tag{14}$$

where $\mathbf{H}_t^{\Pi}=\mathbf{HF}_a\mathbf{F}_d\mathbf{\Pi}_t$, $(\mathbf{W}_{d,t}^{(i)})^{\Pi}=\mathbf{\Pi}_t^H\mathbf{W}_{d,t}^{(i)}\mathbf{\Pi}_t$, $(\mathbf{W}_{a,t}^{(i)})^{\Pi}=\mathbf{\Pi}_t^H\mathbf{W}_{a,t}^{(i)}$, $(\mathbf{W}_{ad,t}^{(i)})^{\Pi}=(\mathbf{W}_{d,t}^{(i)})^{\Pi}(\mathbf{W}_{a,t}^{(i)})^{\Pi}$ and $(\mathbf{B}_{d,t}^{(i)})^{\Pi}=\mathbf{\Pi}_t^H\mathbf{B}_{d,t}^{(i)}\mathbf{\Pi}_t$.

A. *Design of Digital Iterative Space-time Receiver*

Firstly, we design the fully digital iterative space-time receiver based on the IB-DFE principles. The performance of this approach can be regarded as a lower bound for the hybrid iterative block equalizer designed in the next section. The equalizer is designed by minimizing the MSE

$$\begin{aligned}&\left((\mathbf{W}_{ad,t}^{(i)})_{opt}^{\Pi},(\mathbf{B}_{d,t}^{(i)})_{opt}^{\Pi}\right)=\arg\min\ \mathrm{MSE}_t^{(i)}\\&\quad\text{s.t.}\ \sum_{t=1}^{T}\mathrm{diag}((\mathbf{W}_{ad,t}^{(i)})^{\Pi}\mathbf{H}_t^{\Pi})=T\mathbf{I}_{N_s}.\end{aligned} \tag{15}$$

In this case, the number of receiver RF chains is equal to the number of receiver antennas, and thus we only have a digital

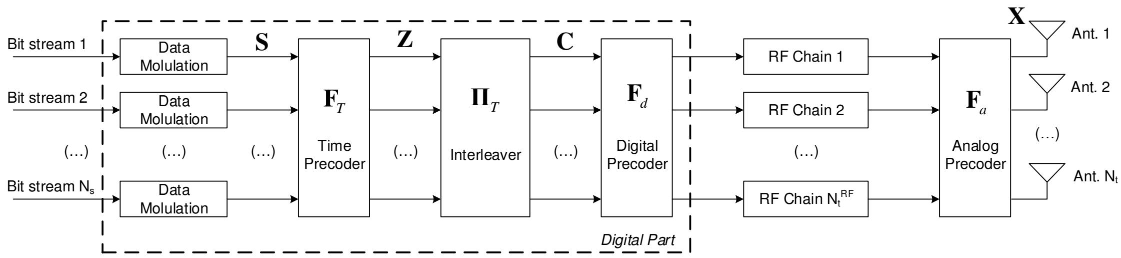

Fig 1. Transmitter block diagram.

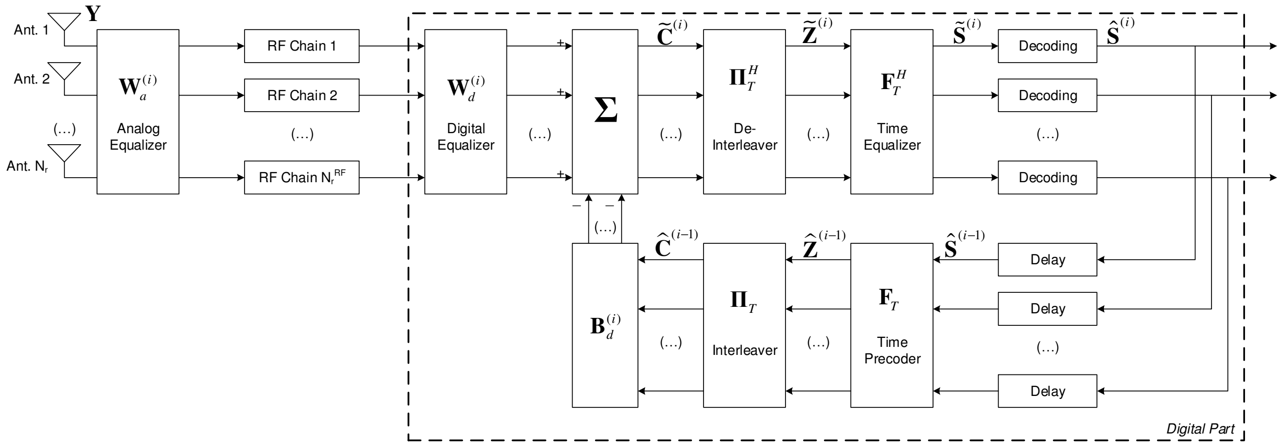

Fig 2. Receiver block diagram.

linear feedforward filter referred as $\mathbf{W}_{ad,t}$ and a feedback filter $\mathbf{B}_{d,t}$. The solution to the optimization problem (15) is

$$(\mathbf{W}^{(i)}_{ad,t})^{\Pi}_{opt} = \mathbf{\Omega}\left(\mathbf{R}_t^{(i-1)}\right)^{-1}(\mathbf{H}_t^{\Pi})^H , \quad (16)$$

$$(\mathbf{B}^{(i)}_{d,t})^{\Pi}_{opt} = \left((\mathbf{W}^{(i)}_{ad,t})^{\Pi}_{opt}\mathbf{H}_t^{\Pi} - \mathbf{I}_{N_s}\right)\left(\mathbf{\Psi}^{(i-1)}\right)^H , \quad (17)$$

$$\mathbf{\Omega} = T\left(\sum_{t=1}^{T}\mathrm{diag}\left(\left(\mathbf{R}_t^{(i-1)}\right)^{-1}(\mathbf{H}_t^{\Pi})^H\mathbf{H}_t^{\Pi}\right)\right)^{-1} , \quad (18)$$

$$\mathbf{R}_t^{(i-1)} = (\mathbf{H}_t^{\Pi})^H\mathbf{H}_t^{\Pi}(\mathbf{I}_{N_s} - |\mathbf{\Psi}^{(i-1)}|^2) + \sigma_n^2\sigma_s^{-2}\mathbf{I}_{N_s} . \quad (19)$$

B. Design of Hybrid Iterative Space-time Receiver

In this section, we design the hybrid iterative block space-time receiver. Clearly, the previous optimization problem of (15) does not take into account the analog domain constraints. Let us denote by $\mathcal{W}_a$ the set of feasible RF equalizers, i.e., the set of $N_t \times N_t^{RF}$ matrices with constant-magnitude entries, then the reformulated optimization problem for the hybrid iterative equalizer is as follows

$$\left((\mathbf{W}^{(i)}_{a,t})^{\Pi}_{opt}, (\mathbf{W}^{(i)}_{d,t})^{\Pi}_{opt}, (\mathbf{B}^{(i)}_{d,t})^{\Pi}_{opt}\right) = \arg\min \mathrm{MSE}_t^{(i)}$$
$$\text{s.t.} \sum_{t=1}^{T}\mathrm{diag}((\mathbf{W}^{(i)}_{d,t})^{\Pi}(\mathbf{W}^{(i)}_{a,t})^{\Pi}\mathbf{H}_t^{\Pi}) = T\mathbf{I}_{N_s} \quad (20)$$
$$(\mathbf{W}^{(i)}_{a,t})^{\Pi} \in \mathcal{W}_a .$$

Due to the digital nature of the feedback equalizer $(\mathbf{B}^{(i)}_{d,t})^{\Pi}$ and since the new constraint does not impose any restriction on this matrix, the feedback equalizer for the hybrid iterative equalizer is similar to the fully digital iterative equalizer discussed in the previous section, and thus given by

$$(\mathbf{B}^{(i)}_{d,t})^{\Pi}_{opt} = \left((\mathbf{W}^{(i)}_{d,t})^{\Pi}_{opt}(\mathbf{W}^{(i)}_{a,t})^{\Pi}_{opt}\mathbf{H}_t^{\Pi} - \mathbf{I}_{N_s}\right)\left(\mathbf{\Psi}^{(i-1)}\right)^H . \quad (21)$$

From (14) and (21), the MSE expression simplifies to is equal (up to a constant) to

$$\overline{\mathrm{MSE}}_t^{(i)} = \left\|\left((\mathbf{W}^{(i)}_{d,t})^{\Pi}(\mathbf{W}^{(i)}_{a,t})^{\Pi} - (\overline{\mathbf{W}}^{(i)}_{ad,t})^{\Pi}_{opt}\right)\left(\tilde{\mathbf{R}}_t^{(i-1)}\right)^{1/2}\right\|_F^2 , \quad (22)$$

$$\tilde{\mathbf{R}}_t^{(i-1)} = \mathbf{H}_t^{\Pi}(\mathbf{I}_{N_s} - |\mathbf{\Psi}^{(i-1)}|^2)(\mathbf{H}_t^{\Pi})^H + \sigma_n^2\sigma_s^{-2}\mathbf{I}_{N_r} , \quad (23)$$

$$(\overline{\mathbf{W}}^{(i)}_{ad,t})^{\Pi}_{opt} = (\mathbf{I}_{N_s} - |\mathbf{\Psi}^{(i-1)}|^2)\mathbf{\Omega}^{-1}(\mathbf{W}^{(i)}_{ad,t})^{\Pi}_{opt} , \quad (24)$$

where $(\overline{\mathbf{W}}^{(i)}_{ad,t})^{\Pi}_{opt}$ and $\tilde{\mathbf{R}}_t^{(i-1)}$ denote a non-normalized version of the optimum fully digital feedforward matrix and the correlation of the ISI plus channel noise.

Due to the non-convex nature of the feasible set $\mathcal{W}_a$, an analytical solution to the problem (20) is difficult to obtain, if not impossible. Nevertheless, we find an approximate solution to problem (20) by assuming that the matrix $(\mathbf{W}^{(i)}_{a,t})^{\Pi}$ is a N_r^{RF} sparse linear combination of vectors $\mathbf{a}_{r,u}(\theta_{j,l}^{r,u})$ or equivalently a N_r^{RF} sparse linear combination of the columns of matrix $\mathbf{A}_r = [\mathbf{A}_{r,1}, \ldots, \mathbf{A}_{t,U}]$. We may say that $(\mathbf{W}^{(i)}_{a,t})^{\Pi}$ has a N_r^{RF} term representation over the dictionary $\mathbf{A}_r$. Therefore, optimization problem can be approximated as follows

$$(\ddot{\mathbf{W}}^{(i)}_{d,t})^{\Pi}_{opt}$$
$$= \arg\min \left\|\left((\ddot{\mathbf{W}}^{(i)}_{d,t})^{\Pi}\mathbf{A}_r^H - (\overline{\mathbf{W}}^{(i)}_{ad,t})^{\Pi}_{opt}\right)\left(\tilde{\mathbf{R}}_t^{(i-1)}\right)^{1/2}\right\|_F^2$$
$$\text{s.t.} \sum_{t=1}^{T}\mathrm{diag}((\ddot{\mathbf{W}}^{(i)}_{d,t})^{\Pi}\mathbf{A}_r^H\mathbf{H}_t^{\Pi}) = T\mathbf{I}_{N_s} \quad (25)$$
$$\left\|\mathrm{diag}(((\ddot{\mathbf{W}}^{(i)}_{d,t})^{\Pi})^H(\ddot{\mathbf{W}}^{(i)}_{d,t})^{\Pi})\right\|_0 = N_t^{RF} ,$$

where $\left\|\mathrm{diag}(((\ddot{\mathbf{W}}^{(i)}_{d,t})^{\Pi})^H(\ddot{\mathbf{W}}^{(i)}_{d,t})^{\Pi})\right\|_0 = N_t^{RF}$ represents the sparsity constraint and enforces that only N_r^{RF} columns of matrix $(\ddot{\mathbf{W}}^{(i)}_{d,t})^{\Pi}$ are non-zero. The optimum digital feedforward matrix $(\mathbf{W}^{(i)}_{d,t})^{\Pi}_{opt}$ is obtained from $(\ddot{\mathbf{W}}^{(i)}_{d,t})^{(i)}_{opt} = [(\mathbf{W}^{(i)}_{d,t})^{\Pi}_{opt}, \mathbf{0}]$ by removing the zero columns and the optimum analogue feedforward matrix $(\mathbf{W}^{(i)}_{a,t})^{\Pi}_{opt}$ is obtained from $\mathbf{A}_r^H$ by selecting the rows corresponding to the non-zero columns of $(\ddot{\mathbf{W}}^{(i)}_{d,t})^{\Pi}$.

From optimality condition (associated Lagrangian equal to zero), we obtain $\mathbf{W}^{(i)}_{res,t}$ that is the residue matrix that is given by

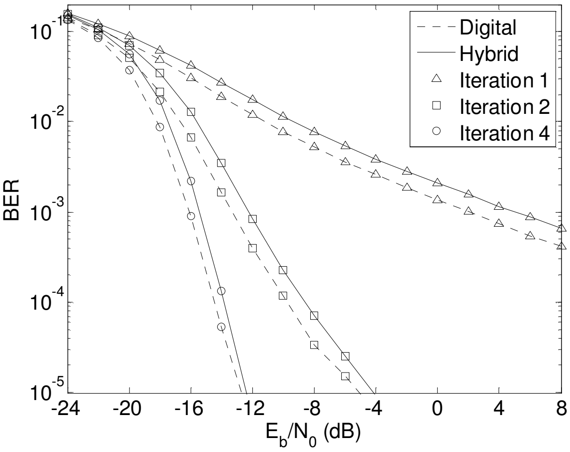

Fig 3. Performance for sparse precoder.

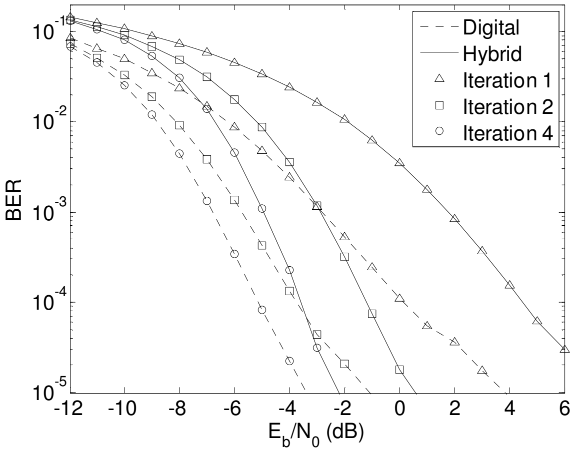

Fig 4. Performance for hybrid random precoder.

$$\mathbf{W}_{res,t}^{(i)} = \left((\ddot{\mathbf{W}}_{d,t}^{(i)})^{\Pi}\mathbf{A}_r^H - (\overline{\mathbf{W}}_{ad,t}^{(i)})^{\Pi}_{opt}\right)\left(\tilde{\mathbf{R}}_t^{(i-1)}\right) + \mathbf{U}_d(\mathbf{H}_t^{\Pi})^H . \quad (26)$$

From the definition of matrices $\tilde{\mathbf{R}}_t^{(i-1)}$, $\mathbf{R}_t^{(i-1)}$, $(\overline{\mathbf{W}}_{ad,t}^{(i)})^{\Pi}_{opt}$ and $(\mathbf{W}_{ad,t}^{(i)})^{\Pi}_{opt}$ equation (26) simplifies to

$$\mathbf{W}_{res,t}^{(i)} = (\ddot{\mathbf{W}}_{d,t}^{(i)})^{\Pi}\mathbf{A}_r^H\tilde{\mathbf{R}}_t^{(i-1)} - \mathbf{\Omega}_d(\mathbf{H}_t^{\Pi})^H , \quad (27)$$

where $\mathbf{\Omega}_d = \mathbf{I}_{N_s} - |\mathbf{\Psi}^{(i-1)}|^2 + \mathbf{U}_d$ denotes a redefined Lagrangian multipliers matrix, that must be selected so that the constraint of the optimization problem (25) is respected. The matrix $\mathbf{U}_d = \mathrm{diag}(\mu_1,\ldots,\mu_{N_s})$ is a diagonal matrix where μ_s, $s\in\{1,\ldots,N_s\}$ are the Lagrange multipliers.

To enforce the sparsity constraint, the best columns of the dictionary $\mathbf{A}_r$ are selected using an iterative greedy method. At each iteration the column of $\mathbf{A}_r$ that is most correlated with the actual value of the residue $\mathbf{W}_{res,t}^{(i)}$ is selected. In the first iteration, the residue is set to the trivial value $\mathbf{W}_{res,t}^{(i)} = -(\overline{\mathbf{W}}_{ad,t}^{(i)})^{\Pi}_{opt}\left(\tilde{\mathbf{R}}_t^{(i-1)}\right)$. Then, after identifying a set of columns of the matrix $\mathbf{A}_r$ (one column per iteration) to form the analog feedforward equalizer matrix $(\mathbf{W}_{a,t}^{(i)})^{\Pi}_{opt}$, we obtain the optimum digital feedforward equalizer matrix $(\mathbf{W}_{d,t}^{(i)})^{\Pi}_{opt}$ using the orthogonality condition. It can be proven that the optimum digital feedforward matrix is

$$(\mathbf{W}_{d,t}^{(i)})^{\Pi}_{opt} = \mathbf{\Omega}_d\left((\mathbf{W}_{a,t}^{(i)})^{\Pi}_{opt}\mathbf{H}_t^{\Pi}\right)^H\left(\mathbf{R}_{d,t}^{(i-1)}\right)^{-1}, \quad (28)$$

where $\mathbf{R}_{d,t}^{(i-1)} = (\mathbf{W}_{a,t}^{(i)})^{\Pi}_{opt}\tilde{\mathbf{R}}_t^{(i-1)}((\mathbf{W}_{a,t}^{(i)})^{\Pi}_{opt})^H$ and to respect the constraint of problem (25) $\mathbf{\Omega}_d$ is given by

$$\mathbf{\Omega}_d = T\left(\sum_{t=1}^{T}\mathrm{diag}\left(\left((\mathbf{W}_{a,t}^{(i)})^{\Pi}_{opt}\mathbf{H}_t^{\Pi}\right)^H\left(\mathbf{R}_{d,t}^{(i-1)}\right)^{-1} \times(\mathbf{W}_{a,t}^{(i)})^{\Pi}_{opt}\mathbf{H}_t^{\Pi}\right)\right)^{-1}, \quad (29)$$

After obtaining the optimum value of the digital feedforward matrix $(\mathbf{W}_{d,t}^{(i)})^{\Pi}$ the residue matrix (26) is updated. The previous steps iterate on the updated residue value to obtain the N_r^{RF} index set to index the dictionary $\mathbf{A}_r$.

The proposed iterative hybrid space-time equalizer is identical to the equalizer proposed in [11] when the block length is equal to one $(T=1)$ and for iteration one $(i=1)$.

V. Performance Results

In this section, we access the performance of the two transmit beamforming approaches combined with the iterative space-time equalizer. We consider a clustered channel model with $N_{cl}=8$ clusters, each with $N_{ray}=10$ rays, with Laplacian distributed azimuth angles of arrival and departure. The average power of all N_{cl} clusters is the same and the angle spread at both the transmitter and receiver is set to 8 degrees. We assume that the transmitter's sector angle is 60º wide in the azimuth domain and the receiver antenna array has omnidirectional antenna elements. The antenna element spacing is assumed to be half-wavelength. The channel remains constant during a block, with size $T=32$, and takes independent values between blocks.

We present results for a scenario whose the parameters are $N_r=32$, $N_t=128$, $N_s=8$, $N_r^{RF}=N_t^{RF}=8$. These results are presented for iteration 1, 2 and 4 of the digital and hybrid iterative space-time receivers, which are referred as *digital* and *hybrid*, in the following.

The performance metric considered is the BER, which is presented as a function of the E_b/N_0, with E_b denoting the average bit energy and N_0 denoting the one-sided noise power spectral density. We consider $\sigma_1^2=\ldots,=\sigma_{N_s}^2=1$ and then the average E_b/N_0 is identical for all streams $s\in\{1,\ldots,N_s\}$.

From Figs. 3 and 4 we can see the performance improves as the number of iterations increases as expected. Furthermore, the proposed hybrid equalizer is quite close to the digital counterpart for the 2-4th iterations. From these results, we verify that the gaps from the 1st to the 2nd iteration are much higher than from the 2nd iteration to the 4th.This larger gap

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is mainly due to the removal of the residual ISI which enables the added benefit of a larger diversity. From the 2nd to the 4th iteration there is also a benefit from ISI removal, but the gains are smaller since most of the ISI is removed in the 4nd iteration. At iteration 1, the BER target of 10^{-3} is achieved for an E_b / N_0 of 6 and 2dB, for sparse precoder and random precoder, respectively. However, at iteration 4, the BER target of 10^{-3} is achieved for an E_b / N_0 of -15.4 and -4.5dB, respectively. Therefore, the random precoder get a better performance for iteration 1, but the gain for sparse precoder was higher and the sparse precoder has a better performance than random precoder at iteration 4. This happens because the mmW massive MIMO are very correlated. The random precoder deals better with interference, but we get a higher beamforming gain with sparse precoder that despite dealing worse with interference, this one can be mitigated with an efficient non-linear equalizer.

VI. CONCLUSION

In this paper, we compared the performance of two hybrid transmit beamforming approaches, with different levels of CSI knowledge, combined with a hybrid iterative space-time equalizer for mmW massive MIMO systems. The analog part considers the specific hardware limitation inherent to the analog domain processing. A space-time encoder was used, before the analog precoders, to ensure transmit Gaussian based signals, which allowed to simplify the receiver optimization and to increase the system diversity.

The results have shown that the sparse precoder with a hybrid iterative space-time receiver achieved the best performance, mainly for low SNR regime, with a very few number of iterations. This happens because a larger beamforming gain can be achieved with sparse precoder and the interference can be efficiently removed with the iterative receiver structure. On the other hand, the random precoder explores the diversity and overcomes the problem of very correlated mmW massive MIMO channels. Therefore, the random precoder achieved the best performance for a linear equalizer (single iteration), because the signal at receiver suffers of less interference. Thus, we can conclude that the random precoder based transmitter structure is interesting for practical mmW massive MIMO based systems, where the channels are very correlated and it does not require CSI.

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