Adaptive Bit Allocation Methods for Multi-cell Joint Processing Systems with Limited Feedback

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Abstract—In this paper, we study multiple-input single-output joint processing (JP) systems with limited feedback where two adjacent base stations exchange both channel state information and their data. To optimize the sum-rate performance of the JP system, we propose a new feedback bit allocation method which maximizes quantization accuracy in the presence of pathloss. The quantization accuracy is formulated by the expectation of the inner product between the actual channel vector and the quantized channel vector. In order to maximize the quantization accuracy, we employ a new method which compensates the phase difference of the two channels. Through numerical evaluations, we show that our proposed feedback bit allocation strategies provide about 50% performance gain in terms of the sum rate performance compared to the conventional method with the equal bit allocation scheme.

I. INTRODUCTION

Base station (BS) cooperation is an effective strategy to increase data rates and reduce network outages in cellular systems [1]–[6]. The cooperation methods at the BS can be used to combat co-channel interference, and allow more aggressive frequency reuse, which can lead to higher throughput. Among BS cooperation methods, coordinated multi-point transmission (CoMP) has recently received a lot of attention [4] [7]. This scheme has been proposed as a technique to mitigate interference in multi-cell downlink networks. In [4], the CoMP scheme with dirty paper coding (DPC) was first introduced for systems with single antenna transmitters and receivers in each cell. Also, the authors in [7] studied the maximum achievable common rate in a coordinated network based on zero-forcing (ZF) receivers and the DPC. In addition, the CoMP scheme has been proposed for general cellular systems in [8], for code division multiple access systems [9] [10], and for orthogonal frequency division multiplexing systems [11]. The CoMP techniques are also being considered for inclusion in the 4G system standard such as Long Term Evolution-advanced [12].

Generally, the CoMP can be categorized into two classes. One is coordinated beamforming (CB) which shares channel state information (CSI) only among BSs, and the other is joint processing (JP) which exchanges the CSI and their data. For both CoMP techniques, the CSI can be obtained at the transmitter via reciprocity in time division duplex (TDD) systems. However, it may be difficult to achieve in frequency division duplex (FDD) systems where receivers use finite bandwidth channels to feed back the CSI to BSs. Therefore, in the FDD systems, the concept of limited feedback is often applied [13] [14] for practical implementation. While the limited feedback for single-cell systems has been well studied [14], comparatively less work has been done in the multicell scenario [15] [16]. In contrast to the single-cell case, the CSI of multiple channels needs to be fed back in multicell systems to consider the difference in the signal strength of the desired and interfering channels for cooperation based strategies.

In multicell limited feedback CB systems, authors in [15] have recently proposed a limited feedback bit allocation scheme where available bits are assigned for the desired channel and the interfering channel. A main goal of their scheme is to minimize a mean loss in the sum-rate due to the quantization using random vector quantization (RVQ). In this paper, unlike the CB system [15] which shares only the CSI, we consider the JP system in the presence of pathloss where two adjacent BSs exchange both the CSI and their data. To optimize the sum-rate performance of the limited feedback JP system, we propose a feedback bit allocation technique to maximize the quantization accuracy, which is formulated by the expectation of the inner product between the actual channel vector and the quantized channel vector. To improve the accuracy of the quantized channel vectors, we take into account the phase compensation factor which adjusts the difference between two channels.

Available bits of our scheme are partitioned into three parts. The first and the second part are assigned for the desired channel and the interfering channel, respectively, while the third part is employed for the compensation of the phase difference. Because of the applied phase compensation technique, the quantization accuracy improves, which leads to the enhanced performance. Simulation results show that using the proposed adaptive feedback algorithm, the sum-rate provide about 50% performance gain over the equal bit partitioning technique.

The paper is organized as follows: In Section II, we describe a system model for the two-cell JP system. Section III proposes feedback bit allocation schemes to improve the performance of limited feedback systems. In Section IV, simulation results are presented to demonstrate the average
sum-rate performance of the proposed feedback bit allocation strategy. Finally, Section V concludes this paper.

Throughout this paper, boldface letters indicate vectors, boldface uppercase letters designate matrices and normal letters stand for scalar values. Moreover, $(\cdot)^H$ represents conjugate transpose. The notation $E[\cdot]$ denotes the expectation operator, and $\| \cdot \|$ is defined as the Euclidean norm of a vector.

II. SYSTEM MODEL

In this section, we provide a system model for two-cell multiple-input single-output (MISO) JP systems with limited feedback as shown in Fig. 1. In this figure, there are two BSs equipped with $M$ transmit antennas, and each BS supports $K$ users with a single antenna. We assume $M \geq K$ as in [17]. User $(i, l)$ represents the $l$-th user in the $i$-th cell for $i = 1, 2$ and $l = 1, \ldots, K$. The channel vector for user $(i, l)$ is written as $
abla(i, l) = \begin{bmatrix} P(i, l) h(i, 1)^H \sqrt{P(i, 2)} h(i, 2)^H \end{bmatrix}^H$, where $P(i, j)$ is the pathloss from the $j$-th BS to user $(i, l)$ for $j = 1, 2$ and $h(i, j)$ indicates the $M \times 1$ channel vector from the $j$-th BS to user $(i, l)$. Here, it is assumed that all channel coefficients are drawn from an independent and identically distributed (i.i.d.) complex Gaussian distribution with zero mean and unit variance.

Defining $\nu(i, l) = \begin{bmatrix} (\nu(i, 1))^H \sqrt{P(i, 2)} (\nu(i, 2))^H \end{bmatrix}^H$ where $\nu(i, j) \in \mathbb{C}^{M \times 1}$ represents the beamforming vector for user $(i, l)$ utilized in the $i$-th BS, the received signal at user $(i, l)$ for $i = 1, 2$ and $l = 1, \ldots, K$ can be given as

$$y(i, l) = (\nabla(i, l))^H s(i, l) + (\nabla(i, l))^H \sum_{k \neq l} s(k, l) + \nu(i, l) \cdot h(i, l) + n(i, l),$$

where $s(i, l)$ equals the data symbol for user $(i, l)$ with $E[|s(i, l)|^2] = 1$, $\bar{\tau}$ stands for the other cell ($\bar{\tau} = 2$ and $\bar{\tau} = 1$), and $n(i, l)$ denotes the complex white Gaussian noise with zero mean and unit variance at user $(i, l)$. We adopt the per-BS power constraint $\sum_{l=1}^K \| \nabla(i, l) \|^2 \leq P_i$ for $i = 1, 2$ [18].

In this paper, we consider limited feedback systems for designing the beamforming vector. The quantized CSI of both the desired and interfering channels is sent over the feedback link. In our work, we assume that users can perfectly estimate the desired and interfering channels using separate training symbols from BS 1 and BS 2. Let us define the channel quality information (CQI) and channel direction information (CDI) as $\| h(i) \|$ and $\| h(i, l) \|$ respectively. We assume that a user feeds back the quantized CSI with total $B_i$ feedback bits.

In this paper, we focus on the quantization process of the CQI assuming that BSs perfectly know the CQI $\| h(i, l) \|$. Thus, substituting $\hat{h}(i, l) = h(i, l)/\| h(i, l) \|$ into the actual channel vector, we rewrite $\hat{h}(i, l)$ for user $(i, l)$ as

$$\hat{h}(i, l) = \sqrt{\| h(i, l) \|} \begin{bmatrix} h(i, 1) \| h(i, 2) \end{bmatrix}. \quad (2)$$

Then the quantized CQI vector $\bar{\nu}(i, l)$ for $\hat{h}(i, l)$ is mapped to a codeword [19] as

$$\bar{\nu}(i, l) = \arg \max_{\nu(i, l)} \| \bar{\nu}(i, l) \| = \arg \max_{\nu(i, l)} | \nu(i, l) \|^2, \quad (3)$$

where $\nu(i, l)$ and $\nu(i, l)$ indicate a codeword and a codebook for the $j$-th BS.

III. PROPOSED LIMITED FEEDBACK BIT ALLOCATION

In this section, we propose feedback bit allocation methods to assign bits for the channel vectors of BS 1, BS 2 and the phase compensation, which are denoted by $B_1^{\nu(i, l)}$, $B_1^{\nu(i, l)}$ and $B_1^{\nu(i, l)}$, respectively, for user $(i, l)$. Note that $B_1^{\nu(i, l)}$ depends on the relative strengths of the desired and interfering signal. Hence, $\bar{\nu}(i, l)$ is quantized using two separate codebooks of size $2^B_1^{\nu(i, l)}$.

A main goal of our proposed scheme is to provide the optimal feedback bit allocation which maximizes the quantization accuracy for a given amount of feedback bits per user. Here, we define the quantization accuracy the expectation of the inner product between the actual channel vector and the quantized channel vector. The quantization accuracy is represented by

$$\Psi = E\left[ \left( \bar{\nu}(i, l) \| \nu(i, l) \|^2 \right)^2 \right], \quad (4)$$

where the quantized channel vector $\bar{\nu}(i, l)$ is denoted by

$$\bar{\nu}(i, l) = \sqrt{\| h(i, l) \|} \begin{bmatrix} h(i, 1) \| h(i, 2) \end{bmatrix}. \quad (5)$$
Note that the maximum quantization accuracy is achieved by $\hat{h}_{i,j} = h_{i,j}$, which accounts for the case with infinite bits. Substituting (2) and (5) into (4), the quantization accuracy is given as

$$\Psi = E \left[ P_{i}^{(i,1)} \| h_{i}^{(1,i)} \| ^{2} \left( \frac{\hat{h}_{i}^{(1,i)}}{h_{i}^{(1,i)}} \right) H_{i}^{(1,i)} \right] + P_{i}^{(i,2)} \| h_{i}^{(2,i)} \| ^{2} \left( \frac{\hat{h}_{i}^{(2,i)}}{h_{i}^{(2,i)}} \right)^{2} \right]. \quad (6)$$

Denoting $a_1$ and $a_2$ as $a_1 = \left( \frac{\hat{h}_{i}^{(1,i)}}{h_{i}^{(1,i)}} \right) H_{i}^{(1,i)}$ and $a_2 = \left( \frac{\hat{h}_{i}^{(2,i)}}{h_{i}^{(2,i)}} \right) H_{i}^{(2,i)}$, respectively, we can see that equation (6) is maximized when the phase between $a_1$ and $a_2$ is zero. Thus, we want to minimize the phase difference between $a_1$ and $a_2$ which is represented as

$$\theta_{1}^{(i)} = \theta_{a1} - \theta_{a2} \mod 2\pi, \quad (7)$$

where $\theta_{a1}$ and $\theta_{a2}$ denote the phase of $a_1$ and $a_2$, and mod represents the modulo operation.

Here, $\vartheta_{1}^{(i)}$ is uniformly distributed in $[0, 2\pi]$. Therefore, uniform quantization is performed on $\vartheta_{1}^{(i)}$ to produce $\vartheta_{1}^{(i)}$ with $B_{i}^{(i)}$ bits. It can be easily shown that the quantized phase difference $\hat{\theta}_{i}^{(i)}$ can be modeled as

$$\hat{\theta}_{i}^{(i)} = \theta_{1}^{(i)} + \gamma U, \quad (8)$$

where $\gamma U$ indicates the quantization error with $\gamma = 2\pi/(2 \cdot 2B_{i}^{(i)})$ and a random variable $U$ is uniformly distributed in $[0, 1]$.

By replacing $a_2$ with $a_2 e^{j\hat{\theta}_{i}^{(i)}}$, we can compensate the phase difference between $a_1$ and $a_2$.

![Fig. 2. An example of the phase compensation](image)

Fig. 2 illustrates an example. When $B_{i}^{(i)} = 1$, the number of available candidates for $a_2$ is two like in Fig. 2(a). Then, $a_2$ moves to $a_2'$ which is the closest candidate to $a_1$ through the uniform phase compensation with fixed $a_1$. Also, when $B_{i}^{(i)} = 2$, there are four available candidates as in Fig. 2(b). In this case, $a_2'$ is selected since this minimizes the phase difference. In this way, we can compensate the phase difference between $a_1$ and $a_2$ with the quantized phase information. Note that the phase difference between $a_1$ and $a_2$ reduces as $B_{i}^{(i)}$ increase.

Now, equation (5) can be rewritten as

$$\hat{h}_{i}^{(i)} = \sqrt{P_{i}^{(i,1)} \| h_{i}^{(1,i)} \| ^{2}} \left( \frac{\hat{h}_{i}^{(1,i)}}{h_{i}^{(1,i)}} \right) e^{j\theta_{1}^{(i)}}. \quad (9)$$

Since the CDI and the CCI are independent, substituting (2) and (9) into (4) yields

$$\Psi = E \left[ P_{i}^{(i,1)} \| h_{i}^{(1,i)} \| ^{2} \left( \frac{\hat{h}_{i}^{(1,i)}}{h_{i}^{(1,i)}} \right) H_{i}^{(1,i)} \right] + P_{i}^{(i,2)} \| h_{i}^{(2,i)} \| ^{2} \left( \frac{\hat{h}_{i}^{(2,i)}}{h_{i}^{(2,i)}} \right)^{2} \right]. \quad (10)$$

In order to compute equation (10), we first obtain

$$E \left[ \| h_{i}^{(1,i)} \| ^{4} \right] \text{ as} \quad E \left[ \| h_{i}^{(1,i)} \| ^{4} \right] = E \sum_{m=l}^{M} \sum_{n=\neq l}^{M} \left| h_{l,m}^{(i,j)} \right|^2 \left| h_{l,n}^{(i,j)} \right|^2 \right] \quad (11)$$

where $h_{l,j}^{(i,j)}$ is defined as channel element from the $j$-th BS in the $i$-th cell to the $k$-th user. Based on the central moments equation [20], we can obtain $E \sum_{m=l}^{M} \left| h_{l,m}^{(i,j)} \right|^2 = 2$ and $E \left[ \left| h_{l,m}^{(i,j)} \right|^2 \right] = 1$. Thus, equation (11) is computed as

$$E \left[ \| h_{i}^{(1,i)} \| ^{4} \right] = 2M + (M^2 - M) \cdot 1 = M = M. \quad (12)$$

Next, from Lemma 1 in [19], $E \left[ \left| h_{l,j}^{(1,i)} \right|^2 \right]$ and $E \left[ \left| h_{l,j}^{(2,i)} \right|^2 \right]$ in (10) can be expressed as

$$E \left[ \left| h_{l,j}^{(1,i)} \right|^2 \right] = E \left[ \left| h_{l,j}^{(2,i)} \right|^2 \right] = M \cos^2(\angle(h_{l,j}^{(1,i)}, h_{l,j}^{(1,i)})) \approx 1 - 2 \frac{\theta_{1}^{(i)}}{2}, \quad (13)$$

where $\angle(x, y)$ indicates the angle between vectors $x$ and $y$. Similar to equation (12), we can see $E \left[ \| h_{i}^{(1,i)} \| ^{2} \right] = M$. Also, we obtain $E \left[ \cos(\gamma U) \right]$ in (10) as

$$E \left[ \cos(\gamma U) \right] = \int_{0}^{1} \cos(\gamma t) dt = \frac{1}{\gamma} \sin \gamma. \quad (14)$$
is the received signal power at the reference distance $D_0$ and $d_{i,j}$ indicates the distance between user $(i, l)$ and the $j$-th BS. In this simulation, we set as $P_0 = 1$. Also, we assume that the cell radius is equal to $D_0 = 4$.

Fig. 3 illustrates the results of the proposed bit allocation scheme obtained from the problem (16) with $M = K = 2$ and $B_t = 8$. This figure plots the feedback bits allocated to the serving BS, the interfering BS and the compensated phase depending on the user position. In the $x$ axis, BS 1 and BS 2 are located at $-4$ and 4, respectively. Thus, the position at 0 represents cell edge. In Fig. 3, we can see that the user location from $-1.8$ to 1.8 becomes the cooperative region. When a user belongs to this region, more precise CSI for the interfering channel is required, because the user tries to reduce inter-cell interference by cooperating with an adjacent BS. In contrast, in the non-cooperative region where the user location is smaller than $-1.8$ or greater than 1.8, the CSI of the interfering channel is no longer necessary since intra-cell interference of the serving BS becomes dominant compared to the inter-cell interference.

The region boundary obtained from this results provides meaningful information on the system design for BS cooperation. If a user in the non-cooperative region is scheduled, then the serving BS can support this user without any cooperation with the interfering BS. Therefore, by simply checking which region user is located at, we can determine whether cooperation with the interfering BS is necessary or not. Obviously, boundary points in Fig. 3 depend on the channel model and pathloss. In addition, we can observe that the number of bits for the compensated phase increases, as a user gets closer to the cell edge. This is show that the compensated phase method is effective in cooperative region, in terms of quantization accuracy.

Fig. 4 exhibits the average sum-rate curves as a function of the system SNR for $M = K = 2$. It is assumed that users are uniformly distributed in the range of $(-4, 4)$. As shown in Fig. 4, the proposed method provides about 52% and 63% performance gains over the equal bit allocation in the system with $B_t = 7$ and $B_t = 9$, respectively.

Similarly, Fig. 5 shows the average sum rate results for $M = K = 3$, we can see that our schemes outperform the equal bit allocation strategy. In this figure, our scheme achieves about 66% and 74% performance gains compared to the equal bit allocation method in $B_t = 7$ and $B_t = 9$, respectively. Comparing with the results in Fig. 4, we can check that a sum rate performance gain over the equal bit allocation of $M = K = 3$ is greater than that of $M = K = 2$. This is due to a fact that as $M$ and $K$ increase, the intra-cell and inter-cell interference increase and thus the performance becomes sensitive to the quantization error. The equal bit allocation scheme is not able to compensate the quantization error, whereas our scheme adaptively allocates the feedback bits per user according to the user location and the phase difference. From the simulation results, it is clear that a performance gain of the proposed feedback allocation scheme
over the equal allocation scheme grows as $B_t$ increases.

V. CONCLUSION

In this paper, we have proposed feedback bit allocation methods to improve the sum-rate performance in MISO multicell JP systems. To enhance the sum-rate performance, we provide a new feedback bit allocation metric which maximizes the quantization accuracy by adaptively allocating the feedback bits per user according to the user location and the phase difference. When a user is close to cell edge, we have better performance by considering the compensated phase factor. Simulation results show that our proposed feedback bit allocation strategies outperform the equal allocation method. We verify that the proposed multi-cell limited feedback scheme is efficient in maximizing the sum-rate performance by utilizing the CDI quantization and the phase compensation.

REFERENCES


