A Virtual Successive Detection for Cooperative MU-MIMO Systems with Reduced CSI

Akihito Taya*, Satoshi Denno†, Koji Yamamoto*, Masahiro Morikura*,
Daisuke Umehara‡, Hidekazu Murata*, and Susumu Yoshida*
* Graduate School of Informatics Kyoto University
Yoshida-hommachi, Sakyo-ku, Kyoto, 606–8501 Japan
E–mail: taya@imc.cce.i.kyoto-u.ac.jp
† Graduate School of Natural Science and Technology Okayama University
3–1–1 Tsushima-naka, Kita-ku, Okayama, 700–8530 Japan
‡ Graduate School of Science and Technology, Kyoto Institute of Technology
Matsugasaki, Sakyo-ku, Kyoto, 606–8585 Japan

***Abstract*—In this paper, we propose a receiver using the concept of a "virtual channel" and Turbo codes for multiuser multiple-input multiple-output (MU-MIMO) systems with cooperating transmitters. In order to reduce feedback information from receivers, a precoding method with reduced channel state information (CSI) is needed. Although signals interfere with each other in such a system, the proposed receiver enables high-speed transmission by detecting signals iteratively. In the proposed iterative scheme, soft information is exchanged between virtual channel detectors and Turbo decoders to improve the bit error rate (BER) performance. We evaluate the performance of the proposed receiver in systems with two transmitters, each equipped with four antennas, and two receivers, each equipped with two antennas. The results show that the proposed receiver can improve the system performance owing to the iteration scheme.**

I. Introduction

Since a large number of mobile wireless devices have been developed, a high-speed wireless access network is needed for connecting them. Therefore, multiple-input multiple-output (MIMO) systems, channel capacity of which increases linearly with the number of antennas [1], have attracted great attention. Furthermore, multiuser MIMO (MU-MIMO) systems [2] are expected to improve the throughput of a downlink in wireless communication systems.

Recently, cooperative transmitter systems have been studied for realizing high-data-rate mobile communication [3]. In these systems, multiple cooperating transmitters send signals to multiple receivers, regarding the whole system as a large-scale MU-MIMO system. However, such systems have some problems. In the downlink of multiuser systems, transmitters require channel state information (CSI) in order to avoid inter-receiver interference by block diagonalization (BD) [4]. CSI is usually fed back from the receivers to the transmitters, and the amount of feedback CSI increases rapidly with the number of cooperating transmitters when the transmitters need full CSI. Since the wireless channel resources are limited, it is very important to reduce the amount of feedback CSI [5]. Another problem is that all the transmitters need to share data and send them in the same frame. It becomes more difficult to solve this problem when the number of transmitters increases.

In order to overcome the above-mentioned problems, we proposed a method for reducing the amount of feedback CSI in [6]. When this method is used, receivers feed CSI back to relevant transmitters, and each transmitter uses only its own channel impulse responses. In addition, transmitters can precode signals without sharing the same data and do not need frame synchronization.

In this paper, in order to further reduce the amount of CSI, we propose to feed back only partial CSI, i.e., each receiver selects one of the transmitters and feeds back CSI that is relevant to the selected transmitter. When the proposed method is used, the transmitters know less CSI than when the method in [6] is used, and limited CSI at transmitters (CSIT) causes signal interference. We propose a novel receiver using the concept of a "virtual channel" in order to overcome this interference. A "virtual channel," which divides signals into real-signal elements and phase elements, has been developed for single-user MIMO (SU-MIMO) systems so that signals can be detected even when the number of their streams exceeds the number of receiving antennas [7]. In addition, we propose an iteration scheme for improving the system performance. For this scheme, Turbo cods [8] are used, although convolutional codes are conventionally used in [6]. In the iterative scheme, virtual channel detectors estimate signals by using the log likelihood ratio (LLR) of coded bit streams calculated at Turbo decoders, and output the LLR of transmitted signals. The iterative detection using LLR is referred to as soft-input soft-output (SISO) detection.

II. System Model

A. MU-MIMO with Cooperating Transmitters

Figure 1 shows a system with M_{T} transmitters and M_{R} receivers. We consider the downlink of the system. Each transmitter has n_{T} antennas, and each receiver has n_{R} antennas. The sum of the transmission and receiving antennas is represented as $N_{\mathrm{T}} = M_{\mathrm{T}} n_{\mathrm{T}}$ and $N_{\mathrm{R}} = M_{\mathrm{R}} n_{\mathrm{R}}$, respectively, and the sum of the transmitted streams to one of the receiver is represented as n_{S}.

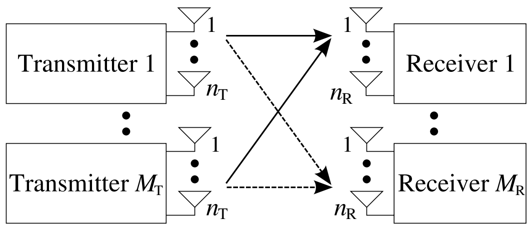

Fig. 1. System model

At the transmitters, data bit streams are encoded using a Turbo code, interleaved, and modulated to quadrature phase-shift keying (QPSK) signals. The signals are then precoded by BD and are orthogonal frequency division multiplexing (OFDM)-modulated by inverse fast Fourier transform (IFFT). Finally, the transmitters send the OFDM signals to the receivers. The receivers demodulate these OFDM signals by FFT, and detect them using a virtual channel detector. The detected signals are then QPSK-demodulated and inputted into de-interleavers. Finally, the coded bit streams are decoded by Turbo decoders.

Let $\boldsymbol{Y}_i \in \mathbb{C}^{n_\mathrm{R}}$ represent a received signal vector at receiver i; $\boldsymbol{X}_j \in \mathbb{C}^{n_\mathrm{T}}$, a transmitted signal vector from transmitter j; $\boldsymbol{H}_{ij} \in \mathbb{C}^{n_\mathrm{R}\times n_\mathrm{T}}$, a channel matrix from transmitter j to receiver i; and $\boldsymbol{N}_i \in \mathbb{C}^{n_\mathrm{R}}$, an additive white Gaussian noise (AWGN) vector. The above signals are the subcarrier signals. Here, the received signal vector at receiver i is written as

$$\boldsymbol{Y}_i = \sum_{j=1}^{M_\mathrm{T}} \boldsymbol{H}_{ij}\boldsymbol{X}_j + \boldsymbol{N}_i. \tag{1}$$

B. BD Algorithm

Let us define $\boldsymbol{H}_i$ and $\boldsymbol{H}^{(i)}$ as $\boldsymbol{H}_i = \begin{bmatrix}\boldsymbol{H}_{i1} & \cdots & \boldsymbol{H}_{iM_\mathrm{T}}\end{bmatrix}$ and $\boldsymbol{H}^{(i)} = \begin{bmatrix}\boldsymbol{H}_1^\mathrm{T} & \cdots & \boldsymbol{H}_{(i-1)}^\mathrm{T} \; \boldsymbol{H}_{(i+1)}^\mathrm{T} & \cdots & \boldsymbol{H}_{M_\mathrm{R}}^\mathrm{T}\end{bmatrix}^\mathrm{T}$, where $[\cdot]^\mathrm{T}$ is the transpose of $[\cdot]$. In the BD algorithm, data signals are precoded to transmission signals. Precoding matrix $\tilde{\boldsymbol{V}}_i \in \mathbb{C}^{n_\mathrm{T}\times n_\mathrm{S}}$ lying in the null space of $\boldsymbol{H}^{(i)}$ can be obtained using singular value decomposition (SVD), where $n_\mathrm{S} = N_\mathrm{T} - N_\mathrm{R} + n_\mathrm{R}$. Data signal vector $\boldsymbol{D}_i \in \mathbb{C}^{n_\mathrm{S}}$, which is sent to receiver i, is precoded to $\boldsymbol{X}$ by $\tilde{\boldsymbol{V}}_i$ as $\boldsymbol{X} = \begin{bmatrix}\tilde{\boldsymbol{V}}_1 & \cdots & \tilde{\boldsymbol{V}}_{M_\mathrm{R}}\end{bmatrix}\begin{bmatrix}\boldsymbol{D}_1^\mathrm{T} & \cdots & \boldsymbol{D}_{M_\mathrm{R}}^\mathrm{T}\end{bmatrix}^\mathrm{T}$, and the transmission signals from each transmitter are defined to satisfy $\boldsymbol{X} = \begin{bmatrix}\boldsymbol{X}_1^\mathrm{T} & \cdots & \boldsymbol{X}_{M_\mathrm{T}}^\mathrm{T}\end{bmatrix}^\mathrm{T}$. Since $\tilde{\boldsymbol{V}}_i$ lies in the null space of $\boldsymbol{H}^{(i)}$, $\boldsymbol{H}_i\tilde{\boldsymbol{V}}_{i'} = \boldsymbol{O}$ for $i \neq i'$, where $\boldsymbol{O}$ is a zero matrix. Therefore, $\boldsymbol{Y}_i$ is rewritten as

$$\boldsymbol{Y}_i = \boldsymbol{H}_i\tilde{\boldsymbol{V}}_i\boldsymbol{D}_i + \boldsymbol{N}_i. \tag{2}$$

Conventionally, each transmitter requires CSI of the entire system, denoted by $\boldsymbol{H}_{ij}$ for all $i = 1, \cdots, M_\mathrm{R}$ and $j = 1, \cdots, M_\mathrm{T}$. Therefore, receiver i feeds back $\boldsymbol{H}_{i1}, \cdots, \boldsymbol{H}_{iM_\mathrm{T}}$ to all the transmitters. The amount of such feedback information is equal to $2QLM_\mathrm{T}^2M_\mathrm{R}n_\mathrm{T}n_\mathrm{R}$ for L-path Rayleigh fading channels, where Q represents a quantization bit number per real element of the channel matrix. This amount increases rapidly with the number of transmitters. Therefore, a method for reducing CSI is needed.

There is another problem in conventional BD. For detecting data signal $\boldsymbol{D}_i$, receiver i requires transmission signals $\boldsymbol{X}_1, \cdots, \boldsymbol{X}_{M_\mathrm{T}}$. Therefore, all the transmitters need to share same signal $\boldsymbol{D}_i$ and send its precoded signal $\boldsymbol{X}_j$ in the same frame. This synchronization is considered to be very difficult to achieve. In order to overcome these problems, we describe a precoding method for reducing feedback CSI. When this method is used, transmitters can precode signals independently and do not need synchronization.

III. Proposed Method for Reducing CSI

First, each receiver selects one of the transmitters. If receiver i selects transmitter j, CSI denoted by $\boldsymbol{H}_{ij}$ is fed back from receiver i to transmitter j. The amount of such feedback information is reduced to $2QLM_\mathrm{R}n_\mathrm{T}n_\mathrm{R}$, which is less than when the conventional method is used by $1/M_\mathrm{T}^2$.

At transmitters, signals are precoded using limited CSIT. In this paper, we restrict ourselves to a system with two transmitters and two receivers. If both receivers select same transmitter 1, transmitter 1 knows CSI $\boldsymbol{H}_{11}$ and $\boldsymbol{H}_{21}$ that represent the channel impulse responses from transmitter 1 to receiver 1 and receiver 2, respectively. Therefore, transmitter 1 can precode signals by BD, as in single-transmitter MU-MIMO systems; $\boldsymbol{X}_1 = \begin{bmatrix}\tilde{\boldsymbol{V}}_{21} & \tilde{\boldsymbol{V}}_{11}\end{bmatrix}\begin{bmatrix}\boldsymbol{D}_{11}^\mathrm{T} & \boldsymbol{D}_{21}^\mathrm{T}\end{bmatrix}^\mathrm{T}$, where $\tilde{\boldsymbol{V}}_{ij} \in \mathbb{C}^{n_\mathrm{T}\times(n_\mathrm{T}-n_\mathrm{R})}$ is defined to lie in the null space of $\boldsymbol{H}_{kj}$, and $\boldsymbol{D}_{ij} \in \mathbb{C}^{n_\mathrm{T}-n_\mathrm{R}}$ represents a data signal vector from transmitter j to receiver i. On the other hand, at transmitter 2, which has no CSI, data signals are transmitted without precoding as $\boldsymbol{X}_2 = \begin{bmatrix}\boldsymbol{D}_{12}^\mathrm{T} & \boldsymbol{D}_{22}^\mathrm{T}\end{bmatrix}^\mathrm{T}$. Here, the dimension of $\boldsymbol{D}_{ij}$ is equal to $n_\mathrm{T}/2$. In this situation, received signal vector $\boldsymbol{Y}_i$ is expressed as

$$\boldsymbol{Y}_i = \begin{bmatrix}\boldsymbol{H}_{i1}\tilde{\boldsymbol{V}}_{k1} & \boldsymbol{H}_{i2}\end{bmatrix}\begin{bmatrix}\boldsymbol{D}_{i1}^\mathrm{T} & \boldsymbol{D}_{i2}^\mathrm{T} & \boldsymbol{D}_{k2}^\mathrm{T}\end{bmatrix}^\mathrm{T} + \boldsymbol{N}_i. \tag{3}$$

In the above equation, $k = 2$ if $i = 1$ and $k = 1$ if $i = 2$.

In the other situation in which the receivers select different transmitters, both transmitters can use partial CSI. Let us assume that receivers 1 and 2 select transmitters 1 and 2, respectively. At transmitter 1, the null space of the channel for receiver 1 can be calculated but the null space of the channel for receiver 2 cannot be calculated because as CSI, only $\boldsymbol{H}_{11}$ is available. Therefore, data signal vector $\boldsymbol{D}_{21}$ can be precoded with $\tilde{\boldsymbol{V}}_{11}$ for not being received at receiver 1, but $\boldsymbol{D}_{11}$ cannot be avoided from being received at receiver 2. In this case, we propose to precode $\boldsymbol{D}_{11}$ with $\boldsymbol{V}_{11} \in \mathbb{C}^{n_\mathrm{T}\times n_\mathrm{R}}$, which is right singular matrix of $\boldsymbol{H}_{11}$. The transmission signal vector is calculated as $\boldsymbol{X}_1 = \begin{bmatrix}\boldsymbol{V}_{11} & \tilde{\boldsymbol{V}}_{11}\end{bmatrix}\begin{bmatrix}\boldsymbol{D}_{11}^\mathrm{T} & \boldsymbol{D}_{21}^\mathrm{T}\end{bmatrix}^\mathrm{T}$ and the received signal vector is written as

$$\begin{aligned}\boldsymbol{Y}_i = &\begin{bmatrix}\boldsymbol{H}_{ii}\boldsymbol{V}_{ii} & \boldsymbol{H}_{ik}\boldsymbol{V}_{kk} & \boldsymbol{H}_{ik}\tilde{\boldsymbol{V}}_{kk}\end{bmatrix}\begin{bmatrix}\boldsymbol{D}_{ii}^\mathrm{T} & \boldsymbol{D}_{kk}^\mathrm{T} & \boldsymbol{D}_{ik}^\mathrm{T}\end{bmatrix}^\mathrm{T} \\ &+ \boldsymbol{N}_i.\end{aligned} \tag{4}$$

In (3) and (4), receiver i receives signals $\boldsymbol{D}_{k1}$ or $\boldsymbol{D}_{k2}$, which are intended for receiver k. Receiver i must detect these interfered signals in addition to the signals intended for

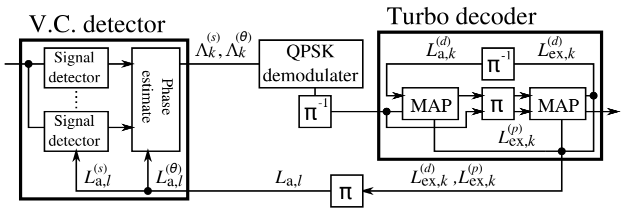

Fig. 2. Receiver architecture

receiver i. In order to solve this problem, we propose a novel receiver using the concept of a "virtual channel" and SISO iterations.

IV. Architecture of Proposed Receiver with Virtual Channels

Figure 2 shows the architecture of the proposed receiver. First, virtual channel detectors estimate data signal vectors from received signal vectors. The estimated signals are QPSK-demodulated and de-interleaved into coded bit streams. Turbo decoders then decode the bit streams iteratively by exchanging LLRs of signals between two maximum *a posteriori* probability (MAP) decoders. After N_{Turbo} iterations, the extrinsic LLRs of coded bit streams are interleaved and inputted into virtual channel detectors as *a priori* LLRs. Using this information, the virtual channel detectors estimate data signal vectors and output soft QPSK signals. The Turbo decoders then again decode the inputted bit streams and output extrinsic LLRs into the virtual channel detectors. The number of iterations of LLR exchange between the virtual channel detectors and Turbo decoders is represented as N_{SISO}. Here, the Turbo decoders decode signals a total of $N_{\mathrm{SISO}} \times N_{\mathrm{Turbo}}$ times.

A. Virtual Channels

The concept of a virtual channel has been developed for SU-MIMO [7], but it can be applied to the considered system. When using the concept of a virtual channel, complex signal vectors and channel matrices are transformed into real vectors and matrices, respectively. Let $\hat{\boldsymbol{Y}}_i$, $\hat{\boldsymbol{D}}_i$, $\hat{\boldsymbol{N}}_i$, and $\hat{\boldsymbol{H}}_i$ represent a received signal vector, a data signal vector, an AWGN vector, and a channel matrix expressed in real value, respectively. The data signal vector includes interfered signals, and the channel matrix represents an equivalent channel including precoding matrices. The lth QPSK-modulated symbol $d_{i,l}$ is divided into real-signal element $s_{i,l}$ and phase element $\theta_{i,l}$ as $d_{i,l} = s_{i,l}\mathrm{e}^{\mathrm{j}\theta_{i,l}}$, where $s_{i,l} = \pm\sqrt{2}$ and $\theta_{i,l} = \pm\pi/4$; j represents the imaginary unit. When the division of a scalar is extended to a vector, $\hat{\boldsymbol{D}}_i$ is rewritten as follows:

$$\hat{\boldsymbol{D}}_i = \boldsymbol{\Omega}_i(\varphi_i)\boldsymbol{S}_i \tag{5}$$

$$\boldsymbol{\Omega}_i(\varphi_i) = \begin{bmatrix} \mathrm{diag}\left[\cos\theta_{i,1}, \cdots, \cos\theta_{i,n_{\mathrm{S}}}\right] \\ \mathrm{diag}\left[\sin\theta_{i,1}, \cdots, \sin\theta_{i,n_{\mathrm{S}}}\right] \end{bmatrix} \tag{6}$$

$$\boldsymbol{S}_i = \begin{bmatrix} s_{i,1} & \cdots & s_{i,n_{\mathrm{S}}} \end{bmatrix}^{\mathrm{T}}, \tag{7}$$

where we call $\boldsymbol{\Omega}_i(\phi)$ a rotation matrix and $\boldsymbol{S}_i$ a real-signal vector. φ_i is the index of the pattern of phase elements $\theta_{i,1}, \cdots, \theta_{i,n_{\mathrm{S}}}$, which is referred to as a "phase pattern" and can have $2^{n_{\mathrm{S}}}$ values. From (5), we can rewrite (3) and (4) as

$$\hat{\boldsymbol{Y}}_i = \boldsymbol{\Phi}_i(\varphi_i)\boldsymbol{S}_i + \hat{\boldsymbol{N}}_i. \tag{8}$$

Here, we refer $\boldsymbol{\Phi}_i(\varphi_i) \left(= \hat{\boldsymbol{H}}_i\boldsymbol{\Omega}_i(\varphi_i)\right)$ as a "virtual channel."

In (8), $\hat{\boldsymbol{Y}}_i$ is a $2n_{\mathrm{R}}$-D vector and $\boldsymbol{S}_i$ is an n_{S}-D vector, i.e., the receivers function as if they have twice as many antennas as they actually do. Therefore, regarding $\boldsymbol{S}_i$ as a transmitted signal over virtual channel $\boldsymbol{\Phi}_i(\varphi_i)$, the receivers can detect $\boldsymbol{S}_i$ for each phase pattern φ by ordered successive detection (OSD), even when $n_{\mathrm{S}} > n_{\mathrm{R}}$. In OSD, signals are alternately detected with a minimum mean-square error (MMSE) filter, and the interference among the signals is canceled in the decreasing order of the signal-to-noise ratio (SNR). The nth-step MMSE filter $\boldsymbol{W}_i^{(n)}(\varphi)$ is defined as

$$\boldsymbol{W}_i^{(n)}(\varphi) = \left(\boldsymbol{\Phi}_i^{(n)}(\varphi)\left[\boldsymbol{\Phi}_i^{(n)}(\varphi)\right]^{\mathrm{T}} + \frac{\sigma_n^2}{\sigma_s^2}\boldsymbol{I}\right)^{-1}\boldsymbol{\Phi}_i^{(n)}(\varphi), \tag{9}$$

where $\boldsymbol{\Phi}_i^{(n)}(\varphi)$, σ_n^2, σ_s^2, and $\boldsymbol{I}$ represent the virtual channel in the nth step, noise power, signal power, and the unity matrix, respectively. The lth-stream signal is estimated as $s_{\mathrm{det},i,l} = \boldsymbol{w}_{i,l}^{\mathrm{T}}\hat{\boldsymbol{Y}}_i$, where $\boldsymbol{w}_{i,l}$ represents the lth column of $\boldsymbol{W}_i$. The soft signal vector $\boldsymbol{S}_{\mathrm{det},i}(\varphi) = \left[s_{\mathrm{det},i,1} \cdots s_{\mathrm{det},i,n_{\mathrm{S}}}\right]^{\mathrm{T}}$ is hard decided into $\overline{\boldsymbol{S}}_{\mathrm{det},i}$, and the phase pattern is then estimated as $\varphi_{\mathrm{det},i} = \underset{\varphi}{\mathrm{argmin}}\left|\hat{\boldsymbol{Y}}_i - \boldsymbol{\Phi}(_i\varphi)\overline{\boldsymbol{S}}_{\mathrm{det},i}(\varphi)\right|^2$. Finally, the estimated data signal vector is expressed as $\boldsymbol{D}_{\mathrm{det},i} = \boldsymbol{\Omega}_i(\varphi_{\mathrm{det},i})\boldsymbol{S}_{\mathrm{det},i}(\varphi_{\mathrm{det},i})$. For the first time in SISO detection, the filter-output signals $\boldsymbol{D}_{\mathrm{det},i}$ are inputted into the Turbo decoders.

B. LLRs of Data and Parity Bits

Let $\boldsymbol{c}_k = (d_k, p_k)$ represent the kth data and parity bits; $\boldsymbol{r}_k = (r_{dk}, r_{pk})$, the received data and parity bits corresponding to $\boldsymbol{c}_k$; $\boldsymbol{R} = (\boldsymbol{r}_1, \cdots, \boldsymbol{r}_K)$, the inputted bit stream to one of the MAP decoders. At the Turbo decoders, LLR $\Lambda_k^{(d)}$ of data bit d_k is calculated with Max-Log-MAP. Extrinsic LLR $L_{\mathrm{ex},k}^{(d)}$ is then obtained as $L_{\mathrm{ex},k}^{(d)} = \Lambda_k^{(d)} - L_{\mathrm{a},k}^{(d)} - L_{\mathrm{in},k}^{(d)}$, where $L_{\mathrm{a},k}^{(d)} = \log\frac{P(d_k=1)}{P(d_k=0)}$ and $L_{\mathrm{in},k}^{(d)} = 2r_{dk}/\sigma_n^2$ represent the *a priori* LLR and intrinsic LLR, respectively. The extrinsic LLR is outputted to the other MAP decoder as an *a priori* LLR. In addition to the LLRs of the data bits, the proposed receivers use the *a priori* LLRs of coded bits c_i when they calculate the LLRs of real-signal and phase elements. Therefore, the LLRs of the parity bits are required. The kth parity bit's LLR $\Lambda_k^{(p)}$ and extrinsic LLR $L_{\mathrm{ex},k}^{(p)}$ are obtained as follows:

$$\begin{aligned} \Lambda_k^{(p)} &= \log\frac{P(p_k=1|\boldsymbol{R})}{P(p_k=0|\boldsymbol{R})} \\ &\simeq L_{\mathrm{a},k}^{(p)} - L_{\mathrm{a},k}^{(d)} + \max_{p_k=1:l\to l'}\left[\log\alpha_{k-1}^l\gamma_k^{l\xrightarrow{d_k=1}l'}\beta_k^{l'}\right] \\ &\quad - \max_{p_k=0:l\to l'}\left[\log\alpha_{k-1}^l\gamma_k^{l\xrightarrow{d_k=0}l'}\beta_k^{l'}\right] \end{aligned} \tag{10}$$

$$L_{\mathrm{ex},k}^{(p)} = \Lambda_k^{(p)} - L_{\mathrm{a},k}^{(p)} - L_{\mathrm{in},k}^{(p)}, \tag{11}$$

where α_{k-1}^{l}, $\beta_k^{l'}$, and $\gamma_k^{l \xrightarrow{d_k} l'}$ are defined as $\gamma_k^{l \xrightarrow{d_k} l'} = P(d_k)P(\boldsymbol{r}_k|\boldsymbol{c}_k)$, $\alpha_{k-1}^{l} = \sum_{l''} \alpha_{k-2}^{l''}\gamma_{k-1}^{l''\to l}$, and $\beta_k^{l'} = \sum_{l''} \beta_{k+1}^{l''}\gamma_{k+1}^{l'\to l''}$, respectively. $L_{\mathrm{a},k}^{(p)} = \log \frac{P(p_k=1)}{P(p_k=0)}$ and $L_{\mathrm{in},k}^{(p)} = 2r_{pk}/\sigma_n^2$ represent the *a priori* LLR and intrinsic LLR of the kth parity bit, respectively. After N_{Turbo} iterations, the extrinsic LLRs of the data and parity bits are interleaved and inputted into the virtual channel detectors as *a priori* LLRs.

C. LLRs of Real-signal and Phase Elements

At the virtual channel detectors, the *a priori* LLRs of real-signal and phase elements are calculated from the inputted LLRs of the coded bit streams denoted as $L_{\mathrm{a},l} = \log \frac{P(c_l=1)}{P(c_l=0)}$. Since c_{2l} becomes the real part and c_{2l+1} becomes the imaginary part when the coded bit streams are QPSK-modulated, the *a priori* LLRs of the real-signal and phase elements are expressed by the equations given below.

$$L_{\mathrm{a},l}^{(s)} = \log \frac{P\left(s_l = +\sqrt{2}\right)}{P\left(s_l = -\sqrt{2}\right)} = L_{\mathrm{a},2l} \tag{12}$$

$$\begin{aligned} L_{\mathrm{a},l}^{(\theta)} &= \log \frac{P\left(\theta_l = +\pi/4\right)}{P\left(\theta_l = -\pi/4\right)} \\ &\simeq \max\left[L_{\mathrm{a},2l} + L_{\mathrm{a},2l+1}, 0\right] - \max\left[L_{\mathrm{a},2l}, L_{\mathrm{a},2l+1}\right] \end{aligned} \tag{13}$$

Using these *a priori* LLRs, LLR $\Lambda_l^{(s)}$ of the lth real-signal element can be obtained as

$$\Lambda_l^{(s)} = \log \frac{P(s_l = +\sqrt{2}|\hat{\boldsymbol{Y}})}{P(s_l = -\sqrt{2}|\hat{\boldsymbol{Y}})} \simeq L_{\mathrm{a},l}^{(s)} + \frac{2\sqrt{2}s_{\mathrm{det},i,l}}{\sigma_n^2|\boldsymbol{w}_{i,l}|^2}. \tag{14}$$

After the LLRs of the real-signal elements are obtained, LLR $\Lambda_l^{(\theta)}$ of the lth phase element is calculated as

$$\begin{aligned} \Lambda_l^{(\theta)} =& \log \frac{P(\theta_l = +\pi/4|\hat{\boldsymbol{Y}})}{P(\theta_l = -\pi/4|\hat{\boldsymbol{Y}})} \\ =& L_{\mathrm{a},l}^{(\theta)} + \max_{\theta_l=+\frac{\pi}{4}} \left[\sum_{\theta_m=+\frac{\pi}{4}, m\neq l}^{n_\mathrm{S}} L_{\mathrm{a},m}^{(\theta)} + J(\varphi)\right] \\ & - \max_{\theta_l=-\frac{\pi}{4}} \left[\sum_{\theta_m=+\frac{\pi}{4}, m\neq l}^{n_\mathrm{S}} L_{\mathrm{a},m}^{(\theta)} + J(\varphi)\right], \end{aligned} \tag{15}$$

where $J(\varphi) = -\frac{1}{2\sigma_n^2}\left|\hat{\boldsymbol{Y}}_i - \boldsymbol{\Phi}_i(\varphi)\overline{\boldsymbol{S}}_i(\varphi)\right|^2$ and $\overline{\boldsymbol{S}}_i(\varphi)$ is a hard-decided signal vector obtained by $\Lambda_l^{(s)}$. The extrinsic LLRs of the phase elements are defined as $\Lambda_{\mathrm{ex},l}^{(\theta)} = \Lambda_l^{(\theta)} - L_{\mathrm{a},l}^{(\theta)}$. The LLRs of the real-signal and phase elements are converted to the LLRs of the QPSK symbols.

$$\Lambda_l^{(\mathrm{r})} = \log \frac{P\left(c_l^{\mathrm{r}} = +1\right)}{P\left(c_l^{\mathrm{r}} = -1\right)} = \Lambda_l^{(s)} \tag{16}$$

$$\begin{aligned} \Lambda_l^{(\mathrm{i})} &= \log \frac{P\left(c_l^{\mathrm{i}} = +1\right)}{P\left(c_l^{\mathrm{i}} = -1\right)} \\ &\simeq \max\left[\Lambda_l^{(s)}, -\Lambda_{\mathrm{ex},l}^{(\theta)}\right] - \max\left[\Lambda_l^{(s)} - \Lambda_{\mathrm{ex},l}^{(\theta)}, 0\right] \end{aligned} \tag{17}$$

The output soft signal $d_{\mathrm{det},i,l}$ is calculated as $\mathrm{Re}\left[d_{\mathrm{det},i,l}\right] = \frac{\mathrm{e}^{\Lambda_l^{(\mathrm{r})}}-1}{\mathrm{e}^{\Lambda_l^{(\mathrm{r})}}+1}$, $\mathrm{Im}\left[d_{\mathrm{det},i,l}\right] = \frac{\mathrm{e}^{\Lambda_l^{(\mathrm{i})}}-1}{\mathrm{e}^{\Lambda_l^{(\mathrm{i})}}+1}$. These signals are inputted into the Turbo decoders through de-interleavers and are then decoded again. The iterative estimation between the virtual channel detectors and Turbo decoders are performed N_{SISO} times.

TABLE I
SIMULATION PARAMETERS

number of antennas	$\{4,4\} \times \{2,2\}$
modulation	OFDM-QPSK
channel model	4-path Rayleigh fading
coding	Turbo code, rate $1/3$
cyclic prefix	10
sub carrier	512
signal length	10

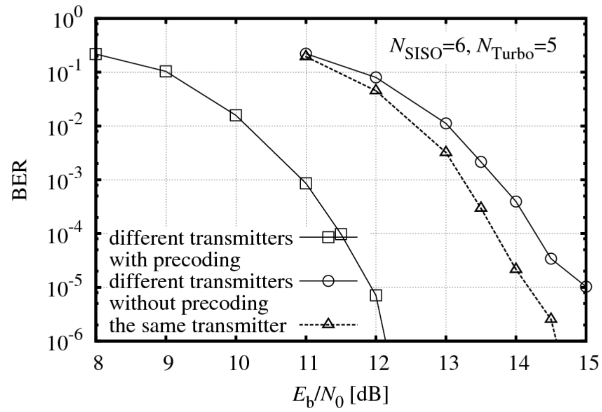

Fig. 3. Performance of Proposed Precoding Method

V. SIMULATION RESULTS

In this section, we evaluate the bit error rate (BER) performance of the proposed method by computer simulation. The simulation parameters are listed in Table I. The simulated system has two transmitters with four antennas each and two receivers with two antennas each. In the system, each receiver detects six-stream signals. The channel impulse responses are independently and identically distributed (i.i.d) and are quasistatic, and the receivers can estimate CSI perfectly and feed them back to the transmitters with no error.

A. Performance of Proposed Precoding Method

The BER performances obtained with different precoding methods are shown in Fig. 3. The solid curves show the performance when the receivers select different transmitters. One is the performance of the proposed precoding method with right singular matrices, and the other is the method without precoding, i.e., data signals are calculated using $\frac{1}{\sqrt{2}}\left[\boldsymbol{I}\ \boldsymbol{I}\right]^{\mathrm{T}}$ instead of right singular matrices. The results show that the proposed precoding method can improve the BER by about 2.8 dB in E_b/N_0 when the receivers select different transmitters. Moreover, the performance of this method is better than the performance when the receivers select the same transmitter.

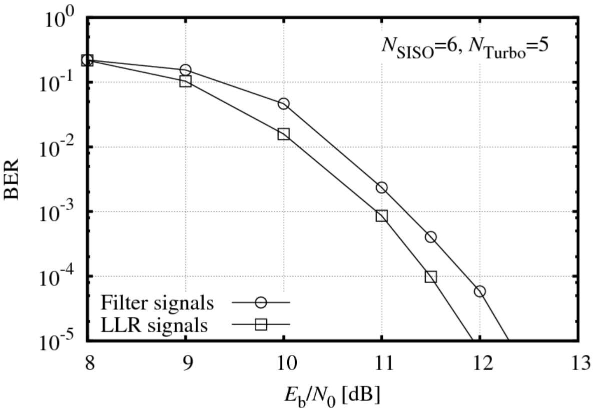

Fig. 4. Output Signals of Virtual Channel Detectors

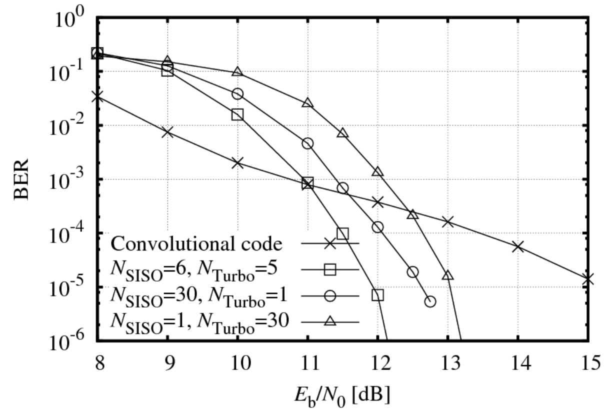

Fig. 6. Performance with Various Number of Iterations

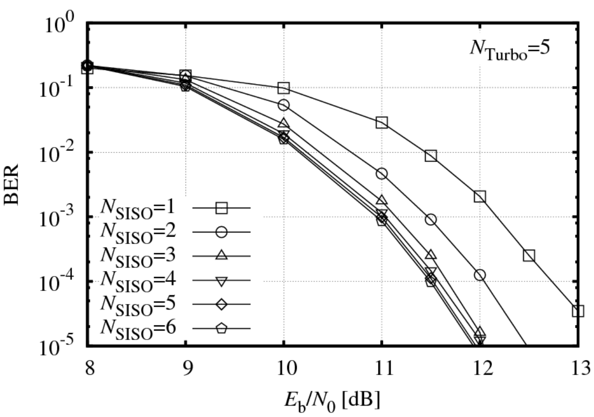

Fig. 5. Iteration Performance

B. Performance of Proposed Receivers

In Fig. 4, the performance of SISO detection with LLR signals $\Lambda_i^{(r)}$ and $\Lambda_i^{(i)}$, and filter-output signals $\boldsymbol{D}_{\text{det},i}$ as the output of the virtual channel detectors are compared. The method with filter-output signals is used in [7]. Using LLR signals as the output of the virtual channel detectors show better performance than using filter-output signals.

Figure 5 shows the BER performance of the proposed iterative scheme for the system in which the receivers select different transmitters and the transmitters precode signals with singular vectors. As shown in Fig. 5, the proposed receiver can improve the BER performance by iterative detection.

Figure 6 shows the performance with different number of iterations and the performance with convolutional code as a conventional method. $N_{\text{SISO}} = 1$ implies that the Turbo decoders do not feed back the LLRs to the virtual channel detectors. On the other hand, when $N_{\text{Turbo}} = 1$, two MAP decoders decode signals only once and feed the LLRs back to the virtual channel detectors. Since the system with $N_{\text{SISO}} = 6$ and $N_{\text{Turbo}} = 5$ shows the best performance, it can be claimed that a double loop architecture improves the system throughput.

VI. Conclusions

We propose a virtual channel receiver with SISO detection for the cooperative MU-MIMO systems with reduced feedback CSI. The proposed receivers can detect signals even when interfered signals are received and the number of signal streams exceeds the number of receiving antennas, because they implement iterative detection between virtual channel detectors and Turbo decoders. We evaluate the performance of a precoding method and the proposed architecture. The results show that feeding back CSI to different transmitters and using right singular matrices for precoding increases BER performance. Furthermore, the proposed iterative scheme improves system throughput.

Acknowledgment

This research is supported by Strategic Information and Communication R&D Promotion Program (SCOPE) conducted by the Ministry of Internal Affairs and Communication (MIC) of Japan.

References

[1] G. J. Foschini and M. J. Gans, “On limits of wireless communications in a fading environment when using multiple antennas,” *Wireless Pers. Commun.*, vol. 6, no. 3, pp. 311–335, March 1998.

[2] Q. H. Spencer, C. B. Peel, A. L. Swindlehurst, and M. Haardt, “An introduction to the multi-user MIMO downlink,” *IEEE Commun. Mag.*, vol. 42, no. 10, pp. 60–67, October 2004.

[3] G. J. Foschini, K. Karakayali, and R. A. Valenzuela, “Coordinating multiple antenna cellular networks to achieve enormous spectral efficiency,” *Commun., IEE Proc.*, vol. 153, no. 4, pp. 548–555, August 2006.

[4] Q. H. Spencer, A. L. Swindlehurst, and M. Haardt, “Zero-forcing methods for downlink spatial multiplexing in multiuser MIMO channels,” *IEEE Trans. Signal Process.*, vol. 52, no. 2, pp. 461–471, February 2004.

[5] J. C. Roh and B. D. Rao, “Transmit beamforming in multiple-antenna systems with finite rate feedback: a vq-based approach,” *Information Theory, IEEE Trans. on*, vol. 52, no. 3, pp. 1101–1112, March 2006.

[6] A. Taya, S. Denno, D. Umehara, M. Morikura, H. Murata, K. Yamamoto, and S. Yoshida, “A virtual layered space-frequency receiver for multiuser MIMO-OFDM systems,” in *VTC Fall’11*, September 2011.

[7] J. Imamura, S. Denno, D. Umehara, and M. Morikura, “A virtual layered space-frequency receiver architecture with iterative decoding,” in *IEICE Trans. Commun.*, vol. E94-B, no. 7, July 2011, pp. 1994–2002.

[8] C. Berrou and A. Glavieux, “Near optimum error correcting coding and decoding: Turbo-codes,” *IEEE Trans. Commun.*, vol. 44, no. 10, pp. 1261–1271, October 1996.