Leakage-Based Hybrid Transceiver Design for Millimeter Wave Multi-User Interference Channel

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***Abstract*—Millimeter wave (mmWave) communication is envisioned as a potential technology for next generation mobile communication as it can offer multi-gigabit-per-second data rate and a huge unlicensed spectrum. In this paper, we present a signal leakage-based low-complexity hybrid analog-digital transceiver design for a mmWave communication system operating in a K-user MIMO interference channel. Owing to higher cost and power consumption of components at mmWave frequencies, optimal system design with reduced hardware and computational complexity is critical to make the commercial use of mmWave communication viable. We achieve reduced hardware complexity of the proposed design by employing sparse approximation techniques for large antenna array. The proposed design is based on the maximization of signal-to-leakage-plus-noise ratio (SLNR) for all users. This leads to the reduced computational complexity of the overall system. Numerical results comparing the proposed design with existing iterative solution demonstrate that it can achieve comparable performance with reduced hardware and computational complexity. Performance of the proposed design with various dictionaries considered for beamforming has also been compared and illustrated in the simulation results.**

***Keywords*—mmWave communication, sparse signal processing, hybrid transceiver, orthogonal matching pursuit, 5G.**

I. Introduction

Recently, millimeter wave communication has been gaining considerable interest for next generation cellular communication in view of its potential to offer vast unlicensed bandwidth. With shorter wavelength at mmWave frequencies, large number of antenna elements can be packed in small volumes enabling high beamforming gain at transceivers, making MIMO an important component of the system architecture [1]–[3]. Unfortunately, providing each antenna element with a separate radio frequency (RF) chain will lead to high cost and power consumption of the mixed signal components at these frequencies, thus making conventional fully-digital systems practically infeasible [1]. In order to combat these challenges, the entire signal processing is divided into a cascade of digital baseband and analog beamformer hybrid structure. More recently, advances in hybrid beamforming strategies for mmWave systems has been discussed in [4], [5]. In [6], hybrid precoders are designed by formulating the overall design problem as a sparse signal recovery problem for a point-to-point mmWave MIMO system. Later in [7], single user hybrid mmWave system with large antenna array at both the base station and mobile station with partial channel knowledge is considered. Therein, the precoding scheme is developed considering a realistic spatial channel model and selecting dominant eigen vectors of the conditional channel covariance matrix. In [8], MMSE-based precoder and receive filters for multi-user MIMO interference channel are jointly designed using orthogonal matching pursuit (OMP) sparse approximation technique. However in [8], the transceiver matrices are obtained by an iterative approach, which leads to high computational cost of the overall mmWave system.

In this paper, we consider the design of a low-complexity hybrid K-user MIMO interference mmWave communication system using an OMP-based algorithm for obtaining hybrid RF/baseband MIMO processors. OMP-based hybrid processing is a sparse approximation technique using a greedy iterative approach, which reduces the residue by selecting the RF beamformers from a set of candidate vectors called dictionaries and optimizes the baseband processors in the least squares sense [9]. We first design the optimal precoder matrices by maximizing the signal-to-leakage-and-noise ratio (SLNR) for all users simultaneously. For a multi-user interference channel, leakage refers to the interference caused by the signal intended for a desired user on the remaining users. More specifically, leakage is the measure of the amount of signal power that leaks into other users. The benefit of solving SLNR-based optimization problem is that it results in K decoupled problems and can provide a closed form solution unlike in the case of MMSE-based optimization problem, which is solved using the iterative approach. This leads to reduced computational complexity as demonstrated later in this paper. SLNR-based system for MIMO downlink channel has been introduced for single stream data input in [10] and extended for MIMO downlink system employing Alamounti coding, enabling simultaneous transmission of two data streams in [11]. In [12], leakage based precoding for multi-user MIMO OFDM in downlink scenario has been discussed. An alternate SLNR-based linear precoding scheme has been discussed in [13], where improved performance was achieved in the case of multiple-stream data input for conventional MIMO downlink scenario. However, to the best of our knowledge, SLNR-based precoding scheme for mmWave system has been reported in the literature. Here, we propose a multi-stream SLNR-based hybrid mmWave system assuming the availability of perfect

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channel state information (CSI) at the transceivers . In the proposed design, we introduce the use of MMSE-type receive filter unlike the use of matched filter detector in the literature [10]–[13], hence improving the overall performance of the mmWave system. We evaluate the performance of the proposed design for various parameters and results are discussed later in the paper. We also compare the performance of our proposed SLNR-based system to that of MMSE-based iterative design in [8] in terms of both hardware and computational complexity. We demonstrate through simulation results that the proposed design results in comparable performance with reduced hardware and computation complexity making the mmWave system practically implementable for 5G.

The rest of the paper is as follows. Sec. II describes the system and channel model. The proposed SLNR-based low complexity system design is discussed in Sec. III. Computational analysis and simulation results are discussed in Sec. IV and V respectively. Finally, the conclusion is given in Sec. VI.

Notations: Throughout this paper, we use bold-faced lowercase letters to denote column vectors and bold-faced uppercase letters to denote matrices. $\underline{\mathbf{X}}$ and $\overline{\mathbf{X}}$ implies that the variable $\mathbf{X}$ corresponds to the baseband and RF block, respectively. $\mathbf{X}^*$,$\mathrm{tr}(\cdot)$, $\mathbb{E}\{\cdot\}$, $\|\cdot\|_0$ and $\|\cdot\|_F$ denotes the conjugate transpose, trace operator, expectation operator, 0-norm and Frobenius-norm respectively.

II. System and Channel Model

In this section, system design for hybrid mmWave communication system operating in a K-user MIMO interference channel is discussed. The block diagram of the proposed system is shown in Fig. 1. We consider multi-stream communication, where data streams $\mathbf{d}_k$ of size N_s for each user are transmitted over nTx transmit antennas and received over nRx receive antennas. Let, the number of RF chains associated with each user be $\overline{N}_t$ and $\overline{N}_r$ for processing at the transmitter and the receiver, respectively, such that $N_s \leq \overline{N}_t < nTx$ and $N_s \leq \overline{N}_r < nRx$. This reduced number of RF chains at the transceivers will lead to reduced hardware complexity . Let $\mathbf{F}_k^o$ be the optimal precoder matrix and $\mathbf{W}_k^o$ be the optimal receive filter at the k-th user for the conventional full-complexity system . Then in an hybrid system, the optimal filters are decomposed into their respective baseband and RF beamformer for each user such that $\mathbf{F}_k^o = \underline{\mathbf{F}}_k\overline{\mathbf{F}}_k$ and $\mathbf{W}_k^o = \underline{\mathbf{W}}_k^*\overline{\mathbf{W}}_k^*$. Let $\mathbf{x}_k$ be the signal transmitted by k-th user over the channel.Then the received signal $\mathbf{y}_k$ at user-k is given by,

$$\mathbf{y}_k = \mathbf{C}_{kk}\mathbf{F}_k\mathbf{d}_k + \sum_{j=1,\, j\neq k}^{K} \mathbf{C}_{kj}\mathbf{F}_j\mathbf{d}_j + \mathbf{n}_k, \tag{1}$$

where, matrix $\mathbf{C}_{kj}$ denotes the channel gain between jth transmitter and kth receiver and $\mathbf{n}_k \sim \mathcal{CN}(0, \sigma^2_{nRx})$ for $k = 1, 2, \cdots, K$, denotes the additive white noise. Thus, the estimated receive signal be,

$$\begin{aligned}\widehat{\mathbf{d}}_k = \mathbf{W}_k^*\mathbf{y}_k &= \underline{\mathbf{W}}_k^*\overline{\mathbf{W}}_k^*\mathbf{C}_{kk}\overline{\mathbf{F}}_k\underline{\mathbf{F}}_k\mathbf{d}_k \\ &+ \underline{\mathbf{W}}_k^*\overline{\mathbf{W}}_k^*\sum_{j\neq k}^{K}\mathbf{C}_{kj}\overline{\mathbf{F}}_j\underline{\mathbf{F}}_j\mathbf{d}_j + \underline{\mathbf{W}}_k^*\overline{\mathbf{W}}_k^*\mathbf{n}_k. \end{aligned}\tag{2}$$

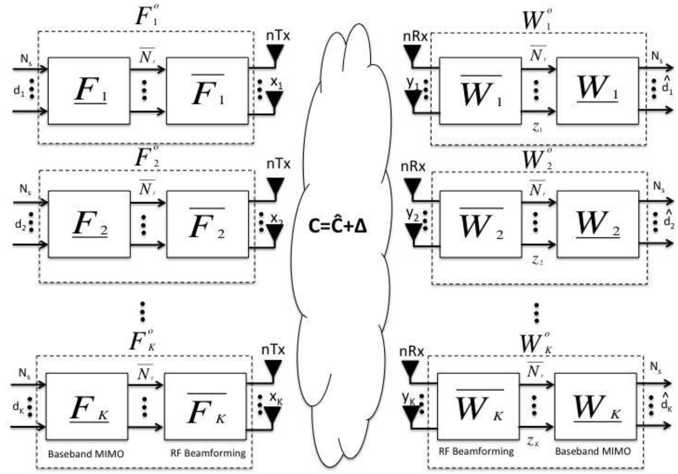

Fig. 1. Hybrid beamforming architecture for a K-user MIMO interference channel.

A. Channel Model

Due to high free-space path-loss and use of large number of tightly-packed antennas at mmWave frequencies, a 2D narrowband parametric clustered channel model is adopted in this paper. We consider the extended Saleh-Valenzuela model [3], in which the channel matrix $\mathbf{C}_{kj}$, from j^{th} transmitter to k^{th} receiver can be characterized as follow,

$$\mathbf{C}_{kj} = \gamma \sum_{m=1}^{N_{cl}} \sum_{n=1}^{N_{ray}} \alpha_{mn}^{kj}\mathbf{a}_r\big(\phi_{mn}^{r(k)}, \theta_{mn}^{r(k)}\big)\mathbf{a}_t\big(\phi_{mn}^{t(j)}, \theta_{mn}^{t(j)}\big), \tag{3}$$

where, N_{ray} is the number of rays in N_{cl} clusters and, the normalization factor $\gamma = \sqrt{\frac{nTxnRx}{N_{cl}N_{ray}}}$ is such that it satisfies $\mathrm{E}\big[\|\mathbf{C}_{\mathrm{kj}}\|_{\mathrm{F}}^2\big] = \mathrm{nTxnRx}$. α_{mn} denotes the complex gain of n^{th} ray in m^{th} cluster and is assumed to be i.i.d. and complex Gaussian random variable with zero mean and variance $\sigma_\alpha^2 \sim \mathcal{N}(0, \sigma_\alpha^2)$. $\mathbf{a}_t\big(\phi_{mn}^{t(j)}, \theta_{mn}^{t(j)}\big)$ and $\mathbf{a}_r\big(\phi_{mn}^{r(k)}, \theta_{mn}^{r(k)}\big)$ are the array response vectors at the transmitter and the receiver respectively, with $\phi_{mn}^{t(j)}, \theta_{mn}^{t(j)}$ and $\phi_{mn}^{r(j)}, \theta_{mn}^{r(j)}$ being azimuthal and elevation angle for transmit and receive antennas, such that,

$$\mathbf{a}\big(\phi_{mn}, \theta_{mn}\big) = \frac{1}{\sqrt{nTx}}\big[\exp^{\iota \mathrm{m}\times\frac{2\pi}{\lambda}\mathrm{d}(\sin(\phi))}\big]^T. \tag{4}$$

III. SLNR-Based Hybrid Transceiver Design for mmWave system

In this section, we present a leakage-based precoder and receive filter design for a K-user MIMO interference channel. We assume that the CSI is perfectly available at the transceivers. First, we will briefly review the formulation of the design problem and obtain the optimal precoder matrices and the receive filters. Then in Sec. III-B, we further decompose the optimal filters into hybrid RF/baseband processors to achieve reduced hardware complexity.

A. SLNR at a Glance

The part of the total signal power that is intended for any user k but that leaks into other users is termed as leakage. From Fig. 2, when user-k is transmitting the data vector $\mathbf{d}_k$ intended for receiver k , then part of this signal gets leaked

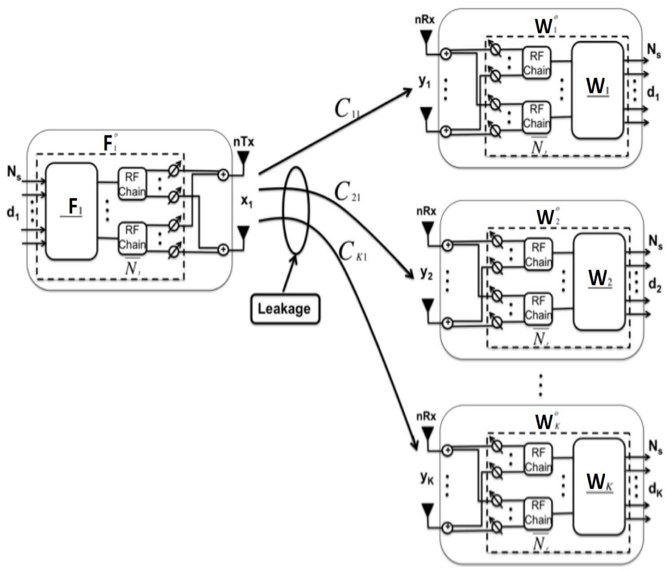

Fig. 2. Leakage from user 1 on other users in millimeter wave system.

towards other users. Thus, the total signal power leaked from user-k to all other users in the system is given by

$$l_k = \sum_{j=1\ ,\ j\neq k}^{K} ||\mathbf{C}_{kj}\mathbf{F}_k||^2. \tag{5}$$

Based on this, SLNR is defined as ratio of total intended signal power for any user k to the noise power plus the total leaked signal power towards other users $\tilde{k}$, where $\tilde{k} = \{1,..,k-1,k+1,..,K\}$. The SLNR for user-$k$ in multi-user interference channel scenario is given by

$$\mathrm{SLNR}_k = \frac{\mathrm{E}[\mathbf{d}_k^*\mathbf{F}_k^*\mathbf{C}_{kk}^*\mathbf{C}_{kk}\mathbf{F}_k\mathbf{d}_k]}{nRx\sigma_k^2 + \mathrm{E}[\sum_{j\neq k}^{K}\mathbf{d}_k^*\mathbf{F}_k^*\mathbf{C}_{jk}^*\mathbf{C}_{jk}\mathbf{F}_k\mathbf{d}_k]}. \tag{6}$$

Evaluating the expectation in (6) and assuming that $\mathrm{E}[\mathrm{d}_\mathrm{k}\mathrm{d}_\mathrm{j}^*] = 0$, for $\mathrm{k} \neq \mathrm{j}$, the expression for SLNR can be simplified as

$$\begin{aligned}\mathrm{SLNR}_k &= \frac{\mathrm{tr}(\mathbf{F}_k^*\mathbf{C}_{kk}^*\mathbf{C}_{kk}\mathbf{F}_k)}{nRx\sigma_k^2 + \mathrm{tr}(\mathbf{F}_k^*\tilde{\mathbf{C}}_{kk}^*\tilde{\mathbf{C}}_{kk}\mathbf{F}_k)}\\ &= \frac{\mathrm{tr}(\mathbf{F}_k^*\mathbf{C}_{kk}^*\mathbf{C}_{kk}\mathbf{F}_k)}{\mathrm{tr}[\mathbf{F}_k^*(\frac{nRx\sigma_k^2}{N_s}\mathbf{I} + \tilde{\mathbf{C}}_{kk}^*\tilde{\mathbf{C}}_{kk})\mathbf{F}_k]},\end{aligned} \tag{7}$$

where, $\tilde{\mathbf{C}}_{kk} = [\mathbf{C}_{1k}...\mathbf{C}_{(k-1)k}\mathbf{C}_{(k+1)k}...\mathbf{C}_{Kk}]^T$. As SLNR-based optimization problem leads to K decoupled problems, it leads to a closed form solution and make the overall processing computationally faster as compared to iterative solution in [8], [14].

Problem formulation: We design the optimal precoder matrices $\mathbf{F}^o$ for each user so as to maximize the SLNR under a constraint on the total transmit power. Thus the optimization problem for the k-th user can be stated as,

$$\mathbf{F}_k^o = \arg\max_{(\mathbf{F}_k)} \frac{\mathrm{tr}(\mathbf{F}_k^*\mathbf{C}_{kk}^*\mathbf{C}_{kk}\mathbf{F}_k)}{\mathrm{tr}[\mathbf{F}_k^*(\frac{nRx\sigma_k^2}{Ns}\mathbf{I} + \tilde{\mathbf{C}}_{kk}^*\tilde{\mathbf{C}}_{kk})\mathbf{F}_k]}, \tag{8}$$
$$\text{subject to: } \mathrm{tr}(\mathbf{F}_k^*\mathbf{F}_k) = N_s \text{ for } k \in \{1,2,...K\}$$

We solve the problem (8), which is of standard Rayleigh quotient form and obtain the precoder matrices by simultaneous diagonalization of two matrices as given in [13]. Thus, as $\mathbf{C}_{kk}^*\mathbf{C}_{kk}$ is Hermitian and $(\frac{nRx\sigma_k^2}{Ns}\mathbf{I} + \tilde{\mathbf{C}}_{kk}^*\tilde{\mathbf{C}}_{kk})$ is Hermitian and positive definite, then from the generalized eigen space,

TABLE I

SLNR based precoder design $\mathbf{F}_k$
Input: $\mathbf{A}_k = \mathbf{C}_{kk}^*\mathbf{C}_{kk}$ and $\mathbf{B}_k = (\mathbf{C}_{kk}^*\mathbf{C}_{kk} + \frac{nRx\sigma^2}{N_s}\mathbf{I} + \tilde{\mathbf{C}}_{kk}^*\tilde{\mathbf{C}}_{kk})$
1. Compute $\mathbf{G}_k$ by Cholesky decomposition of $\mathbf{B}_k$ such that $\mathbf{B}_k = \mathbf{G}_k\mathbf{G}_k^*$ where, $\mathbf{G}_k$ is a lower triangular matrix with positive diagonal elements. Then, we have $(\mathbf{G}_k^{-1})^* = \mathbf{Z}_k$.
2. Compute $\mathbf{A}_k^{'} = \mathbf{Z}_k^*\mathbf{A}_k\mathbf{Z}_k$
3. Perform eigen decomposition of $\mathbf{A}_k^{'}$ as $\mathbf{A}_k^{'}\mathbf{U}_k = \mathbf{U}_k\mathbf{\Lambda}_k$ to find the unitary matrix $\mathbf{U}_k$.
4. Compute $\mathbf{T}_k = \mathbf{Z}_k\mathbf{U}_k$.
Output: $\mathbf{F}_k = \gamma\mathbf{T}_k(\mathbf{I}_{N_s};\mathbf{0})$ so that $\mathrm{Tr}(\mathbf{F}_k\mathbf{F}_k^* = N_s)$

there exists a non-singular matrix $\mathbf{T}_k$ such that

$$\begin{aligned}\mathbf{T}_k^*\mathbf{C}_{kk}^*\mathbf{C}_{kk}\mathbf{T}_k &= \mathbf{\Lambda}_k\\ \mathbf{T}_k^*(\frac{nRx\sigma_k^2}{N_s}\mathbf{I} + \tilde{\mathbf{C}}_{\mathbf{kk}}^*\tilde{\mathbf{C}}_{\mathbf{kk}})\mathbf{T}_\mathbf{k} &= \mathbf{\Omega}_\mathbf{k},\end{aligned} \tag{9}$$

where, $\mathbf{T}_k$ spans the range space for generalized eigenspace of the pair $\left(\mathbf{C}_{kk}^*\mathbf{C}_{kk}, (\frac{nRx\sigma_k^2}{N_s}\mathbf{I} + \tilde{\mathbf{C}}_{kk}^*\tilde{\mathbf{C}}_{kk})\right)$. Let $\mathbf{B}_k = \left(\mathbf{C}_{kk}^*\mathbf{C}_{kk} + \frac{nRx\sigma_k^2}{N_s}\mathbf{I} + \tilde{\mathbf{C}}_{kk}^*\tilde{\mathbf{C}}_{kk}\right)$, then, from the Proposition 1 in [13], $\mathbf{T}_k$ can be obtained as $\mathbf{T}_k = \mathbf{Z}_k\mathbf{U}_k$, where $\mathbf{Z}_k \in \mathbb{C}^{nTx\times nTx}$ such that, $\mathbf{Z}_k^*\mathbf{B}_k\mathbf{Z}_k = \mathbf{I}$. Hence, we obtain the precoder matrices by using the algorithm given in Table I. Specifically, the precoder matrix for any user k is given by

$$\mathbf{F}_k^o = \gamma\mathbf{T}_k(\mathbf{I}_{N_s};\mathbf{0}), \text{ such that } \mathrm{Tr}(\mathbf{F}_k\mathbf{F}_k^*) = N_s. \tag{10}$$

In order to obtain the optimal receive filter matrices, a matched filter is used in [10], [13] which can be realized as

$$\mathbf{W}_k^{o*} = \frac{\mathbf{F}_k^{o*}\mathbf{C}_{kk}^*}{N_s||\mathbf{C}_{kk}\mathbf{F}_k^o||_F}. \tag{11}$$

In this paper, we use an MMSE-type receive filter for processing the received signal $\mathbf{y}_k$. Based on the available channel $\mathbf{C}_{kk}$ and the optimal precoder $\mathbf{F}_k^o$ obtained earlier, the sum-mean square error for multi-user interference channel can be obtained as,

$$\begin{aligned}&\mathrm{MSE}_k = \mathbb{E}\{\|\widehat{\mathbf{d}}_k - \mathbf{d}_k\|^2\}\\ &= \mathrm{tr}\big(\mathbf{W}_k\sum_{i=1}^{K}(\mathbf{C}_{ki}\mathbf{F}_i\mathbf{F}_i^*\mathbf{C}_{ki}^*)\mathbf{W}_k^* - (\mathbf{W}_k\mathbf{C}_{kk}\mathbf{F}_k + \mathbf{F}_k^*\mathbf{C}_{kk}^*\mathbf{W}_k^*)\\ &+\sigma_n^2\mathbf{W}_k\mathbf{W}_k^* + \mathbf{I}\big).\end{aligned} \tag{12}$$

We design the optimal receive filter by minimizing the sum-MSE under a the constraint on the total transmit power. Thus, the optimization problem can be mathematically expressed as,

$$\min_{\{\mathbf{W_k}\}}\sum_{k=1}^{K}\mathrm{MSE}_k \quad s.t. \quad \mathrm{tr}(\mathbf{F}_k^*\mathbf{F}_k) = N_s, \forall k \in \{1..K\}, \tag{13}$$

For a given $\mathbf{F}_k$, $k \in \{1,..,K\}$, we obtain the optimal MMSE receive filter as

$$\mathbf{W}_k^o = \mathbf{F}_k^*\mathbf{C}_{kk}^*\big[\sum_{i=1}^{K}\mathbf{C}_{ki}\mathbf{F}_i\mathbf{F}_i^*\mathbf{C}_{ki}^* + \sigma_n^2\mathbf{I}\big]^{-1}, k \in \{1,..,K\} \tag{14}$$

Once we obtain the optimal precoder and receive filter matrices, we introduce hybrid architecture leading to reduced hardware complexity by decomposing these optimal filters into digital baseband and analog RF beamformer matrices. We perform hybrid decomposition by using the OMP-based sparse approximation technique. The details of the hybrid design are discussed in the following subsection.

B. Hybrid OMP-Based Precoder/Receive Filter Design

We obtain the decomposed hybrid RF/baseband matrices by using orthogonal matching pursuit (OMP) sparse approximation technique. OMP is a greedy algorithm that constructs the sparse approximation through an iterative process. At each iteration, the residue is reduced, which starts out being equal to the full complexity optimal matrices, by appropriately selecting the column vectors from the predefined set of dictionaries. OMP is a well known signal processing algorithm, and has been extensively studied in literature for various applications [8], [9], [15]. To obtain the decomposed matrices for the given optimal matrices in (10) and (14), we select the RF beamforming vectors from a dictionary that is most strongly correlated to the residue computed at each iteration and obtain the baseband matrices by solving the least squares problem.

We first consider the decomposition of the optimal receive filters obtained in (14). Note first the optimal receive filter matrix can be expressed as,

$$\mathbf{W}_k^o = \mathbf{\Gamma}_{\mathbf{y}_k} \mathbf{\Gamma}_{\mathbf{y}_k \mathbf{s}_k}^{-1}, \tag{15}$$

where $\mathbf{\Gamma}_{\mathbf{y}_k} = \mathbb{E}[\mathbf{y}_k \mathbf{y}_k^*]$ and $\mathbf{\Gamma}_{\mathbf{y}_k \mathbf{d}_k} = \mathbb{E}[\mathbf{y}_k \mathbf{d}_k^*]$. The baseband receive filter matrix can be written as,

$$\underline{\mathbf{W}}_k^o = \mathbf{\Gamma}_{\mathbf{z}_k}^{-1} \mathbf{\Gamma}_{\mathbf{z}_k \mathbf{d}_k} = \left(\overline{\mathbf{W}}_k^* \mathbf{\Gamma}_{\mathbf{y}_k} \overline{\mathbf{W}}_k\right)^{-1} \overline{\mathbf{W}}_k^* \mathbf{\Gamma}_{\mathbf{y}_k \mathbf{d}_k}, \tag{16}$$

where $\mathbf{\Gamma}_{\mathbf{z}_k} = \mathbb{E}[\mathbf{z}_k \mathbf{y}_k^*]$ and $\mathbf{\Gamma}_{\mathbf{z}_k \mathbf{d}_k} = \mathbb{E}[\mathbf{z}_k \mathbf{d}_k^*]$. Using this baseband matrix, the problem for obtaining hybrid RF beamforming matrix can be formulated as,

$$\begin{aligned} \overline{\mathbf{W}}_k^o &= \underset{\overline{\mathbf{W}}_k}{\operatorname{argmin}} \, \mathbb{E}\|\mathbf{d}_k - \underline{\mathbf{W}}_k^{o*} \overline{\mathbf{W}}_k\|^2, \\ &= \underset{\overline{\mathbf{W}}_k^o}{\operatorname{argmin}} \, \|\mathbf{\Gamma}_{\mathbf{y}_k}^{\frac{1}{2}} \mathbf{W}_k^o - \mathbf{\Gamma}_{\mathbf{y}_k}^{\frac{1}{2}} \overline{\mathbf{W}}_k \underline{\mathbf{W}}_k^o\|_F^2. \end{aligned} \tag{17}$$

Introducing the dictionary $\mathbf{S}_{BF}$, the optimization problem for a sparse receive filter can be rephrased as,

$$\begin{aligned} \underline{\tilde{\mathbf{W}}}_k^o &= \underset{\underline{\tilde{\mathbf{W}}}_k}{\operatorname{argmin}} \, \|\mathbf{\Gamma}_{\mathbf{y}_k}^{\frac{1}{2}} \mathbf{W}_k^o - \mathbf{\Gamma}_{\mathbf{y}_k}^{\frac{1}{2}} \mathbf{S}_{BF} \underline{\tilde{\mathbf{W}}}_k^o\|_F^2, \\ & \textit{s.t. } \|\text{diag } (\underline{\tilde{\mathbf{W}}}_k^o \underline{\tilde{\mathbf{W}}}_k^{o*})\|_0 = \overline{N}_r. \end{aligned} \tag{18}$$

Similarly, the optimization problem for designing the sparse precoder matrix can be written as,

$$\overline{\mathbf{F}}_k^o = \underset{\overline{\mathbf{F}}_k}{\operatorname{argmin}} \, \|\mathbf{\Gamma}_{\mathbf{y}_k^r}^{\frac{1}{2}} \mathbf{F}_k^o - \mathbf{\Gamma}_{y_k^r}^{\frac{1}{2}} \overline{\mathbf{F}}_k \underline{\mathbf{F}}_k^o\|_F^2. \tag{19}$$

We have introduced the dictionary as above,

$$\begin{aligned} \overline{\mathbf{F}}_k^o &= \underset{\tilde{\mathbf{F}}_k}{\operatorname{argmin}} \, \|\mathbf{\Gamma}_{\mathbf{y}_k^r}^{\frac{1}{2}} \mathbf{F}_k^o - \mathbf{\Gamma}_{\mathbf{y}_k^r}^{\frac{1}{2}} \mathbf{S}_{BF} \tilde{\mathbf{F}}_k^o\|_F^2, \\ \text{s.t.} \quad & \|\text{diag } (\underline{\tilde{\mathbf{F}}}_k^o \underline{\tilde{\mathbf{F}}}_k^{o*})\|_0 = \overline{N}_t \text{ and } \|\underline{\tilde{\mathbf{F}}}_k\|_F^2 = \|\mathbf{F}_k^o\|_F^2. \end{aligned} \tag{20}$$

To solve this optimization problem, the RF beamforming matrix is selected from the set of candidate vectors $\mathbf{S}_{BF}$. At each iteration, the effective residue is updated, hence, minimizing the overall error with each iteration. A generalized detailed OMP based iterative algorithm for jointly designing the RF and baseband hybrid filters at transceivers is given in Table II. The set of candidate beamforming vectors are acquired from different dictionary sets. In this paper, we evaluate the performance of the proposed designs over different dictionaries including, eigen beamforming, discrete Fourier transform(DFT), discrete Cosine transform(DCT), discrete Hadamard transform(DHT) and antenna selection beamforming and is illustrated through simulation results. In general, different dictionaries have different properties and can be selected depending on the application. Eigen beamforming is most complex of all above mentioned dictionaries and consists of eigenvectors of $\mathbf{\Gamma}_{\mathbf{y}_k}$. However, it offers the perfect decomposition of optimal full-complexity matrices into hybrid baseband and RF units which is proved in lemma-2 in [8], and hence offers very good performance, whereas, DFT, DCT and DHT dictionaries can be designed by using the columns of Fourier transform, Cosine transform and Hadamard transform matrices respectively. In the case of implementation complexity, antenna selection beamforming method is considered to be simplest of all, as it can be implemented by using a switching circuit, where the dictionary consisting of columns of I_{nRx} are considered.

TABLE II

OMP-based iterative algorithm
Require $\mathbf{P}_k^o, \mathbf{\Phi}, \mathbf{S}_{BF}$
$\overline{\mathbf{Q}}_k = [\ ]$
$\mathbf{R}_0 = \mathbf{P}_k^o$
for $i = 1$ to $\overline{N}$ do
$\quad \mathbf{\Psi}_{i-1} = (\mathbf{\Phi}\mathbf{S}_{BF})^*(\mathbf{\Phi}\mathbf{R}_{i-1})$
$\quad l = arg\ max_{m=1 \ldots M}(\mathbf{\Psi}_{i-1}\mathbf{\Psi}_{i-1}^*)_{m,m}$
$\quad \overline{\mathbf{Q}}_k = [\overline{\mathbf{Q}}_k \| \mathbf{\Phi}(:,k)]$
$\quad \underline{\mathbf{Q}}_k = (\overline{\mathbf{Q}}_k^* \overline{\mathbf{Q}}_k)^{-1} \overline{\mathbf{Q}}_k^* \mathbf{P}_k^o$
$\quad \mathbf{R}_i = \frac{\mathbf{P}_k^o - \overline{\mathbf{Q}}_k \underline{\mathbf{Q}}_k}{\|\mathbf{P}_k^o - \overline{\mathbf{Q}}_k \underline{\mathbf{Q}}_k\|_F}$
end for
$\underline{\mathbf{Q}} = \sqrt{N_s} \frac{\underline{\mathbf{Q}}}{\|\overline{\mathbf{Q}}\underline{\mathbf{Q}}^*\|_F}$, when $\zeta = 1$
return $\overline{\mathbf{Q}}, \underline{\mathbf{Q}}$
Precoder: $\overline{N} = \overline{N}_t, \mathbf{P}^o = \mathbf{F}^o, \mathbf{\Phi} = \mathbf{\Gamma}_{\mathbf{y}_k^r}^{\frac{1}{2}}, \zeta = 1,$ $\overline{\mathbf{Q}} = \overline{\mathbf{F}}$, and $\underline{\mathbf{Q}} = \underline{\mathbf{F}}$
Receive filter: $\overline{N} = \overline{N}_r, \mathbf{P}^o = \mathbf{W}^{o*}, \mathbf{\Phi} = \mathbf{\Gamma}_{\mathbf{y}_k}^{\frac{1}{2}}, \zeta = 0,$, $\overline{\mathbf{Q}} = \overline{\mathbf{W}}$, and $\underline{\mathbf{Q}} = \underline{\mathbf{W}}$

IV. Computational Complexity

In this section, we analyze the computational complexity of both the MMSE-based iterative and proposed SLNR based optimization problems. We compare the complexity for each user based on two factors. First, we consider the count of basic multiplication/division and addition/subtraction for different processing parts in the algorithms, the details of which are given in Table III. Then, we provide the asymptotic computational complexity in big-O notations as given in Table IV. It is observed that iterative solution is more complex then SLNR based system due to the coupled optimization. While, SLNR based precoder design decomposes the complete problem into decoupled optimization problems and reduces the complexity by almost a factor of N. From the results, it is observed that precoder design using iterative method depends on the number of users present in the system. Thus, as the number of users increases, the slower the iterative method becomes. However, SLNR-based precoder performs equivalent computations even with increasing number of users. From Table IV, the overall complexity of MMSE-based iterative

TABLE III

Parameter	Multiplication/Division		Addition/Subtraction	
	Iterative	SLNR	Iterative	SLNR
Precoder	$(8K+4)N^2Ns+N^2(4NK+4N+10)$	$12N^3+9N^2+1$	$3N^2Ns(2K+1)+N^2(3KN+3N+8)$	$12N^3+8N^2$
Receiver	$(8K+4)N^2Ns+N^2(4NK+4N+10)$	$2N+3N^3K+3N^3+1$	$3N^2Ns(2K+1)+N^2(3KN+3N+8)$	$2N+N^3-2$
Langrangian	$8KN^2Ns+4KN^3+3N+10$	—	$6KN^2Ns+3KN^3+3N+2$	—
Total	$8N^2Ns(3K+1)+4N^3(3K+2)+20N^2+3N+10$	$3N^3(5+K)+9N^2+2N+2$	$6N^2Ns(3K+1)+3N^3(3K+2)+16N^2+3N+2$	$13N^3+8N^2+2N-2$

TABLE IV

Parameter	Iterative	SLNR
Precoder	$\mathcal{O}(KN^3)$	$\mathcal{O}(N^3)$
Receiver	$\mathcal{O}(KN^3)$	$\mathcal{O}(N^3)$
Langrangian	$\mathcal{O}(KN^3)$	—
Total	$\mathcal{O}(N_{iter}KN^3)$	$\mathcal{O}(N^3)$

solution is of the order of O($N_{iter}KN^3$), which increases by order of O(N) when number of users increases in the system, whereas, the complexity of leakage-based system is of the order of O(N^3). Hence, the proposed system results in reduced computational complexity for mmWave system.

V. Simulation Results

In this section, we present the simulation results for the proposed leakage-based mmWave system in terms of sum-mean square error (Sum-MSE), bit-error-rate (BER) and sum-rate performance. The sum-rate for multi-user MIMO system is evaluated as, $sumrate = \sum_{k=1}^{K}\sum_{i=1}^{Ns}\log_2(1+SINR_{k,i})$ where $SINR_{k,i} = \frac{\mathbf{W}_{ki}\mathbf{C}_{kk}\mathbf{F}_{ki}\mathbf{F}_{ki}^*\mathbf{C}_{kk}^*\mathbf{W}_{ki}^*}{\mathbf{W}_{ki}\mathbf{\Psi}_{ki}\mathbf{W}_{ki}^*}$ and $\mathbf{\Psi}_{ki} = \sigma_n^2\mathbf{I} + \sum_{l=1}^{k}\mathbf{C}_{kl}\mathbf{F}_l\mathbf{F}_l^*\mathbf{C}_{kl}^* - \mathbf{C}_{kk}\mathbf{F}_{ki}\mathbf{F}_{ki}^*\mathbf{C}_{kk}^*$. We compare the performance amongst following schemes,

1) Proposed SLNR-based mmWave system with MMSE type receive filter (*ref.* SLNR-MMSE).
2) Proposed SLNR-based mmWave system with matched filter type receive filter (*ref.* SLNR-MF).
3) Original SLNR-based system in [11] was extended for mmWave system and compared with proposed design (*ref.* SLNR-ORIG).
4) MMSE-based joint iterative transceiver design for mmWave system in [8] (*ref.* ITR-MMSE).

Throughout our simulations, we assume $\{K \in 2,4,8\}$ users in a multi-user interference system equipped with $nTx = nRx = 2\overline{N}_t$ number of transmit and receive antennas. The RF chain associated with each user at both the transmitter and receiver unit is $\overline{N}_t = \overline{N}_r = 2N_s + 2$, hence achieving atleast half of the hardware complexity as compared to conventional fully-digital system. The channel assumed is with uniform linear antenna arrays with inter-element spacing as $\lambda/2$. The number of clusters and number of rays are assumed to be $N_{cl} \in \{1,2,4\}$ and $N_{ray} = 5$ respectively. BPSK modulation scheme is assumed for generation of data. All the simulations parameters have been tested for $N = 10000$ data samples.

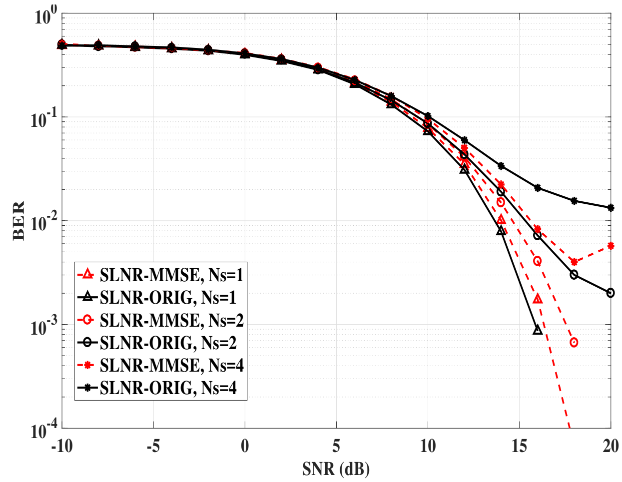

Fig. 3. BER performance versus varying SNR for proposed SLNR-based and original SLNR-based system with varying number of data streams at $K = 4$.

In Fig. 3 and Fig. 4, we demonstrate the BER performance for various schemes (SLNR-MMSE, SLNR-MF and SLNR-ORIG) by considering different number of input data streams. It is observed that BER decreases with increasing SNR but system performance also degrades at larger number of data streams due to increased channel interference. Through both the figures, we also observed that the proposed SLNR-MMSE scheme performs best amongst all the other schemes. In Fig. 5, the sum-MSE performance has been compared for SLNR-MMSE to ITR-MMSE over different number of users $K \in \{2,4,8\}$. From Fig. 5, we observed that the proposed SLNR-based system performs equivalently to the iterative solution at higher SNR and results in overall comparable performance with reduced computational complexity by the factor O(N), as discussed previously in Section. IV. In Fig. 6, we demonstrate the performance of SLNR-MMSE in terms of sum-rate, for different dictionaries that includes eigen beamforming, DCT, DFT, DHT and antenna selection, considered while designing the RF beamformers during the hybrid system design. Sum-rate is observed to increase with increasing SNR for all the aforementioned dictionaries, with eigen beamformer performing best amongst all. Other dictionaries show considerable performance for the proposed system, but due to the ease of their implementation they can be considered for the applications having non-critical criterion.

VI. Conclusion

In this paper, we proposed hybrid analog/digital transceiver designs for a K-user MIMO interference channel for mmWave

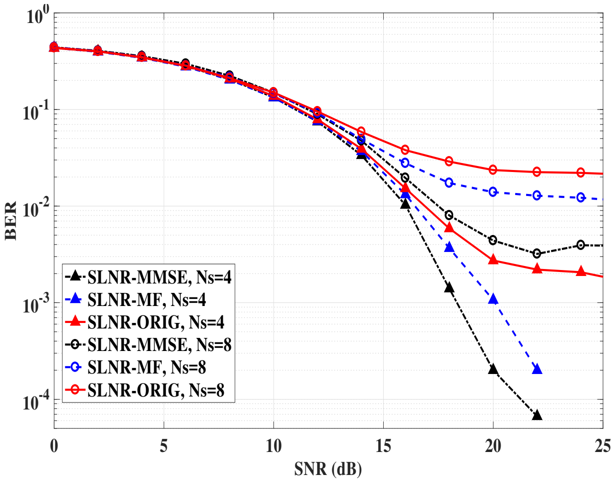

Fig. 4. BER performance versus varying SNR for proposed system with MMSE-type and MF type receiver and original SLNR-based system for different number of data streams at $K = 4$.

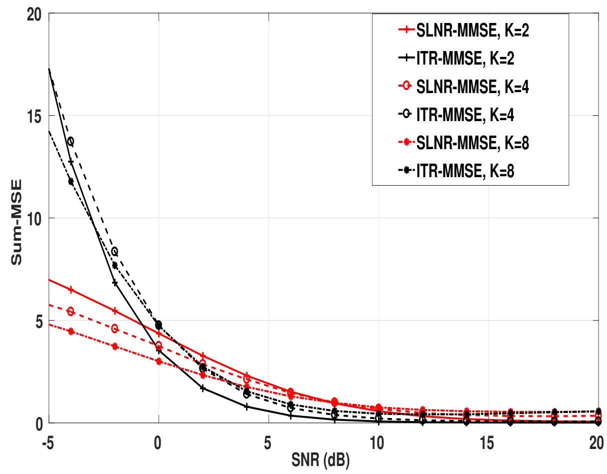

Fig. 5. Sum-MSE performance versus varying SNR for proposed SLNR-based system and MMSE-based iterative system for different number of users at $Ns = 4$.

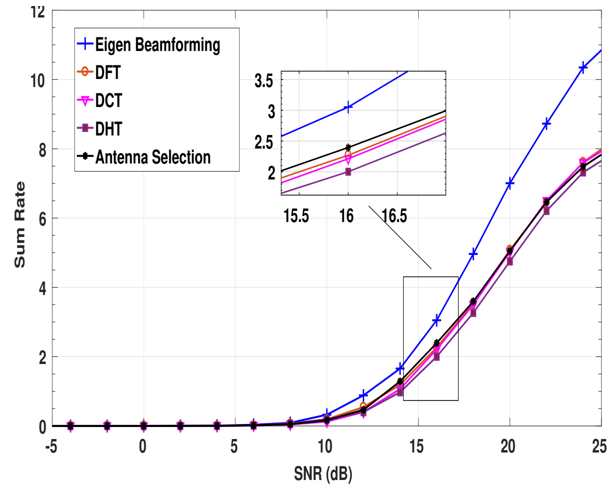

Fig. 6. Sumrate performance versus varying SNR for proposed system using different beamforming techniques at $K = 4$ and $Ns = 4$.

communication system that is less expensive in terms of both hardware and computational cost. Low hardware complexity is achieved by using hybrid system architecture, where the analog RF and digital baseband units were designed by the use of OMP-based sparse approximation technique. SLNR-based design lead to reduction of computational cost as it resulted in decoupled problems and provided a closed form solution. It is observed through simulations that the use of MMSE type receiver outperforms the use of matched filter type detector in the system. We also compared the performance of proposed system with that of joint transceiver design for mmWave system in literature [8]. Simulations results over various performance parameters demonstrate that the proposed design achieves comparable performance with reduced hardware and computational complexity.

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