Feedback Bit Allocation Schemes for Multi-User Distributed Antenna Systems

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***Abstract*—In this paper, we propose a feedback bit allocation algorithm for multi-user downlink distributed antenna systems with limited feedback. We consider a composite fading channel with small scale fadings and path loss, and assume the case where each user is served by only one distributed antenna (DA) port while each DA port can support any number of users. In order to efficiently determine bit allocation, we propose an iterative algorithm which minimizes an upper bound of a mean rate loss. Compared to conventional bit allocation methods, the proposed algorithm can be applied to more general system configurations. Simulation results show that our proposed algorithm offers a performance gain of 20% over an equal bit allocation scheme.**

***Index Terms*—Distributed antenna systems, limited feedback, feedback bit allocation, SLNR maximizing beamforming.**

I. INTRODUCTION

RECENTLY, distributed antenna systems (DAS) have been introduced as a new cellular communication structure. Unlike conventional centralized antenna systems (CAS) where all antennas are co-located at the cell center, distributed antenna (DA) ports in the DAS are separated geographically. Thus, the DAS can reduce the access distance for each user along with the transmit power and co-channel interference, which results in improved cell-edge performance [1]. By considering a composite fading channel, performance analysis and transmission schemes for the DAS were studied in [2] and [3].

In practical limited feedback systems, a transmitter acquires channel state information (CSI) through feedback from receivers based on a codebook whose size is determined by the allocated number of feedback bits. The authors in [4] and [5] presented feedback bit allocation methods which minimize an upper bound of a mean rate loss due to quantization errors for the multi-cell CAS. In [4], the bit allocation algorithm was proposed under the assumption that the zero-forcing beamforming is employed, and thus the number of transmit antennas is set to be greater than or equal to the number of users. The work in [5] considered a Wyner channel model and adopted generalized eigenvector beamforming which is similar to the signal-to-leakage plus noise ratio (SLNR) maximization approach [6]. In [5], it is assumed that each cell has one user interfered by only one adjacent cell due to the Wyner channel model. Also, the application of the bit allocation method in [5] is limited to a system where each transmitter has two antennas. In [7], by applying the SLNR maximizing beamforming, a feedback bit allocation algorithm which maximizes a lower bound of the expected signal-to-interference plus noise ratio (SINR) was introduced for the multi-user single-cell DAS assuming that each DA port supports only one user and each user is also served by only one DA port.

This paper extends our prior work in [7] to a more general case where each DA port can support any number of users. We assume that each user selects the nearest DA port, and the chosen DA ports transmit the signal to the corresponding users. This pairing has some advantages in terms of the desired signal power compared to the conventional pairing considered in [7], since users may be supported by the second closest or the third closest DA port in [7]. Under these assumptions, a DA port can support any number of users, and thus there exist both inter- and intra- DA port interferences. In this case, the bit allocation problem is more complex than that of [7] which considers only inter-DA port interferences.

To determine bit allocation, we first examine an upper bound of the mean rate loss using the results in [5] and [8]. Then, we propose an iterative algorithm which specifies the number of feedback bits. Note that in the case where the number of DA ports which are turned on is smaller than three, our algorithm does not need an iterative procedure and has a closed-form solution. Also, unlike the methods in [4] and [5], there is no limitation on system configurations in our algorithm. In addition, our proposed solution includes the results in [7] as a special case. Simulation results demonstrate that the DAS with the proposed feedback bit allocation algorithm outperforms the system with an equal bit allocation scheme and our pairing provides a sum rate gain over the conventional pairing considered in [7].

Throughout this paper, bold lower case letters denote vectors, and the superscripts $(\cdot)^H$ and $(\cdot)^{-1}$ stand for Hermitian and the inverse operation, respectively. Also, $\mathcal{E}(\cdot)$ represents expectation.

II. SYSTEM MODEL

In this section, we describe a system model for the multi-user downlink DAS with limited feedback in single-cell environments. We consider the DAS which has N DA ports with M antennas and K users with a single antenna. The locations of DA ports are determined by the results in [9] and we employ the composite channel model which encompasses not only small scale fading (i.e. Rayleigh fading) but also large scale fading (i.e. path loss). In this paper, we apply per-DA port power constraint P and set the cell radius as R. Also, it is assumed that each user and DA port know distances between all linked channels. For the pairing between DA ports and users, we generalize the case of [7]. Each user selects the nearest DA port, and the chosen DA ports transmit the signal to the corresponding users. Also, the DA ports which are not selected by any user are turned off. Here, we denote $\mathcal{D}$ as the

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set of DA ports which are turned on and $\mathcal{U}_n$ as the set of users which are served by the n-th DA port.

In this paper, we consider that each user has perfect knowledge of linked channel state and quantizes the channel direction information (CDI) of the small scale fading based on a codebook to feed the CDI back to the DA ports. We assume that the codebook vectors are independently determined by utilizing RVQ [10]. Let us denote the set of codewords for the k-th user and the n-th DA port as $\mathcal{C}_{n,k} = \left\{\mathbf{c}_{n,k,1}, \mathbf{c}_{n,k,2}, \cdots, \mathbf{c}_{n,k,2^{B_{n,k}}}\right\}$ where each element is a unit norm column vector of length M and $B_{n,k}$ indicates the allocated number of feedback bits for the channel between the k-th user and the n-th DA port.

Also, we denote the channel column vector of length M between the n-th DA port and the k-th user as $\mathbf{g}_{n,k} = d_{n,k}^{-\alpha/2}\mathbf{h}_{n,k}$ where $d_{n,k}$ stands for the distance between the n-th DA port and the k-th user, α indicates the path loss exponent and $\mathbf{h}_{n,k}$ equals the channel column vector for small scale fading. The elements of $\mathbf{h}_{n,k}$ are independent and identically distributed complex Gaussian random variables with zero mean and unit variance. Here, the CDI of $\mathbf{h}_{n,k}$ is defined as $\tilde{\mathbf{h}}_{n,k} \triangleq \mathbf{h}_{n,k}/\left\|\mathbf{h}_{n,k}\right\|$. Then, the k-th user quantizes $\tilde{\mathbf{h}}_{n,k}$ as $\hat{\mathbf{h}}_{n,k} = \arg\max_{\mathbf{c}\in\mathcal{C}_{n,k}} \left|\tilde{\mathbf{h}}_{n,k}^H\mathbf{c}\right|^2$ where $\hat{\mathbf{h}}_{n,k}$ indicates the quantized channel between the k-th user and the n-th DA port. Each DA port acquires the quantized CDI of the linked channels via feedback of the codeword index from each user, and then computes the beamforming vector by utilizing the quantized CDI.

We assume equal power allocation for users supported by the same DA port and define $P_n \triangleq \frac{P}{|\mathcal{U}_n|}$ where $|\mathcal{U}_n|$ represents the cardinality of $\mathcal{U}_n$, i.e., the number of users served by the n-th DA port. Then, the received signal for the k-th user is written as $y_k = \sum_{n\in\mathcal{D}}\sum_{i\in\mathcal{U}_n}\sqrt{P_n}\mathbf{g}_{n,k}^H\hat{\mathbf{w}}_{n,i}s_{n,i} + z_k$ where $\hat{\mathbf{w}}_{n,i}$ stands for the i-th user beamforming column vector of length M with unit norm ($\|\hat{\mathbf{w}}_{n,i}\| = 1$) which is computed by utilizing the quantized channel information at the n-th DA port, $s_{n,i}$ denotes the desired signal of the i-th user transmitted from the n-th DA port with $\mathcal{E}\left[|s_{n,i}|^2\right] = 1$ and z_k indicates the additive complex Gaussian noise variable with zero mean and variance σ_z^2.

Assuming that the j-th DA port is selected by the k-th user, the received signal for the k-th user can be rewritten as $y_k = \sqrt{P_j}\mathbf{g}_{j,k}^H\hat{\mathbf{w}}_{j,k}s_{j,k} + \sum_{\substack{i\in\mathcal{U}_j\\ i\neq k}}\sqrt{P_j}\mathbf{g}_{j,k}^H\hat{\mathbf{w}}_{j,i}s_{j,i} + \sum_{\substack{n\in\mathcal{D}\\ n\neq j}}\sum_{i\in\mathcal{U}_n}\sqrt{P_n}\mathbf{g}_{n,k}^H\hat{\mathbf{w}}_{n,i}s_{n,i} + z_k$. The first term indicates the desired signal and the second and the third terms account for the intra- and inter-DA port interferences, respectively.

In this paper, we employ the SLNR maximizing beamforming [6] to simplify the analysis. Then, the k-th user beamforming vector is given as

$$\hat{\mathbf{w}}_{j,k} = \max\mathrm{ev}\left(\left(\sigma_z^2\mathbf{I} + P_j\sum_{i=1,i\neq k}^{K} d_{j,i}^{-\alpha}\hat{\mathbf{h}}_{j,i}\hat{\mathbf{h}}_{j,i}^H\right)^{-1} P_j d_{j,k}^{-\alpha}\hat{\mathbf{h}}_{j,k}\hat{\mathbf{h}}_{j,k}^H\right) \quad (1)$$

where $\max\ \mathrm{ev}(\mathbf{A})$ denotes the eigenvector corresponding to the largest eigenvalue of $\mathbf{A}$. Albeit suboptimum in terms of the sum rate, the SLNR maximizing beamforming offers good performance in a distributed manner and is commonly used in multi-cell systems due to its simple computation.

Let us define $\mathrm{SINR}_k^{\mathrm{LFB}}$ as SINR with limited feedback for the k-th user. Then, the average sum rate with limited feedback can be represented as $C^{\mathrm{LFB}} = \mathcal{E}\left[\sum_{k=1}^{K}\log_2\left(1+\mathrm{SINR}_k^{\mathrm{LFB}}\right)\right]$ where

$$\mathrm{SINR}_k^{\mathrm{LFB}} \triangleq \frac{P_j\left|\mathbf{g}_{j,k}^H\hat{\mathbf{w}}_{j,k}\right|^2}{\sigma_z^2 + \sum_{\substack{i\in\mathcal{U}_j\\ i\neq k}}P_j\left|\mathbf{g}_{j,k}^H\hat{\mathbf{w}}_{j,i}\right|^2 + \sum_{\substack{n\in\mathcal{D}\\ n\neq j}}\sum_{i\in\mathcal{U}_n}P_n\left|\mathbf{g}_{n,k}^H\hat{\mathbf{w}}_{n,i}\right|^2}. \quad (2)$$

Next, we will formulate an objective function based on an upper bound of the mean rate loss for feedback bit allocation and propose an iterative algorithm which determines the number of feedback bits between all links.

III. Feedback Bit Allocation

We denote $B_k^t = \sum_{n\in\mathcal{D}}B_{n,k}$ as the total number of feedback bits for the k-th user. In practical systems, finite B_k^t leads to quantization errors. To enhance performance, we should properly allocate feedback bits to all links by utilizing a reliable metric. In this paper, as the optimization metric, we utilize an upper bound of the mean rate loss.

A. Problem Formulation

We define $\mathbf{w}_{j,k}$ as the k-th user beamforming vector with full CSI by substituting $\mathbf{h}$ instead of $\hat{\mathbf{h}}$ in (1). Also, the average sum rate with full CSI is represented as $C^{\mathrm{full}} = \mathcal{E}\left[\sum_{k=1}^{K}\log_2\left(1+\mathrm{SINR}_k^{\mathrm{full}}\right)\right]$ where $\mathrm{SINR}_k^{\mathrm{full}}$ is obtained by replacing $\hat{\mathbf{w}}_{j,k}$ with $\mathbf{w}_{j,k}$ in (2). Then, the mean rate loss is defined as

$$\Delta C = C^{\mathrm{full}} - C^{\mathrm{LFB}} = \sum_{k=1}^{K}\Delta C_k \quad (3)$$

where $\Delta C_k = \mathcal{E}\left[\log_2\left(1+\mathrm{SINR}_k^{\mathrm{full}}\right) - \log_2\left(1+\mathrm{SINR}_k^{\mathrm{LFB}}\right)\right]$. By employing a high SINR assumption and dropping the denominator of $\mathrm{SINR}_k^{\mathrm{full}}$, ΔC_k can be bounded by $\Delta C_k < \mathcal{E}\left[\log_2\left(P_j\left|\mathbf{g}_{j,k}^H\mathbf{w}_{j,k}\right|^2\right) - \log_2\left(\mathrm{SINR}_k^{\mathrm{LFB}}\right)\right]$. Also, a lower bound of the numerator of $\mathrm{SINR}_k^{\mathrm{LFB}}$ in (2) is given as $P_j d_{j,k}^{-\alpha}\left|\mathbf{h}_{j,k}^H\hat{\mathbf{w}}_{j,k}\right|^2 \geq P_j d_{j,k}^{-\alpha}\left\|\mathbf{h}_{j,k}\right\|^2\left|\tilde{\mathbf{h}}_{j,k}^H\hat{\mathbf{h}}_{j,k}\right|^2\left|\hat{\mathbf{h}}_{j,k}^H\hat{\mathbf{w}}_{j,k}\right|^2$ [5].

Then, an upper bound of ΔC_k is expressed as (4) on the top of the next page. Equality (5) is derived from the fact $\mathcal{E}\left[\log_2\left|\tilde{\mathbf{h}}_{j,k}^H\mathbf{w}_{j,k}\right|^2\right] = \mathcal{E}\left[\log_2\left|\hat{\mathbf{h}}_{j,k}^H\hat{\mathbf{w}}_{j,k}\right|^2\right]$ and the approximation (6) is obtained from applying Jensen's inequality to both the numerator and the denominator [11].

Through simulations, we have verified that (6) is greater than (5) and the gap between (5) and (6) decreases as B_k^t increases. Also, since the denominator becomes dominant at the high SNR region, (5) is upper-bounded by (6). Thus, we can say that (6) is an upper bound of ΔC_k and the approximation does not affect our results. Utilizing the probability density function of $\left|\tilde{\mathbf{h}}_{j,k}^H\hat{\mathbf{h}}_{j,k}\right|^2$, a lower bound of $\mathcal{E}\left[\left|\tilde{\mathbf{h}}_{j,k}^H\hat{\mathbf{h}}_{j,k}\right|^2\right]$ is given as [8]

$$\mathcal{E}\left[\left|\tilde{\mathbf{h}}_{j,k}^H\hat{\mathbf{h}}_{j,k}\right|^2\right] > 1 - 2^{-\frac{B_{j,k}}{M-1}}. \quad (7)$$

Now, we will derive a bound of $\mathcal{E}\left[\left|\mathbf{h}_{j,k}^H\hat{\mathbf{w}}_{j,i}\right|^2\right]$ and $\mathcal{E}\left[\left|\mathbf{h}_{n,k}^H\hat{\mathbf{w}}_{n,i}\right|^2\right]$ ($i \neq k$). Since $\mathcal{E}\left[\left|\mathbf{h}_{j,k}^H\hat{\mathbf{w}}_{j,i}\right|^2\right]$ and $\mathcal{E}\left[\left|\mathbf{h}_{n,k}^H\hat{\mathbf{w}}_{n,i}\right|^2\right]$ have the same form, we only derive the

$$\Delta C_k < \mathcal{E}\Big[\log_2\big|\tilde{\mathbf{h}}_{j,k}^H \mathbf{w}_{j,k}\big|^2\Big] - \mathcal{E}\Big[\log_2\big|\hat{\mathbf{h}}_{j,k}^H \hat{\mathbf{w}}_{j,k}\big|^2\Big] - \mathcal{E}\Bigg[\log_2 \frac{\big|\tilde{\mathbf{h}}_{j,k}^H \hat{\mathbf{h}}_{j,k}\big|^2}{\sigma_z^2 + \sum_{i\in\mathcal{U}_j, i\neq k} P_j \big|\mathbf{g}_{j,k}^H \hat{\mathbf{w}}_{j,i}\big|^2 + \sum_{n\in\mathcal{D}, n\neq j}\sum_{i\in\mathcal{U}_n} P_n \big|\mathbf{g}_{n,k}^H \hat{\mathbf{w}}_{n,i}\big|^2}\Bigg] \quad (4)$$

$$= -\mathcal{E}\Bigg[\log_2\big|\tilde{\mathbf{h}}_{j,k}^H \hat{\mathbf{h}}_{j,k}\big|^2 \Big/ \Big(\sigma_z^2 + \sum_{i\in\mathcal{U}_j, i\neq k} P_j \big|\mathbf{g}_{j,k}^H \hat{\mathbf{w}}_{j,i}\big|^2 + \sum_{n\in\mathcal{D}, n\neq j}\sum_{i\in\mathcal{U}_n} P_n \big|\mathbf{g}_{n,k}^H \hat{\mathbf{w}}_{n,i}\big|^2\Big)\Bigg] \quad (5)$$

$$\approx -\log_2 \mathcal{E}\Big[\big|\tilde{\mathbf{h}}_{j,k}^H \hat{\mathbf{h}}_{j,k}\big|^2\Big] \Big/ \Big(\sigma_z^2 + \sum_{i\in\mathcal{U}_j, i\neq k} P_j d_{j,k}^{-\alpha} \mathcal{E}\Big[\big|\mathbf{h}_{j,k}^H \hat{\mathbf{w}}_{j,i}\big|^2\Big] + \sum_{n\in\mathcal{D}, n\neq j}\sum_{i\in\mathcal{U}_n} P_n d_{n,k}^{-\alpha} \mathcal{E}\Big[\big|\mathbf{h}_{n,k}^H \hat{\mathbf{w}}_{n,i}\big|^2\Big]\Big) \quad (6)$$

bound of $\mathcal{E}\big[\big|\mathbf{h}_{n,k}^H \hat{\mathbf{w}}_{n,i}\big|^2\big]$. By applying the triangle inequality, $\mathcal{E}\big[\big|\mathbf{h}_{n,k}^H \hat{\mathbf{w}}_{n,i}\big|^2\big]$ can be bounded by $\mathcal{E}\big[\big|\mathbf{h}_{n,k}^H \hat{\mathbf{w}}_{n,i}\big|^2\big] \leq M\mathcal{E}\big[1 - \big|\tilde{\mathbf{h}}_{n,k}^H \hat{\mathbf{h}}_{n,k}\big|^2\big]$ [5]. Similar to the case of $\big|\tilde{\mathbf{h}}_{j,k}^H \hat{\mathbf{h}}_{j,k}\big|^2$, an upper bound of $\mathcal{E}\big[\big|\mathbf{h}_{n,k}^H \hat{\mathbf{w}}_{n,i}\big|^2\big]$ is obtained as [8]

$$\mathcal{E}\Big[\big|\mathbf{h}_{n,k}^H \hat{\mathbf{w}}_{n,i}\big|^2\Big] < M \cdot 2^{-\frac{B_{n,k}}{M-1}}. \quad (8)$$

Substituting (7) and (8) into (6), an upper bound of ΔC_k is computed as

$$\Delta C_k < -\log_2 \frac{1 - 2^{-\frac{B_{j,k}}{M-1}}}{\sigma_z^2 + (|\mathcal{U}_j|-1) P_j d_{j,k}^{-\alpha} M 2^{-\frac{B_{j,k}}{M-1}} + \sum_{n\in\mathcal{D}, n\neq j} P d_{n,k}^{-\alpha} M 2^{-\frac{B_{n,k}}{M-1}}} \triangleq -\Omega_k.$$

We can achieve the minimum of ΔC in (3) by minimizing ΔC_k for all k, and thus we focus on the upper bound of ΔC_k. Since the minimization of the upper bound of ΔC_k is equivalent to the maximization of Ω_k, we can formulate the problem which allocates the feedback bits for the k-th user with $n \in \mathcal{D}$ as

$$B_{n,k}^{\text{real}} = \arg \max_{B_{n,k}\in[0, B_k^t]} \Omega_k, \quad \text{s.t.} \sum_{n\in\mathcal{D}} \big\lceil B_{n,k}^{\text{real}} \big\rfloor = B_k^t \quad (9)$$

where $B_{n,k}^{\text{real}}$ is a real value and $\lceil\cdot\rfloor$ indicates the round operation. Next, we propose a new feedback bit allocation algorithm which iteratively maximizes Ω_k in the following.

B. Bit Allocation Algorithm

Since (9) is a coupled problem, i.e. all $B_{n,k}$ for $n \in \mathcal{D}$ are related with each other, we propose an iterative method in order to obtain the feedback bits. The partial derivative of Ω_k with respect to $B_{l,k}$ for $l \in \mathcal{D}$ and $l \neq j$ is written as

$$\frac{\partial \Omega_k}{\partial B_{l,k}} = \frac{-2^{-\frac{B_k^t - a_{j,l}}{M-1}}(C_1+C_3)2^{\frac{B_{l,k}}{M-1}} + 2^{-\frac{B_{l,k}}{M-1}} C_2 - 2\cdot 2^{-\frac{B_k^t - a_{j,l}}{M-1}} C_2}{(M-1)\, C_4\Big(C_1 + 2^{-\frac{B_{l,k}}{M-1}} C_2 + C_3 2^{-\frac{B_k^t - a_{j,l}}{M-1}} 2^{\frac{B_{l,k}}{M-1}}\Big)^2}$$

where $a_{j,l} = \sum_{n\in\mathcal{D}, n\neq j,l} B_{n,k}$, $C_1 = \sigma_z^2 + \sum_{n\in\mathcal{D}, n\neq j,l} P d_{n,k}^{-\alpha} 2^{-\frac{B_{n,k}}{M-1}} M$, $C_2 = P d_{l,k}^{-\alpha} M$, $C_3 = (|\mathcal{U}_j|-1) P_j d_{j,k}^{-\alpha} M$ and $C_4 = \Big(1 - 2^{-\frac{B_k^t - a_{j,l}}{M-1}} 2^{\frac{B_{l,k}}{M-1}}\Big) / \Big(C_1 + 2^{-\frac{B_{l,k}}{M-1}} C_2 + 2^{-\frac{B_k^t - a_{j,l}}{M-1}} 2^{\frac{B_{l,k}}{M-1}} C_3\Big)$.

With fixed $a_{j,l}$, $B_{l,k}$ for $l \neq j$ which maximizes (9) is obtained either at the critical point of $\frac{\partial \Omega_k}{\partial B_{l,k}} = 0$, or at the boundary points. Thus, for the k-th user, the feedback bits are given by

$$B_{l,k}^{\text{real}} = \left((M-1)\log_2 \frac{-C_2 + \sqrt{C_2^2 + 2^{\frac{B_k^t - a_{j,l}}{M-1}} (C_1+C_3)\, C_2}}{C_1 + C_3}\right)^+ \quad \text{for } l \neq j \quad (10)$$

$$B_{l,k}^* = \big\lceil B_{l,k}^{\text{real}} \big\rfloor, \quad B_{j,k}^* = B_k^t - \sum_{l\neq j} B_{l,k}^* \quad (11)$$

where $(A)^+$ denotes $\max(0, A)$. Note that $B_{l,k}^{\text{real}}$ is dependent only on the transmit power and distance, and thus each user can determine the number of feedback bits for all links regardless of channel realizations. Also, when each DA port supports only one user and each user is also served by only one DA port, C_3 vanishes in (10) and (10) becomes equivalent to the solution in [7].

In the case of $|\mathcal{D}| \leq 2$, we can directly determine the feedback bits from (10) and (11). However, when $|\mathcal{D}| \geq 3$, to compute the number of feedback bits for all links, we need an iterative procedure, since $B_{l,k}$ calculated from (10) affects the solution for other channel links. Thus, we first initialize $B_{l,k}$ for $l \in \mathcal{D}$ and $l \neq j$, then compute $a_{j,l}$ and update $B_{l,k}^*$ for $l \in \mathcal{D}$ and $l \neq j$. Finally, we obtain the sub-optimal feedback bits by repeating this process until convergence. The proposed algorithm is summarized as follows:

for $k = 1 : K$
  1) Initialize $B_{l,k}$ for $l \in \mathcal{D}$ and $l \neq j$
  2) Obtain $B_{l,k}^{\text{real}}$ from (10) for $l \in \mathcal{D}$ and $l \neq j$
  3) Update $B_{l,k} = \lceil B_{l,k}^{\text{real}} \rfloor$
  4) Repeat 2) and 3) until convergence
  5) Obtain $B_{j,k} = B_k^t - \sum_{l\neq j}^K B_{l,k}$
end

Since $B_{l,k}$ determined from (10) maximizes Ω_k, Ω_k with the updated feedback bits increases in every iteration. Also, Ω_k and $B_{l,k}$ converge, since Ω_k is upper-bounded by $\mathcal{E}\left[\log_2\left(1 + \text{SINR}_k^{\text{LFB}}\right)\right]$, and thus our algorithm is guaranteed to converge. Simulations show that our algorithm needs only 2 to 4 iterations for convergence.

IV. Simulation Results

In this section, we present simulation results to demonstrate the efficacy of our proposed algorithms. In the simulation, the cell radius and the path loss exponent are set to $R = 1$ and $\alpha = 3.75$, respectively. Also, SNR is defined as P/σ_z^2. We assume that all users have the same number of total feedback bits. First, we consider a system with the fixed locations of DA ports and users depicted in Figure 1. Since we assume that the pairing between users and DA ports is determined by the distance, user 1 is supported by DA port 1, while both user 2 and 3 are served by DA port 2 and other DA ports are turned off.

In this environment, Figure 2 plots the number of allocated bits for all users with respect to SNR with $B_k^t = 18$. For user 1, more bits are allocated for the desired channel link at low SNR, i.e., the noise limited regime, while more CSI for the interfering channel link is required in order to reduce interference as SNR increases. On the other hand, the number

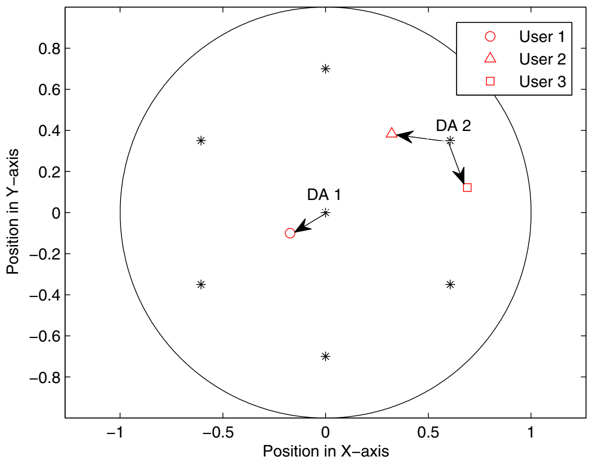

Fig. 1. Locations of 7 DA ports and 3 users

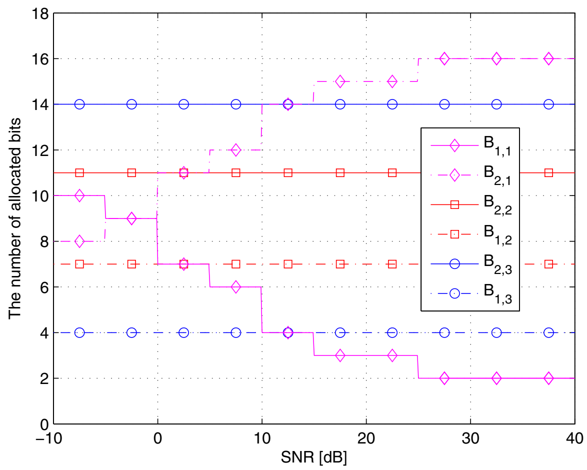

Fig. 2. Feedback bit allocation with respect to SNR

of allocated feedback bits does not change with SNR for user 2 and 3. The reason for this is that user 2 and 3 receive both the desired signal and the intra-DA port interference from DA port 2. Thus, for both user 2 and 3, several bits should be allocated to the channel linked to DA port 2 even at high SNR and the number of allocated feedback bits remains constant regardless of SNR. Also, since user 2 is closer to DA port 1 than user 3, the inter-DA port interference received at user 2 is much larger. Therefore, the number of allocated bits $B_{1,2}$ is greater than $B_{1,3}$.

Next, we consider a system where all users are uniformly distributed in a cell. We do not compare the performance of our proposed algorithm with [4] and [5] since the bit allocation methods of [4] and [5] cannot be directly applied to our system due to the limitation on the system configuration. In Figure 3, we present the average sum rate curves as a function of SNR for DAS with $K = 3$, $N = 7$ and $M = 3$. In this figure, full CSI means perfect knowledge of the linked channels at the transmitter and equal bit allocation indicates that feedback bits for all channel links are equally allocated. At SNR = 40dB, the proposed bit allocation has performance gains of 6%, 14% and 20% over the equal bit allocation for $B_k^t = 6$, 12 and 18, respectively. Also, it is shown that our pairing provides a sum rate gain over the pairing considered in [7]. For the pairing in [7], it is assumed that each DA port supports only one user and

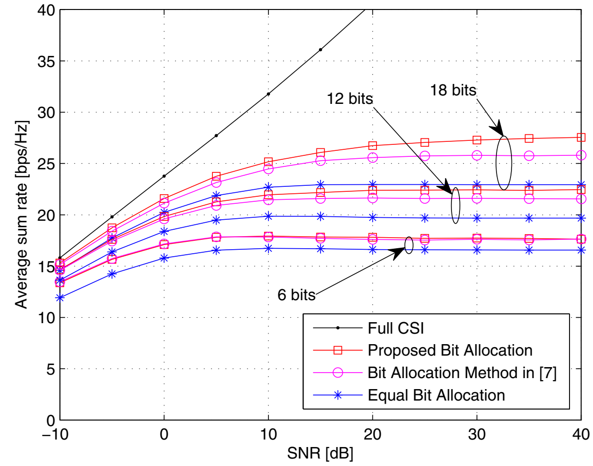

Fig. 3. The average sum rate with $K = 3$, $N = 7$ and $M = 3$

each user is also served by only one DA port, and thus users may be supported by the second closest or the third closest DA port. For this reason, the desired signal power with the pairing in [7] can be less than that with our scheme where each user is supported by the nearest DA port. The simulation results demonstrate that our proposed algorithm enhances the average sum rate compared to the conventional bit allocation methods.

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