

Transmit Diversity Scheme with Sounding-based Channel Phase Feedback in OFDMA Systems

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Abstract—Existing feedback schemes of Channel State Information (CSI) in OFDMA systems can be mainly sorted into two categories as follows: 1) sounding channel feedback, and 2) precoder matrix indicator (PMI) feedback. However, both these two categories have some performance drawbacks and limitation of usage. In this paper, two Sounding-based Phase Feedback (SPF) schemes, namely, *Type_1* SPF and *Type_2* SPF, are proposed as alternatives of CSI feedback methods. With the proposed feedback schemes, the OFDMA transmitters can obtain the information of channel phases and perform an approximate Equal Gain Combining (EGC) transmission. Comparing to the current sounding channel feedback schemes, these two schemes have the flexibility to be employed in both TDD and FDD modes with comparable overhead. Comparing to the PMI feedback scheme, the proposed schemes have lower computation complexity at the receiver sides since the procedures to find the optimal precoder is not needed. Moreover, the fundamental performance limit due to quantizing the channel status also does not exist in the proposed schemes. Theoretically, the proposed schemes could convey near perfect channel phase information if appropriate estimation algorithms are adopted in the transmitter for channel estimation. Numerical analysis is also provided to evaluate the average receive SNR achieved by the two proposed schemes for comparing their performance under non-ideal channel estimations. Simulation results show that the performance of the proposed schemes can achieve close performance to that of EGC (within 0.5 dB) which needs perfect channel phase information at the transmitter.

Keywords— Sounding channel, CSI feedback, OFDMA, Transmit diversity, Maximal ratio combining, Equal gain combining

I. INTRODUCTION

Transmit diversity is an important technique that brings significant gain to wireless systems with multiple transmit antennas. However, the optimal transmission strategy, namely, Maximal Ratio Combining (MRC), needs full Channel State Information (CSI) at the transmitter that is difficult to obtain in practice. In the next generation broadband communication systems such as LTE and IEEE 802.16m, two main categories of CSI feedback are supported: 1) sounding channel feedback, and 2) precoder matrix indicator (PMI) feedback. Literature [1] first provided a comprehensive overview for the sounding feedback technique. Till now, several works [2-7] have paid attention to the related issues to improve the system performance in individual aspects. However, both these two

categories have some performance drawbacks and limitation of usage that have not entirely been handled so far.

In the sounding channel feedback scheme, the receiver sends sounding channels that carry pre-known signals to the transmitter. The transmitter measures the channel responses via the sounding channels and utilizes the measurement results to the reverse link under the assumption of DL/UL channel reciprocity. In general, this feedback approach can only be applied in TDD systems since DL/UL channels are allocated in separate frequency bands in FDD systems and thus the DL/UL instantaneous channel responses are typically uncorrelated. Moreover, the property of DL/UL channel reciprocity may not always perfectly stand due to several factors such as the phase mismatch of TX/RX RF chains at the transmitter and non-identical channel impulse responses of the forward and reverse links in environments with rich scatters. The loss of reciprocity property will degrade the accuracy of CSI and make the performance unreliable. On the other hand, in PMI feedback mode, the receiver chooses the most appropriate precoder matrix from a set of finite precoder matrix and feedback its index to the transmitter. This approach is workable in both TDD and FDD modes, but performance loss is unavoidable due to the quantization of channel status. Moreover, procedures to find the most appropriate precoder may introduce considerable computation complexity at the receiver especially in wideband systems.

In this paper, two sounding-based partial feedbacks of CSI schemes for Multiple-Input-Single-Output (MISO) channels are proposed as new candidates of CSI feedback methods. The basic concept is using pairs of sounding feedback channels to convey full phase information or differential phase information of the channel responses of each spatial channel. With the phase information, the transmitter is allowed to employ an approximate Equal Gain Combining (EGC), which is known to have the close-optimal performance. Additionally, the proposed schemes have the flexibility to be employed in both TDD and FDD modes. The receiver only needs to reflect the channel phase information and does not need to search for the most appropriate precoders. We also compare the performance of the two sounding-based feedback schemes under channel estimation error by analyzing the average receive SNR. Finally, simulations are conducted to compare the symbol error rate

performance of the two proposed sounding-based feedback schemes with that of EGC.

The rest of this paper is organized as follows. In Section II, the system model and the proposed schemes are presented. Next, Section III analyzes the average receive SNR of the proposed schemes. Then, simulation results are demonstrated in Section IV. Finally, conclusions are given in Section V.

II. SYSTEM MODEL AND PROPOSED SCHEME

In this paper, we consider a wireless OFDMA system with M transmit antennas and a single receive antenna. In such a system, it is reasonable to assume that each subcarrier in a OFDMA symbol serves as a narrow band channel so that the data symbols carried in each subcarrier undergoes a frequency nonselective channel. Let $\mathbf{h}_{t,c}^H = [h_{t,c}^1 \ h_{t,c}^2 \ \dots \ h_{t,c}^M]$ be the $1 \times M$ frequency-domain channel response vector of the subcarrier c of OFDMA symbol t in which $h_{t,c}^n$ is the channel response from n -th transmit antenna to the receive antenna, and can be expressed in the form $\alpha_{t,c}^n e^{j\theta_{t,c}^n}$, where $\alpha_{t,c}^n$ is the channel gain and $\theta_{t,c}^n$ is the phase shift. Notice that $(\cdot)^H$ denotes the conjugate transpose for symbol t in subcarrier c .

Denote the complex number data symbol carried by a subcarrier in a given OFDMA symbol as s , with unity average power. The transmitted signals from each transmission antennas at the transmitter can be expressed as $\sqrt{E_s} \mathbf{w} s$. E_s is the energy of the data symbol and $\mathbf{w}$ is a $M \times 1$ precoding vector (also usually refer as beamforming vector) with the 2 norm is normalized to unity, i.e. $\|\mathbf{w}\|^2 = \mathbf{w}^H \mathbf{w} = 1$. At the receiver end, the received signal is given by $r_{t,c} = \sqrt{E_s} \mathbf{h}_{t,c}^H \mathbf{w} s + n_{t,c}$, $n_{t,c}$ is additive white noise which is assumed to be a circularly symmetric complex Gaussian random variable with zero mean and variance N_0 . We also define the transmit SNR $\rho = E_s / N_0$. Then the receive SNR at the receiver can be expressed as $\rho_{eff} = \mathbf{h}_{t,c}^H \mathbf{w} \mathbf{w}^H \mathbf{h}_{t,c} \rho$ [9].

There are several strategies to decide the precoding vector $\mathbf{w}$, depending on the CSI at the transmitter. When the transmitter has the full CSI, the optimal (i.e. maximize the received SNR) is MRC. The precoding vector of MRC is $\mathbf{h} / \|\mathbf{h}\|$, with the receive SNR $\rho_{eff}^{MRC} = \sum_{n=1}^M |\alpha_{t,c}^n|^2 \rho$. MRC is proved to achieve full diversity gain and antenna array gain [8], however, full CSI includes amplitudes and phases of channel responses is in general hard to be obtained in most practical systems because of the difficulty for the receivers to feedback the information within a reasonable latency. A suboptimal transmission strategy which is of practical interest is EGC. EGC does not need the information of the amplitudes of the channel responses. The precoding vector of EGC is given by

$\frac{1}{\sqrt{M}} [e^{-j\theta_{t,c}^1} \ e^{-j\theta_{t,c}^2} \ \dots \ e^{-j\theta_{t,c}^M}]$. It can be seen that EGC rotates the phases of the channel responses and make the received signal to be combined coherently. The receive SNR of EGC is $\rho_{eff}^{EGC} = \frac{[\sum_{n=1}^M \alpha_{t,c}^n]^2}{M} \rho$. In [9], it is proved that EGC achieve the same diversity order as that of MRC. They have the same logarithmic slope of error rate, but with slight shift due to power loss. (see results, e.g. [10]). In general, EGC has close performance to that of MRC.

Proposed Sounding-based Phase Feedback (SPF) Scheme

In the following, we introduce the proposed Sounding-based Phase Feedback (SPF) scheme, which is designed for OFDMA system to convey the channel phase information from the receiver to the transmitter and facilitate the transmitter to perform EGC transmission.

For convenience of illustration, in the remained part of this paper, we consider DL channel only. Moreover, we assume the time domain is divided into equal-sized subframes which consists of T OFDMA symbols, and a subband is a resource unit defined by the system which is a rectangular block occupying K adjacent subcarriers in a subframe. When the base station starts to trigger a CSI feedback procedure, it signals the user and indicates one or more subbands for channel information. This is a general system architecture existing in many OFDMA-based systems. The relationship of subframe and subband is shown in Fig. 1.

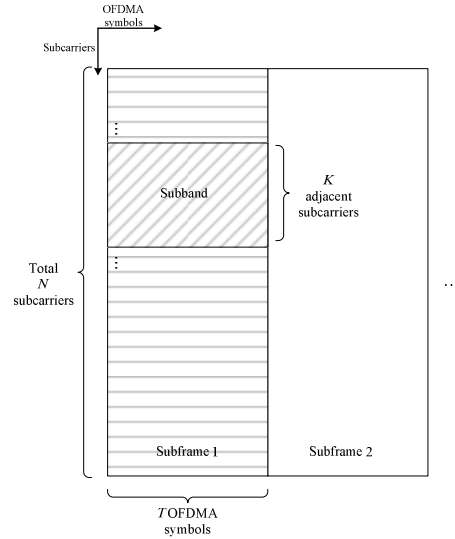


Fig. 1 Illustration of the architecture of subframe and subband

Based on the architecture described below, Fig. 2 is the structure of the proposed sounding-based phase feedback

channels (SPFC). As depicted in Fig. 2, SPFC composes of several sounding cluster (regions with the same color in the figure) which contains x adjacent subcarriers and two OFDMA symbols. The number of adjacent subcarriers x depends on the size of the subband and a system parameter, *decimation order* D , which will be described in more detail later. Each sounding cluster corresponding to a spatial channel (i.e. the channel from a transmit antenna to the receive antenna) of the subband indicated by base station.

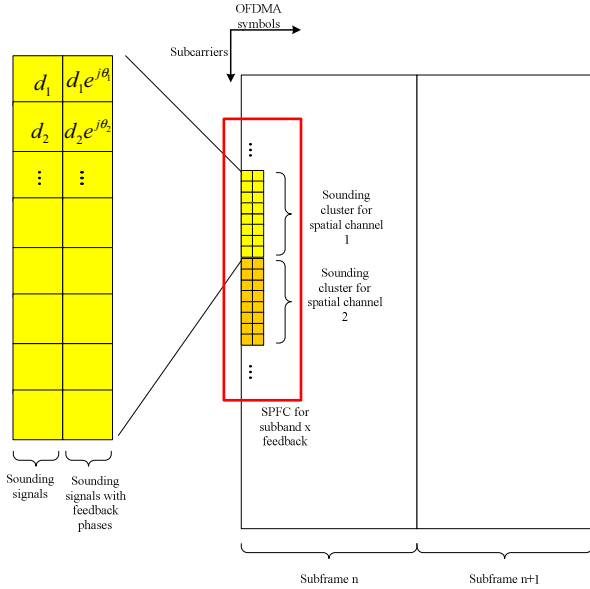


Fig. 2. The structure of the proposed Sounding-based Phase Feedback Channel (SPFC).

The first OFDMA symbol of a sounding cluster is operated as a normal sounding channel. Each subcarrier carries sounding signals $s_1, s_2, \dots, s_x$ which is pre-known by the base station and can be used to estimate the UL channel responses by a variety of conventional estimating algorithms such as Least Square (LS) [12] or Linear Minimal Mean Square Error (LMMSE) [12]. On the other hand, for the second OFDMA symbol, each subcarrier carries the sounding signals with additional phase shifts $\theta_1, \theta_2, \dots, \theta_x$, that is, the signals carried are $s_1 e^{j\theta_1}, s_2 e^{j\theta_2}, \dots, s_x e^{j\theta_x}$. After transmitting through the UL channel, the received signals after the DFT at the base station in the two OFDMA symbols are $h_1^1 s_1, h_2^1 s_2, \dots, h_x^1 s_x$ and $h_1^2 s_1 e^{j\theta_1}, h_2^2 s_2 e^{j\theta_2}, \dots, h_x^2 s_x e^{j\theta_x}$, respectively, if noise is negligible. h_c^s denotes the (frequency domain) channel responses in the c -th subcarrier of the s -th OFDMA symbol. In most scenario of interest, the channel responses of two adjacent OFDMA symbols can be considered to be with high correlation, thus we can assumed $h_k^1 = h_k^2 = h_k$, for $k = 1, 2, \dots, x$. Due to this property, if the base station can obtain

h_k^1 and $h_k^2 e^{j\theta_k}$ by channel estimation techniques via the sounding signal s_k , it can extract the additional phase information θ_k carried in the sounding cluster. This is the basic idea of the proposed phase feedback schemes.

Based on this approach, the user could convey the phases of DL channel responses to the base station and allow the base station to perform EGC transmission by adjusting the initial phases of each transmit antennas. To do this, we have two types of SPF schemes with different phase information carried and different overhead, which will be illustrated as follows. Moreover, it is unavoidable that the error of channel estimation may somewhat degrade the accuracy of phase information extracted and thus affect the performance of data transmission. We will analyze and compare the performance of the two types of SPF under phase estimation error by numerical analysis and simulation in Section III and Section IV, respectively.

Type_1 SPF:

In *Type_1* SPF, the carried phase information θ_k in the k -th subcarrier of a sounding cluster n is $\theta_n^{DL, l+(k-1)D}$. Let l and u are the smallest and largest subcarrier indices of the subband indicated by the base station, $\theta_n^{DL, l+(k-1)D}$ is the phase of the DL channel response of the subcarrier with index $(l+(k-1)D)$ from the n -th transmit antenna. That is, *Type_1* SPF samples the phases of the DL channel responses from each transmit antenna with a period D and convey the sampled phases with the corresponding sounding clusters. D is the decimation order as mentioned above. It is a system parameter and may be adjust dynamically to meet the appropriate tradeoff of estimation accuracy and feedback overhead. Each sounding cluster occupies x subcarriers, $x = \arg \max_k (l + (k-1)D) \leq u$.

Based on the sounding signals, the base station could calculate estimator of the two effective channel responses in the adjacent OFDMA symbols, i.e. h_k and $h_k e^{j\theta^{DL, l+(k-1)D}}$, and then extract the phase by subtracting the phases of the two estimators. For example, if LS method is used for channel estimation, for the k -th subcarrier in the sounding cluster, the received signals in the two OFDMA symbol are $r_k^1 = h_k s_k + n_1$ and $r_k^2 = h_k e^{j\theta^{DL, l+(k-1)D}} s_k + n_2$, where n_1 and n_2 are i.i.d. complex Gaussian random variables. The estimators of effective channel responses in LS are $\lambda_k^1 = r_k^1 s_k^{-1}$ and $\lambda_k^2 = r_k^2 s_k^{-1}$ and the extracted value of phase $\theta^{DL, l+(k-1)D}$ is obtained by $\hat{\theta}^{DL, l+(k-1)D} = \text{angle}(\lambda_k^2) - \text{angle}(\lambda_k^1)$. The other DL channel phases in the subband that are not carried in the sounding clusters may be estimated by interpolation methods. In the simulation, linear interpolation is adopted. After obtaining $\hat{\theta}^{DL, l+(k-1)D}$, the corresponding transmit antenna then precodes

the data symbol d by multiplying $e^{-j\theta^{DL,J+(k-1)D}}$, similar to the typical EGC transmission described in Section II.

On the other hand, if LMMSE method is used to obtain the estimators, the phase information can also be extracted in the same manner. However, it is worth to note that to find the LMMSE estimator of $h_k e^{j\theta^{DL,J+(k-1)D}}$, the base station needs to have the knowledge of the correlations of DL channel phases. More specifically, assuming the channel responses of DL and UL channels are uncorrelated (e.g. in a FDD system), the correlation of $h_{k_1} e^{j\theta^{DL,J+(k_1-1)D}}$ and $h_{k_2} e^{j\theta^{DL,J+(k_2-1)D}}$ is

$$E[h_{k_1} e^{j\theta^{DL,J+(k_1-1)D}} h_{k_2}^* e^{-j\theta^{DL,J+(k_2-1)D}}] = E[h_{k_1} h_{k_2}^*] E[e^{j\theta^{DL,J+(k_1-1)D}} e^{-j\theta^{DL,J+(k_2-1)D}}].$$

The first term is the correlation of the UL channel responses and can be simply obtained by UL reference signals. The second term is the correlation of the DL channel phases. The base station can also obtain this statistic from other *Type_1* SPF channels.

Type_2 SPF:

Type_1 SPF is a straightforward approach which directly conveys the DL channel phase information of each transmit antennas. However, in many possible scenarios, the channel status is quasi-static and approximately unchanged in a short interval which contains the time of channel phase feedback and corresponding data transmission. In such situations, more efficient scheme, *Type_2* SPF can be employed with lower overhead and less sensitivity to the estimation error (as proved in the next section).

Instead of conveying the complete phase value $\theta_n^{DL,J+(k-1)D}$, *Type_2* SPF conveys the differential phase value $\delta_n^{DL,J+(k-1)D} = \theta_n^{DL,J+(k-1)D} - \theta_1^{DL,J+(k-1)D}$. Then the base station obtains the estimated value of the differential phase $\hat{\delta}_n^{DL,J+(k-1)D}$ with the same method as that of *Type_1* SPF and precode the data symbol by multiplying $e^{-j\hat{\delta}_n^{DL,J+(k-1)D}}$ when transmission. By doing this, all the transmit antennas align to the DL channel phase of the first transmit antenna. Comparing to the *Type_1* SPF scheme in which all transmit antenna align to zero phase, the *Type_2* SPF can also achieve coherent combining at the receiver side like EGC with a phase shift $e^{j\theta_1^{DL,J+(k-1)D}}$. However, the sounding cluster of the first transmit antenna is not needed and thus one sounding cluster can be saved. The phase shift $e^{j\theta_1^{DL,J+(k-1)D}}$ can be simply compensated by the receiver, since the receiver has the full DL channel information.

Type_2 SPF schemes does not only have less overhead, in the next section, it is also proved that if LS is used for channel estimation. Additionally, *Type_2* SPF is also less sensitive to the channel estimation error and has higher average receive signal power per subcarrier (and thus higher receive SNR) than that of *Type_1* SPF.

III. PERFORMANCE ANALYSIS

In this section, we analyze and compare the performance of *Type_1* and *Type_2* SPF scheme under channel estimation error is presented as LS is used in channel estimation. The performance metric used is the average power of the received combined signal of a subcarrier. Before analysis, we first summarize some important settings and assumptions.

1. Due to the property of the LS, it can be assumed that the channel estimation error in different time and different devices (i.e. transmitter or receiver) are independent random variable (may not identically distributed).
2. Because of the circular symmetry of complex Gaussian additive noise, the phase estimation error of each time of channel estimations can be modeled by even probability density functions.
3. Under reasonable noise power of practical interest, we assume the values of phase estimation error are bounded in $[-\pi/4, \pi/4]$ (i.e., $\Pr(|\text{error of estimated phase}| \leq \pi/4) \sim 1$).
4. The wireless channel is quasi-static such that in a short interval containing the time of channel phase feedback and corresponding data transmission, the variation of channel status is relatively slow to be considered as approximately unchanged.

To prove the main result, a Lemma is provided as follows:

Lemma 1: Let $x_1, x_2, \dots, x_n$ ($n \geq 1$) are independent continuous random variable with even probability density function, that is, $f_{x_k}(x) = f_{x_k}(-x)$, for $k=1, 2, \dots, n$. Then, $E(\cos(x_1 + x_2 + \dots + x_n)) = E(\cos(x_1))E(\cos(x_2)) \dots E(\cos(x_n))$.

Proof:

It can be proved by induction.

For $n = 1$, the statement is true. Assume when $n = j-1$ ($j \geq 2$), $E(\cos(x_1 + x_2 + \dots + x_{j-1})) = E(\cos(x_1))E(\cos(x_2)) \dots E(\cos(x_{j-1}))$. Then, for $n = j$,

$$\begin{aligned} & E(\cos(x_1 + x_2 + \dots + x_j)) \\ &= E(\cos(x_1 + x_2 + \dots + x_{j-1}))E(\cos(x_j)) - \\ & E(\sin(x_1 + x_2 + \dots + x_{j-1}))E(\sin(x_j)). \end{aligned} \quad (1)$$

Because x_j has a even probability density function and $\sin(\cdot)$ is an odd function, $E(\sin(x_j)) = 0$. Thus, we can get

$$\begin{aligned} & E(\cos(x_1 + x_2 + \dots + x_j)) \\ &= E(\cos(x_1 + x_2 + \dots + x_{j-1}))E(\cos(x_j)) \\ &= E(\cos(x_1))E(\cos(x_2)) \dots E(\cos(x_j)). \end{aligned} \quad (2)$$

■

By the two Lemmas, it is sufficient to prove that the average received signal power of the *Type_2* SPF scheme is larger than or equal to that of the *Type_1* SPF scheme.

Some notations are first defined as follows:

- $\Delta_1^{DL,n}$: The phase estimation error of the DL channel response from the n -th transmit antenna of a given subcarrier at the receiver side.
- $\Delta_2^{DL,n}$: The phase estimation error of the UL channel response of a given subcarrier from the n -th transmit antenna in the first OFDMA symbol of sounding cluster at the BS side.
- $\Delta_3^{DL,n}$: The phase estimation error of the effective UL channel response of a given subcarrier from the n -th transmit antenna in the second OFDMA symbol of sounding cluster at the BS side.
- d : The modulated signal, without loss of generality, we assume d has average power 1.
- α_n : The amplitude of the DL channel response of a given subcarrier from the n -th transmit antenna

Now we can derive and compare the average power of the received signals of the two schemes.

- a. *Type_1* SPF: The average power the received signal in a given subcarrier at the receiver can be expressed as

$$P_{Type1} = \left| d \sum_{n=1}^M \alpha_n e^{-j(\Delta_1^{DL,n} - \Delta_2^{DL,n} + \Delta_3^{DL,n})} \right|^2, \quad (3)$$

$$= \sum_{n=1}^M \alpha_n^2 + E \left(\underbrace{\sum_{n=2}^M \alpha_n \cos(\Delta_1^{DL,n} - \Delta_2^{DL,n} + \Delta_3^{DL,n} - \Delta_1^{DL,1} + \Delta_2^{DL,1} - \Delta_3^{DL,1})}_{M_1} \right)$$

$$+ E \left(\sum_{n=2}^M \sum_{m=2}^M \alpha_n \alpha_m \cos(\Delta_1^{DL,n} - \Delta_2^{DL,n} + \Delta_3^{DL,n} - \Delta_1^{DL,m} + \Delta_2^{DL,m} - \Delta_3^{DL,m}) \right)$$

- b. *Type_2* SPF: The average power the received signal in a given subcarrier at the receiver can be expressed as

$$P_{Type2} = \left| \left(\alpha_1 + \sum_{n=2}^M \alpha_n e^{-j(\Delta_1^{DL,n} - \Delta_2^{DL,n} + \Delta_3^{DL,n})} \right) d e^{j\theta^{DL,1}} \right|^2, \quad (4)$$

$$= \sum_{n=1}^M \alpha_n^2 + E \left(\underbrace{\alpha_1 \sum_{n=2}^M \alpha_n \cos(\Delta_1^{DL,n} - \Delta_2^{DL,n} + \Delta_3^{DL,n})}_{M_1'} \right)$$

$$+ E \left(\sum_{n=2}^M \sum_{m=2}^M \alpha_n \alpha_m \cos(\Delta_1^{DL,n} - \Delta_2^{DL,n} + \Delta_3^{DL,n} - \Delta_1^{DL,m} + \Delta_2^{DL,m} - \Delta_3^{DL,m}) \right)$$

where $\theta^{DL,1}$ is the phase of DL channel response from the first transmit antenna in a given subcarrier. Also, notice that the response of the n th DL channel, denoted as $h^{DL,n}$, can be expressed as $h^{DL,n} = \alpha_n e^{j\theta^{DL,n}}$. Combining **Lemma1** and the fact that $0 \leq E(\cos(x)) \leq 1$ if x is bound in $[-\pi/4, \pi/4]$, we can conclude that $M_1' \geq M_1$, and consequently, $P_{Type2} \geq P_{Type1}$.

IV. SIMULATION RESULTS

Simulations are conducted to compare the symbol error rate (SER) of the proposed schemes using different estimation methods with that of the optimal EGC in which the perfect channel phase information is known by the transmitter. The main simulation settings are summarized in Table 1.

Our simulation parameter settings are based on the FDD LTE systems with DL and UL bandwidth 10MHz, respectively. In such system, each DL or UL subframe occupies 1ms interval and includes 600 data subcarriers and 14 OFDMA symbols. The channel responses of DL and UL channels are assumed independent. In the simulated scenario, the transmitter is the base station and the receiver is the user. We further assume each device (i.e. base station and user) has a processing delay of three subframes. Thus, the round trip processing delay from the user estimates the DL channel by the DL reference signals to the base station transmits data signals based on the phase information extracted from the proposed sounding channels is six subframes (6ms).

The channel model adopted in the simulations is Extended Pedestrian A (EPA) model [11]. The maximal Doppler frequency is 5Hz which is corresponding to 2.6 km/h user speed under 2.5GHz carrier frequency.

The subband for channel phase feedback contains 72 adjacent subcarriers. The decimation order is set to 8. Thus $9 \times 2 = 18$ sounding signals are needed for each sounding clusters. Assume the base station has two transmit antennas, the number of sounding signals needed for *Type_1* and *Type_2* schemes are 36 and 18, respectively. The user uses LMMSE estimator for DL channel estimation, and the base station uses LS+interpolation or LMMSE to estimate the sounding channels. The performance of the two estimation methods are presented and compared in the simulation results.

TABLE 1

SIMULATION SETTINGS

System bandwidth	10MHz(1024 FFT, 600 data subcarriers)
Duplex mode	FDD
Subband size	72 subcarrier $\times$ 14 OFDMA symbols
Decimation order	8
Round trip processing delay	6 subframe (6ms)
Channel model	ITU-T Extended PedA (EPA)
Max. Doppler frequency	5Hz (2.6 km/h user speed in 2.5GHz band)
Antenna configuration	2 $\times$ 1 and 4 $\times$ 1 MISO
Modulation	QPSK (uncoded)
Estimator in receiver	DL: LMMSE UL: 1. LS+ linear interpolation 2. LMMSE

Fig.3 and Fig.4 are the symbol error rates of QPSK signals for 2 $\times$ 1 and 4 $\times$ 1 MISO channels, respectively. It can be

observed that if LMMSE is used in the transmitter, the SER performance of both *Type_1* and *Type_2* scheme is quite close to achieve the performance of EGC. In 2×1 case, they have almost the same performance. While in 4×1 case, the difference is also within about 0.5 dB. There is a performance loss compared to LMMSE when LS is used. However, it still brings significant improvement over SISO channel. The performance loss of LS is not surprised since the estimation accuracy of LMMSE is typically much better than that of LS.

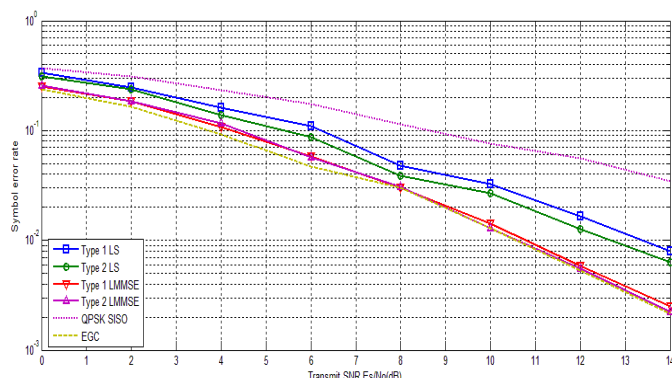


Fig. 3. The symbol error rates of QPSK signals for 2×1 MISO channel

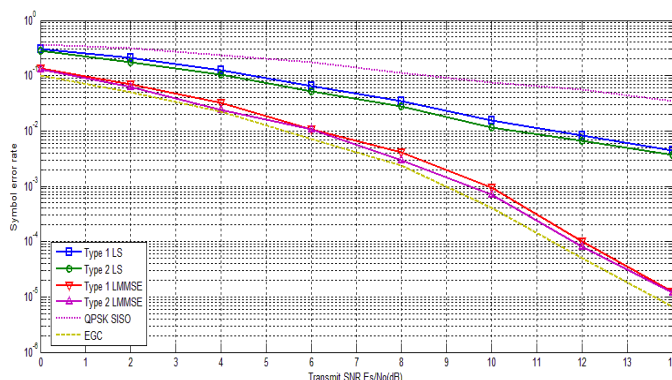


Fig. 4. The symbol error rates of QPSK signal for 4×1 MISO channel

When LS is used, the *Type_2* scheme has obviously better performance than *Type_1* scheme does, as expected according to the analysis in Section III. Furthermore, according to the simulation results, *Type_2* scheme also performs slightly better than *Type_1* scheme when LMMSE is adopted. In the usage scenario in which the channel status does not vary fast, *Type_2* scheme is a better choice with both lower overhead and better performance.

V. CONCLUSIONS

Two sounding-based channel phase feedback schemes (*Type_1* SPF and *Type_2* SPF) for MISO environments in OFDMA systems are successfully proposed. The proposed schemes can convey the channel phase information and allow

the transmitters to perform an approximate EGC transmission. Comparing to current sounding channel feedback scheme, the proposed schemes have the flexibility to be employed in both TDD and FDD modes with comparable overhead. Compare to PMI feedback scheme, the computation complexity of the proposed scheme is lower at the receiver side since the receiver does not need to search the most appropriate precoder matrix.

Simulation results show that when LMMSE method is adopted in the transmitter to estimate the channel responses of sounding channels, the proposed schemes can achieve almost the same performance. Also, the performance is within 0.5 dB compared to that of EGC when the number of transmit antenna are 2 and 4, respectively. In addition, simulation results reveal that in the slow fading environments where the channel status is quasi-static in an interval, the *Type_2* SPF scheme can achieve better performance than that of the *Type_1* SPF scheme while also with lower overhead.

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