Capacity Evaluation with Limited Feedback for Amplify-and-Forward MIMO Relay Channels in Urban Environment

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***Abstract*—The combination of multiple-input multiple-output (MIMO) and relay technology promises to increase coverage and system capacity, while the actual performance of MIMO relay system highly depends on the practical channel conditions. Based on the propagation measurement data in urban environment, we deal with the capacity evaluation for a two-hop amplify-and-forward MIMO relay channel in this paper. Since achieving the theoretic capacity requires the channel state information of both hops at the relay node, a limited feedback scheme is introduced. Distinguished from the conventional schemes, both precoding matrix and channel eigenvalues are quantized and fed back, and a novel codebook design method for the eigenvalues is proposed. Adopting the presented scheme, the values of ergodic capacity for various antenna configurations, precoding codebooks, feedback quantities and power allocation strategies are finally evaluated by real channel responses. The numerical results indicate that the proposed scheme outperforms the existing ones and the channel eigenvalues are essential for feedback.**

I. Introduction

The introduction of wireless relaying in cellular network shows lots of virtues to provide reliable transmission, broad coverage and high capacity. The combination of multiple-input multiple-output (MIMO) and relay technology promises even higher throughput for radio communication systems. In recent years, many literature have investigated on the information theoretic limits for the MIMO relay channels with different modes. For decode-and-forward (DF) relaying, the bounds on ergodic capacity has been developed in [1], where a single MIMO relay node is considered and the channels are assumed to be Gaussian or Rayleigh fading. In the case that amplify-and-forward (AF) mode is enabled, an asymptotical quantitative result of capacity is presented in [2], where multiple relay nodes cooperate to achieve distributive diversity.

This paper concentrates on the AF MIMO relay system since the AF strategy has potential advantages on flexibility and simplicity [3]. Considering an AF MIMO relay station (RS) that supports a dual-hop transmission between the base station (BS) and mobile station (MS), the optimal linear transceiver design for the RS has been developed [4] [5] via optimal power allocation (OPA). It is proved that such a dual-hop channel can be decomposed into several orthogonal subchannels, only if the channel state information (CSI) of both BS-RS and RS-MS links is perfectly known to the relay. Nevertheless, it is infeasible in practical frequency division duplex (FDD) systems. To solve the problem, a limited feedback joint precoding scheme for AF MIMO relaying is presented in [6]. A distributive codeword selection criterion for the precoders at both BS and RS is derived to reduce the feedback delay, and the codebook design in such a joint precoding system is also investigated.

Note that this limited feedback precoding scheme is evaluated by the Rayleigh fading channels, whereas the eigenvalues of channel matrices are close to each other in this case. Therefore, considering the marginal improvement of OPA, a simple uniform power allocation (UPA) strategy is adopted instead in [6]. However, in real channel conditions, hardly can the Rayleigh fading condition hold, and the channel eigenvalues are commonly distinct. As a result, the performance of OPA strategy may sharply increases. Hence the OPA is indispensable, when the channel eigenvalues are available at the RS.

In this paper, we deal with the capacity evaluation for an AF MIMO relay channel based on the propagation measurement data, from which the channel impulse responses (CIRs) for different antenna configurations are reconstructed. In order to approach to the theoretic capacity, a limited feedback scheme is introduced, in which both precoding matrix and channel eigenvalues are quantized and fed back. Specially, a novel codebook design method employing the minimum mean square error (MMSE) criterion is proposed for the channel eigenvalues, since the codewords derived from the MMSE criterion provide good approximations of the inputs. With the presented scheme, the values of ergodic capacity for various antenna configurations, precoding codebooks, feedback quantities and power allocation strategies are finally evaluated by real channel responses. The numerical results indicate that the proposed scheme has a better performance and the channel eigenvalues are essential for the further improvement of ergodic capacity.

II. Measurement Description

A. Measurement System

An extensive measurement campaign was performed at the center frequency of 2.35 GHz with 50 MHz bandwidth, utilizing the Elektrobit PropSound channel sounder. As described in detail in [7], the sounder works in a time-division multiplexing (TDM) mode. Thus periodic pseudo random binary signals are transmitted between different Tx-Rx antenna

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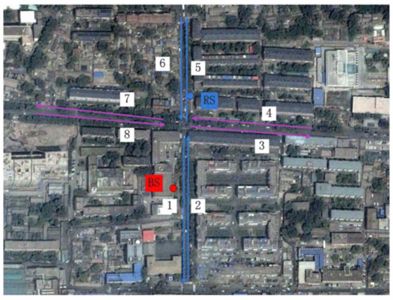


Fig. 1. Measurement environment and route plan

pairs in sequence. The interval within which all antenna pairs are sounded once is defined as a measurement cycle. In order to capture the spatial characteristics with a high accuracy, the omni-directional array consisting of 16 cross-polarized elements was employed at BS, RS, and MS. All the elements are aligned in a circle, and the spacing between the neighboring elements is half a wavelength.

B. Measurement Environment

The field measurement was conducted in a typical urban area of Beijing, China. The airscape of the measurement scenario is illustrated in Fig. 1. The antennas of the BS were installed on the rooftop of a 5-floor building, which was approximately $22\mathrm{m}$ in height. In accordance with the practical network deployment, the height of RS antennas is commonly considered to be lower than the BS to reduce the maintenance cost. Therefore the RS antennas were adjusted to $7\mathrm{m}$ above the street. The horizontal distance between the BS and the RS is $107\mathrm{m}$, and a clear line-of-sight (LoS) propagation exists. In order to simulate a user device, the MS antennas were fixed on a trolley with the height of $1.8\mathrm{m}$. The MS antennas moved along eight continuous routes throughout the measurement. Investigated on the RS-MS link, routes 1,2,5,6 are in LoS condition, and routes 3,4,7,8 are in non line-of-sight (NLoS) condition. The positions of MS were recorded by Global Positioning System (GPS).

C. Data Post Processing

In data post processing, the CIRs are converted from the raw data firstly. Then Space-Alternating Generalized Expectation maximization (SAGE) algorithm [8] is applied to extract channel parameters from the CIRs. As an extension of Maximum-Likelihood (ML) method, the SAGE algorithm provides a joint estimation of the parameter set $\boldsymbol{\eta}_p = [\tau_p, \nu_p, \phi_p, \varphi_p, \mathbf{A}_p]$ with no constrains on the response of antenna array. $\tau_p, \nu_p, \phi_p, \varphi_p, \mathbf{A}_p$ denote the propagation delay, the Doppler shift, the azimuth of departure (AoD), the azimuth of arrival (AoA), and the polarization matrix of the p^{th} propagation path, respectively. In order to capture all dominant paths that characterize the propagation environment exactly, totally 50 paths of strongest power are extracted for each measurement cycle, namely $P = 50$.

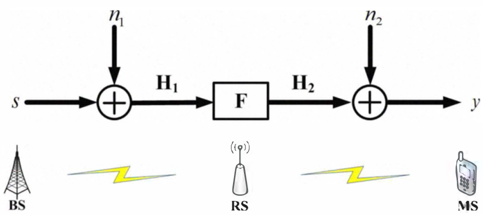


Fig. 2. System model

III. Capacity of Amplify-and-Forward MIMO Relay Channel

Consider a dual-hop MIMO relay system that consists of three nodes, in which a relay node is deployed to assist the transmission from the BS to the MS. The system model is illustrated in Fig. 2. Assume that the numbers of antennas equipped at the BS, MS and RS are M, N and L, respectively, and thus we can use $\mathbf{H}_1^{L\times M}$, $\mathbf{H}_2^{N\times L}$, and $\mathbf{F}^{L\times L}$ to denote the channel response between the BS and the RS, the RS and the MS, and the transfer function of the AF relay node. Here we focus on the relay applications intended for coverage expansion, hence the direct link between the BS and the MS can be neglected. Therefore the received baseband signal at the MS can be written as

$$\boldsymbol{y} = \mathbf{H}_2\mathbf{F}\mathbf{H}_1\boldsymbol{s} + \mathbf{H}_2\mathbf{F}\boldsymbol{n}_1 + \boldsymbol{n}_2, \tag{1}$$

where $\boldsymbol{n}_1$ and $\boldsymbol{n}_2$ denote the complex circular white Gaussian noise, i.e., $\boldsymbol{n}_1 \sim \mathcal{CN}(0, \sigma_1^2\mathbf{I}_L)$ and $\boldsymbol{n}_2 \sim \mathcal{CN}(0, \sigma_2^2\mathbf{I}_N)$, where $\mathbf{I}_{(r)}$ denotes the r-order identity matrix.

As proved in [4], the instantaneous capacity between the BS and the MS is derived as

$$C_I = \log_2\left|\mathbf{I}_L + \rho_1\mathbf{H}_1\mathbf{H}_1^\dagger - \rho_1\mathbf{H}_1\mathbf{H}_1^\dagger\mathbf{S}^{-1}\right|, \tag{2}$$

where $|\cdot|$, $(\cdot)^\dagger$, $(\cdot)^{-1}$ denote the determinant, the conjugate transpose and the inverse of the matrix, respectively, ρ_1 represents the normalized signal-to-noise ratio (SNR) at the RS, and

$$\mathbf{S} = \mathbf{I}_L + \frac{\sigma_1^2}{\sigma_2^2}\mathbf{F}^\dagger\mathbf{H}_2^\dagger\mathbf{H}_2\mathbf{F}. \tag{3}$$

Equation (2) and (3) indicate that the relay matrix $\mathbf{F}$ is essential to be designed in order to maximize the capacity of the AF MIMO relay channel. Let the eigenvalue decompositions of $\mathbf{H}_1\mathbf{H}_1^\dagger$ and $\mathbf{H}_2^\dagger\mathbf{H}_2$ be

$$\mathbf{H}_1\mathbf{H}_1^\dagger = \mathbf{U}_1\mathbf{\Sigma}_1\mathbf{U}_1^\dagger \tag{4}$$

$$\mathbf{H}_2^\dagger\mathbf{H}_2 = \mathbf{V}_2\mathbf{\Sigma}_2\mathbf{V}_2^\dagger \tag{5}$$

where $\mathbf{U}_1$ and $\mathbf{V}_2$ are unitary matrices, $\mathbf{\Sigma}_1 = \mathrm{diag}\{\alpha_1, \alpha_2, \ldots, \alpha_L\}$ with $\alpha_l \geq 0$, and $\mathbf{\Sigma}_2 = \mathrm{diag}\{\beta_1, \beta_2, \ldots, \beta_L\}$ with $\beta_l \geq 0$. We assume that all eigenvalues are arranged in the descending order for convenience. It has been disclosed in [4] that the optimal relay matrix $\mathbf{F}$ is given by

$$\mathbf{F} = \mathbf{V}_2\mathbf{\Lambda}_F\mathbf{U}_1^\dagger, \tag{6}$$

where $\mathbf{\Lambda}_F$ is a diagonal matrix that can be written as

$$\mathbf{\Lambda}_F = \frac{\sigma_2}{\sigma_1}\mathbf{\Lambda}_X(\mathbf{I}_L + \rho_1\mathbf{\Sigma_1})^{-\frac{1}{2}}, \quad (7)$$

with $\mathbf{\Lambda}_X$ representing the power allocation matrix. Denote $\mathbf{\Sigma}_X = \mathbf{\Lambda}_X{}^2 = \mathrm{diag}\{x_1, x_2, \ldots, x_L\}$, and the power constraint satisfies

$$\sum_{l=1}^{L} x_l \leq \rho_2 L, \quad (8)$$

where ρ_2 denotes the normalized SNR at the MS. According to [4], the optimal power allocation factor x_l is derived as

$$x_l = \frac{1}{2\beta_l}\max\left[\sqrt{\rho_1^2\alpha_l^2 + 4\rho_1\alpha_l\beta_l\mu} - \rho_1\alpha_l - 2\right]^+ \quad (9)$$

where $[a]^+ = \max(a, 0)$ and μ is chosen to meet (8). Define a function $g(\mu)$,

$$g(\mu) = \frac{1}{2}\sum_{l=1}^{L}\frac{1}{\beta_l}\left[\sqrt{\rho_1^2\alpha_l^2 + 4\rho_1\alpha_l\beta_l\mu} - \rho_1\alpha_l - 2\right]^+ - \rho_2 L \quad (10)$$

thereby μ is the unique root of (10), which can be solved by a numerical root-finding algorithm.

With the optimal power allocation matrix $\mathbf{\Lambda}_F$, the relay matrix given by (6) is also optimized. Therefore the capacity of AF MIMO relay channel attains to the maximum value,

IV. Limited Feedback for Amplify-and-Forward Relaying

Designing the relay matrix $\mathbf{F}$ requires full CSI of both BS-RS and RS-MS links at the relay node. Although it is reasonable to assume that $\mathbf{H}_1$ is available at the RS and $\mathbf{H}_2$ available at MS, unfortunately $\mathbf{H}_2$ is unreachable for the RS in practical FDD systems. To overcome the problem, a feedback of $\mathbf{H}_2$ is demanded at the RS. According to (6) - (9), the feedback values should not only include the precoding matrix $\mathbf{V}_2$, but also the eigenvalues $\{\beta_k\}$ of the second link, since they are necessary for the OPA.

Reviewing the limited feedback for precoding matrix, we can conclude that the existing proposals are all based on two aspects, namely codebook design and quantization criterion. The basic ideas of codebook design includes vector quantization (VQ) method [9], Grassmannian line packing method [10], etc. The details of the codebook design methods are omitted here for the limited space. Nonetheless, the codebook is fixed beforehand and is known to both RS and MS. Consider a codebook $\mathcal{C}$ consisting of 2^B matrices in $\mathbb{C}^{K_1\times K_2}$ i.e., $\{\mathbf{W}_1, \ldots, \mathbf{W}_{2^B}\}$, where B is the feedback bits for the precoding matrix $\mathbf{V}$. The quantization of a precoding matrix $\mathbf{V}$, say $\hat{\mathbf{V}}$, is chosen from the codebook $\mathcal{C}$ according to the following rule:

$$\hat{\mathbf{V}} = \arg\min_{\mathbf{W}\in\mathcal{C}} d^2(\mathbf{V}, \mathbf{W}) \quad (11)$$

where $d(\mathbf{V}, \mathbf{W})$ denotes the distance metric. Here we apply the Chordal distance that is expressed as

$$d(\mathbf{V}, \mathbf{W}) = \sqrt{\mathbf{K_2} - \sum_{\mathbf{j=1}}^{\mathbf{K_2}}\lambda_{\mathbf{j}}^{\mathbf{2}}\{\mathbf{V}^\dagger\mathbf{W}\}} \quad (12)$$

where $\lambda_j\{\mathbf{V}^\dagger\mathbf{W}\}$ denotes the j^{th} singular value of $\mathbf{V}^\dagger\mathbf{W}$. Note that the latest study [6] has proposed the optimal codeword selection scheme for the relay system. Nevertheless it is not adopted in this paper due to the high computational complexity.

Next we will discuss the limited feedback for channel eigenvalues. A novel codebook design method for the eigenvalues is proposed as follows.

Design Criterion: Let $\boldsymbol{\beta} = \{\beta_1, \ldots, \beta_L\}$ with β_l denoting the l^{th} eigenvalue of $\mathbf{H}_2^\dagger\mathbf{H}_2$. Referring to (9), we can find that the power allocation coefficients are in connection with the proportional relation among the channel eigenvalues, rather than their absolute values. Thus firstly we normalize each $\boldsymbol{\beta}$ to satisfy $\sum_{l=1}^{L}\beta_l = k$, where k is an arbitrary integer for all $\boldsymbol{\beta}$. In this case, the codebook is designed for $\overline{\boldsymbol{\beta}} = \{\beta_1, \ldots, \beta_{L-1}\}$, since the L^{th} eigenvalue can be calculated as $\beta_L = k - \sum_{l=1}^{L-1}\beta_l$.

Let $\hat{\boldsymbol{\beta}}$ denote the quantized version of $\overline{\boldsymbol{\beta}}$. We design a codebook $\mathcal{C}_\beta$ leading to the MMSE, that is,

$$\min_{\hat{\boldsymbol{\beta}}\in\mathcal{C}_\beta} E\|\overline{\boldsymbol{\beta}} - \hat{\boldsymbol{\beta}}\|^2, \quad (13)$$

where $E(\cdot)$ denotes the expectation operator and $\|\cdot\|$ stands for the module of the vector. The intention of the MMSE criterion is to obtain the codewords which approximate to any given $\overline{\boldsymbol{\beta}}$ as far as possible.

Design Algorithm: One of the virtues of the MMSE criterion is that it has a closed-form VQ design algorithm (Lloyd algorithm). The Lloyd algorithm is based on two conditions called the nearest neighborhood condition (NNC) and the centroid condition (CC). The approach is summarized as:

1. *NNC*: For given code vectors $\{\hat{\boldsymbol{\beta}}_i; i = 1, \ldots, 2^{B_\beta}\}$, where B_β is the feedback bits for each $\overline{\boldsymbol{\beta}}$, the optimum partition cells satisfy

$$\mathcal{R}_i = \{\boldsymbol{\beta}\in\mathbb{R}^L : \|\overline{\boldsymbol{\beta}} - \hat{\boldsymbol{\beta}}_i\|^2 \geq \|\overline{\boldsymbol{\beta}} - \hat{\boldsymbol{\beta}}_j\|^2, \forall i \neq j\}, i = 1, \ldots, 2^{B_\beta} \quad (14)$$

where $\mathcal{R}_i$ is the partition cell (Voronio region) for the i^{th} code vector.

2. *CC*: For a given partition $\{\mathcal{R}_i; i = 1, \ldots, 2^{B_\beta}\}$, the optimum code vectors satisfy

$$\hat{\boldsymbol{\beta}}_i = \arg\min_{\hat{\boldsymbol{\beta}}\in\mathcal{R}_i} E[\|\overline{\boldsymbol{\beta}} - \hat{\boldsymbol{\beta}}\|^2], i = 1, \ldots, 2^{B_\beta} \quad (15)$$

The above two conditions are iterated until MMSE converges. As the codebook is generated by MMSE criterion, it is natural to apply the squared error as the quantization metric. Thus for a given $\overline{\boldsymbol{\beta}}$, its quantized version $\hat{\boldsymbol{\beta}}$ is selected by the following equation.

$$\hat{\boldsymbol{\beta}} = \arg\min_{\hat{\boldsymbol{\beta}}_j\in\mathcal{C}_\beta} \|\overline{\boldsymbol{\beta}} - \hat{\boldsymbol{\beta}}_j\|^2 \quad (16)$$

V. Numerical Results

A. Channel Reconstruction

The measurement campaign described in section II provides the real CIRs in the given propagation scenario. Hence substituting $\mathbf{H}_1$ and $\mathbf{H}_2$ in (2) by the measured CIRs, we can calculate the channel capacity of the corresponding relay system. However, note that the antenna numbers (M, L, N) of the measurement platform are as large as $(16, 16, 16)$, which are almost prohibitive for the practical network deployment and equipment manufacture at present. To get more sensible results, we proceed to reconstruct $\mathbf{H}_1$ and $\mathbf{H}_2$ for different antenna configurations, e.g. antenna numbers, element spacings, field patterns, etc. Referring to [11], the channel reconstruction is of the form as follows.

$$\begin{aligned} \mathbf{H}(\tau)_{n,m} &= \sum_{p=1}^{P(\tau)} \mathbf{F}_{Rx,n}^{T}(\varphi_p)\mathbf{A}_p\mathbf{F}_{Tx,m}(\phi_p) \\ &\quad \cdot \exp\left(jd_m 2\pi\lambda_0^{-1}\sin(\phi_p)\right) \\ &\quad \cdot \exp\left(jd_n 2\pi\lambda_0^{-1}\sin(\varphi_p)\right) \end{aligned}$$

where $\tau_p, \phi_p, \varphi_p, \mathbf{A}_p$ are of the same denotations interpreted in section II and are extracted by SAGE algorithm. Moreover, $(\cdot)^T$ denotes the matrix transpose, $P(\tau)$ represents the number of paths at the given delay τ, λ_0 is the wavelength of the carrier, d_n stands for the distance between the n^{th} Rx antenna element and the first element, $\mathbf{F}_{Rx,n}$ is the field pattern of the n^{th} Rx antenna element. For Tx antenna elements, d_m, $\mathbf{F}_{Tx,m}$ hold the same meanings with d_n, $\mathbf{F}_{Rx,n}$, respectively.

In the following studies, all Tx and Rx antennas are assumed to be isotropic for simplicity, i.e., the field patterns can be rewritten as [12]

$$\mathbf{F}_{Tx,m}(\phi_p) = \begin{bmatrix} F_{Tx,m,V}(\phi_p) \\ F_{Tx,m,H}(\phi_p) \end{bmatrix} = \begin{bmatrix} \cos\theta_{Tx} \\ \sin\theta_{Tx}\cos\phi_p \end{bmatrix}$$
$$\mathbf{F}_{Rx,n}(\varphi_p) = \begin{bmatrix} F_{Rx,m,V}(\varphi_p) \\ F_{Rx,m,H}(\varphi_p) \end{bmatrix} = \begin{bmatrix} \cos\theta_{Rx} \\ \sin\theta_{Rx}\cos\varphi_p \end{bmatrix}$$

where θ indicates the slant angle between the antenna element and the vertical direction; the subscript V, H stand for the vertical and horizontal polarizations, respectively.

It should be addressed that the CIRs reconstructed by (17) represent the frequency-selective fading channels. Thus the capacity of AF MIMO relay channel is modified as

$$C_I = \frac{1}{Q}\sum_{q=1}^{Q}\log_2\left|\mathbf{I}_L + \rho_1\mathbf{H}_{1,q}\mathbf{H}_{1,q}^{\dagger} - \rho_1\mathbf{H}_{1,q}\mathbf{H}_{1,q}^{\dagger}\mathbf{S}_q^{-1}\right|$$

where $\mathbf{H}_q$ is the channel matrix in frequency domain and Q is the number of frequency bins. It is well known that $\mathbf{H}_q$ can be converted from $\mathbf{H}(\tau)$ applying Discrete Fourier Transform.

B. Channel Normalization

In order to isolate the small scale characteristics of the channel from the effects of path-loss and shadowing, we need to normalize the reconstructed channel matrix. Then we can evaluate the channel capacity at different SNRs. The normalization should be calculated differently for various antenna types. When vertically polarized antenna arrays are applied, the power of each $\mathbf{H}_q$ is normalized to satisfy

$$\|\mathbf{H}_q\|_F^2 = N_{Rx} \times N_{Tx} \quad (17)$$

where N_{Rx} and N_{Tx} denote the antennas numbers at Rx side and Tx side, respectively. While for dual-polarized antenna arrays, the cross-polarized subchannels suffer from severe power losses, which need to be taken into account in the capacity calculations [13]. Therefore we normalize the eigenvalues of $\mathbf{H}_q$ to the SNR of the strongest subchannle, that is

$$\max\left(\text{eig}\left(\mathbf{H}_q\mathbf{H}_q^{\dagger}\right)\right) = 1 \quad (18)$$

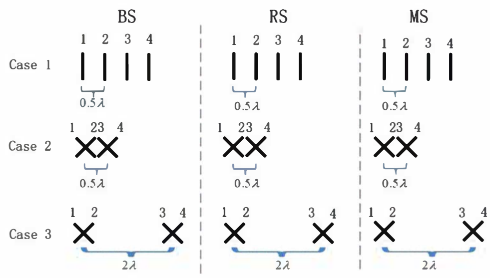


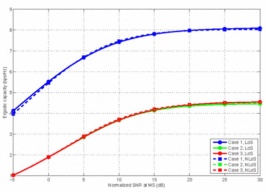
Fig. 3. Antenna configurations

C. Capacity Evaluation

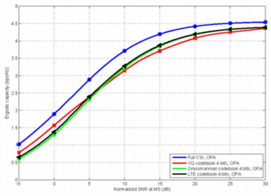
We focus on the AF MIMO relay system in which the antenna numbers (M, L, N) are set as $(4, 4, 4)$. Three kinds of antenna configurations are considered as illustrated in Fig. 3. For case 1, vertically polarized antennas are deployed and the spacing between any adjacent antennas is half a wavelength. For case 2 and 3, we use dual-polarized antennas instead, whose slant angles are $\pm 45°$. The difference between the latter two cases lies on the antenna spacings. In addition, the normalized SNR at the RS is fixed to 10 dB throughout the evaluation, i.e., $\rho_1 = 10$ dB.

Assuming the CSI of both hops is perfectly known to the RS, the maximum values of ergodic capacity under different propagation conditions (LoS/NLoS for RS-MS link) are depicted in Fig. 4(a) for all three cases. It indicates that the performance of the concerned antenna configurations is not sensitive to the propagation conditions, since the ergodic capacity does not present any significant difference under LoS and NLoS conditions. However, once dual-polarized antennas are deployed, the ergodic capacity decreases significantly, owing to the power losses incurred by the cross-polarized subchannels. The capacity decline compared to case 1 is about 3 bps/Hz when $\rho_2 = -5$ dB, and attains to 4 bps/Hz when ρ_2 is as high as 30 dB. Furthermore, comparing case 2 and case 3, we also note that enlarging the antenna spacings does not provide a visible benefit in terms of ergodic capacity in this scenario.

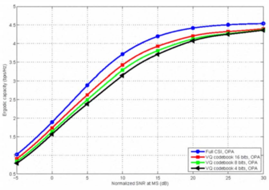
In the following studies, we concentrate on the antenna case 3 in NLoS condition and the limited feedback scheme is introduced. As shown in Fig. 4(b), the codebook derived by VQ method in consistent with [9], the Grassmannia codebook given by [14], and the codebook defined by LTE standard [15]



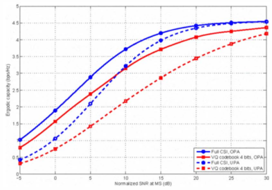
(a) Capacity comparison for antenna configurations.



(b) Capacity comparison for precoding codebooks.



(c) Capacity comparison for feedback quantities.



(d) Capacity comparison for power allocation strategies.

Fig. 4. Numerical results

are employed to quantize the precoding matrix respectively. The size of all these codebooks is fixed to 4 bits in order to make a fair comparison. Besides, for OPA strategy, another 2 bits are used to quantize the eigenvalues of $\mathbf{H}_2^{\dagger}\mathbf{H}_2$, adopting the method proposed in section IV. It shows that with the same quantity of feedback, the performance of Grassmannian codebook and LTE codebook appears extremely close to each other, but the VQ codebook presents a tiny advantage when ρ_2 is below 5 dB, and shows a little drawback thereafter.

The capacity disparity caused by feedback quantities is illustrated in Fig. 4(c). The codebook designed by VQ method is adopted for the precoder as it is flexible to change the codebook size. It is logical that when 16 bits are utilized to encode the 4×4 precoding matrix, the capacity loss due to quantization distortion is no more than 0.3 bps/Hz. However, it will lead to a considerable feedback overhead.

Fig. 4(d) shows the performance of different power allocation strategies. To implement the OPA, 4 bits are used to encode the precoding matrix and 2 bits for the channel eigenvalues. While for the UPA strategy, channel eigenvalues are not demanded thereby all 6 bits are allocated to the quantization of precoding matrix. It is observed that with the same feedback quantity, the OPA strategy always performs better. Especially when ρ_2 is lower than 10 dB, the OPA strategy with limited feedback even surpasses the UPA with full CSI. While with the increasing of ρ_2, the advantage of OPA decreases, and is nearly invisible when ρ_2 attains to 30 dB. According to Fig. 4(d), it is convinced that the OPA strategy offers prominent benefit on ergodic capacity in real channel conditions, and the channel eigenvalues are definitely worth feeding back.

VI. Conclusions

In this paper we present a limited feedback scheme for a dual-hop AF MIMO relay system. Both precoding matrix and channel eigenvalues are quantized and fed back, and a novel codebook design method for the eigenvalues is proposed. The performance of the presented scheme is evaluated in terms of ergodic capacity by real channel responses. Various antenna configurations, precoding codebooks, feedback quantities and power allocation strategies are considered. It is revealed that for a given feedback quantity, different codebooks for precoding matrix provides similar performance, whereas different power allocation strategies shows distinct effects specifically. The OPA strategy always outperforms the UPA significantly, which indicates that the channel eigenvalues are critical to be fed back.

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