Co-channel Interference Cancellation for Downlink Block Transmission with Cyclic Prefix

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***Abstract*— This paper proposes co-channel interference cancellation / equalization methods for downlink block transmission with cyclic prefix in multi-cell environments using multiple receiving antennas and iterative least-square (LS) interference cancellation. Due to circulant structure of channel matrices in block transmission schemes with cyclic prefix, pseudo inverse matrices, which is required in the iterative LS method, can be obtained with low computational complexity and is multiplied to the received signal in an efficient manner. Computer simulation results show that the proposed iterative LS interference cancellation and equalization method can achieve almost the same bit error rate (BER) performance as an minimum mean-square-error (MMSE) approach with lower computational complexity and without noise variance estimation.**

I. Introduction

Block transmission schemes with cyclic prefix (CP), such as orthogonal frequency division multiplexing (OFDM)[1] and single-carrier block transmission with CP (SC-CP)[2]-[4], have been applied to mobile communications systems, such as WiMAX system, where orthogonal frequency division multiple access (OFDMA)[5] is adopted for the physical layer /medium access control layer protocol. One of the most significant problems in such mobile communications environments is the interference from neighboring cells, especially for the case with frequency reuse factor of 1. In order to cope with the problem, we have proposed a downlink channel estimation scheme for multi-cell block transmission systems[6], where not only a channel response between a desired base station (BS) and a mobile terminal (MT) but also responses between interfering BSs and the MT are obtained simultaneously, by taking advantage of the signal structure of the block transmission with cyclic prefix.

In this paper, assuming the availability of the channel state information (CSI) of the desired and the interference signals, we propose co-channel interference cancellation / equalization methods for downlink block transmission with cyclic prefix using multiple receiving antennas and iterative least square (LS) interference cancellation, which is also employed in [6]. Due to the circulant structure of the channel matrices in the block transmission systems with cyclic prefix, the pseudo inverse matrices, which is required in the iterative LS method, can be obtained with low computational complexity and is multiplied to the received signal vector in an efficient manner. Computer simulation results show that the proposed iterative LS interference cancellation and equalization method can achieve almost the same bit error rate (BER) performance as an minimum mean-square-error (MMSE) approach with lower computational complexity and without noise variance estimation.

II. Signal Modeling

The following notations are used for describing the proposed system. K is the length of the guard interval, M is the FFT size, and L is the channel order. An $M \times M$ identity matrix will be denoted as $\mathbf{I}_M$, and a discrete Fourier transform (DFT) matrix of size $M \times M$, whose (i,j) element is $\frac{1}{\sqrt{M}} e^{-j\frac{2\pi(i-1)(j-1)}{M}}$, as $\mathbf{D}$. We will use $E[\cdot]$ to denote ensemble average, $(\cdot)^T$ for transpose, $(\cdot)^H$ for Hermitian transpose, diag$[\cdot]$ for diagonal matrix, and $(\cdot)^*$ for complex conjugate.

A. OFDMA

Fig.1 shows a system model considered in this paper, where signals from one desired base station and U interfering base stations are received at a mobile terminal, which has Q antenna elements. Assuming that the DFT window timings of all the base stations are synchronized, and that the length of the cyclic prefix K is greater than or equal to channel order L, the received signal vector at the q-th antenna after the cyclic prefix removal can be written as

$$\mathbf{r}_{of}^{(q)} = \mathbf{C}_d^{(q)} \mathbf{D}^H \mathbf{s}_d + \sum_{i=1}^{U} \mathbf{C}_{u,i}^{(q)} \mathbf{D}^H \mathbf{s}_{u,i} + \mathbf{n}^{(q)}, \qquad (1)$$

where $\mathbf{s}_d = [s_0^d, \ldots, s_{M-1}^d]^T$ denotes a transmitted signal block of the desired base station, $\mathbf{s}_{u,i} = [s_0^{u,i}, \ldots, s_{M-1}^{u,i}]^T$, $(i = 1, \ldots, U)$ is a transmitted signal block from the i-th interfering base station and $\mathbf{n}^{(q)}$ is additive white noise with a covariance matrix $\sigma_n^2 \mathbf{I}_M$. Here, note that all the elements of $\mathbf{s}_d$ will be detected in the mobile terminal regardless of subcarrier allocation. $\mathbf{C}_d^{(q)}$ and $\mathbf{C}_{u,i}^{(q)}$, $(q = 1 \ldots Q,\ i = 1 \ldots, U)$ are $M \times M$ circulant channel matrices defined as

$$\mathbf{C}_d^{(q)} = \begin{bmatrix} h_{0,q}^d & & h_{K,q}^d & \cdots & h_{1,q}^d \\ \vdots & \ddots & & \ddots & \vdots \\ h_{K,q}^d & & \ddots & & h_{K,q}^d \\ & \ddots & & \ddots & \\ & & h_{K,q}^d & \cdots & h_{0,q}^d \end{bmatrix}, \qquad (2)$$

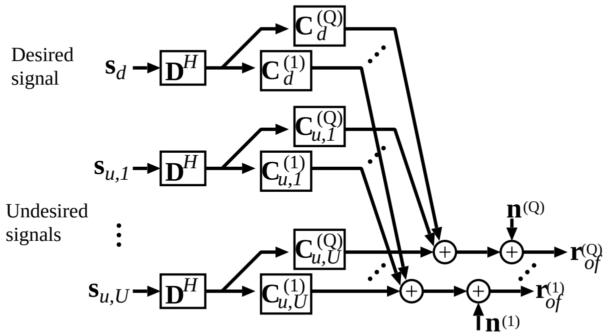

Fig. 1. System Model of OFDMA

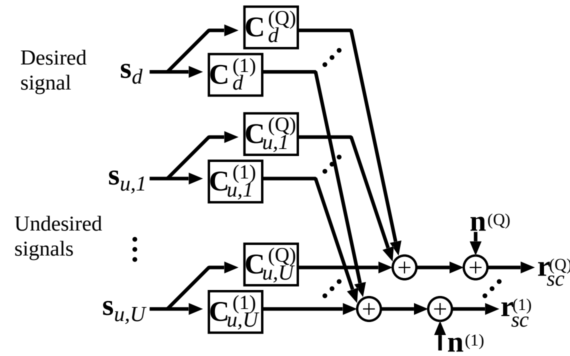

Fig. 2. System Model of SC-CP

where $\{h_{0,q}^d, \ldots, h_{K,q}^d\}$ denotes a channel impulse response between the desired base station and the q-th antenna of the terminal, and

$$\mathbf{C}_{u,i}^{(q)} = \begin{bmatrix} h_{0,q}^{u,i} & & h_{K,q}^{u,i} & \ldots & h_{1,q}^{u,i} \\ \vdots & \ddots & & \ddots & \vdots \\ h_{K,q}^{u,i} & & \ddots & & h_{K,q}^{u,i} \\ & \ddots & & \ddots & \\ & & h_{K,q}^{u,i} & \ldots & h_{0,q}^{u,i} \end{bmatrix}, \tag{3}$$

where $\{h_{0,q}^{u,i}, \ldots, h_{K,q}^{u,i}\}$, $(i = 1, \ldots, U)$ denotes a channel impulse response between the i-th interfering base station and the q-th antenna of the terminal, respectively.

After the DFT at the receiver, the received signal vector is given by

$$\begin{aligned} \mathbf{R}_{of}^{(q)} &= \mathbf{D}\mathbf{r}_{of}^{(q)} \\ &= \mathbf{\Lambda}_d^{(q)}\mathbf{s}_d + \sum_{i=1}^{U} \mathbf{\Lambda}_{u,i}^{(q)}\mathbf{s}_{u,i} + \mathbf{N}^{(q)}, \end{aligned} \tag{4}$$

where $\mathbf{N}^{(q)} = \mathbf{D}\mathbf{n}^{(q)}$ and

$$\mathbf{\Lambda}_d^{(q)} = \mathrm{diag}[\lambda_d^{(q,0)}, \ldots, \lambda_d^{(q,M-1)}] = \mathbf{D}\mathbf{C}_d^{(q)}\mathbf{D}^H, \tag{5}$$

$$\mathbf{\Lambda}_{u,i}^{(q)} = \mathrm{diag}[\lambda_{u,i}^{(q,0)}, \ldots, \lambda_{u,i}^{(q,M-1)}] = \mathbf{D}\mathbf{C}_{u,i}^{(q)}\mathbf{D}^H. \tag{6}$$

By stacking Q received signal vectors $\mathbf{R}_{of}^{(q)}$, $(q = 1 \ldots Q)$, we obtain a received signal vector of size $QM \times 1$ as

$$\begin{aligned} \mathbf{R}_{of} &\overset{\text{def}}{=} \left[\mathbf{R}_{of}^{(1)T} \quad \ldots \quad \mathbf{R}_{of}^{(Q)T}\right]^T \\ &= \begin{bmatrix} \mathbf{\Lambda}_d^{(1)} & \mathbf{\Lambda}_{u,1}^{(1)} & \ldots & \mathbf{\Lambda}_{u,U}^{(1)} \\ \vdots & \vdots & & \vdots \\ \mathbf{\Lambda}_d^{(Q)} & \mathbf{\Lambda}_{u,1}^{(Q)} & \ldots & \mathbf{\Lambda}_{u,U}^{(Q)} \end{bmatrix} \begin{bmatrix} \mathbf{s}_d \\ \mathbf{s}_{u,1} \\ \vdots \\ \mathbf{s}_{u,U} \end{bmatrix} + \begin{bmatrix} \mathbf{N}^{(1)} \\ \vdots \\ \mathbf{N}^{(Q)} \end{bmatrix} \\ &= \mathbf{\Lambda}_d\mathbf{s}_d + \sum_{i=1}^{U} \mathbf{\Lambda}_{u,i}\mathbf{s}_{u,i} + \mathbf{N}, \end{aligned} \tag{7}$$

where $\mathbf{\Lambda}_d = \left[\mathbf{\Lambda}_d^{(1)} \ldots \mathbf{\Lambda}_d^{(Q)}\right]^T$, $\mathbf{\Lambda}_{u,i} = \left[\mathbf{\Lambda}_{u,i}^{(1)} \ldots \mathbf{\Lambda}_{u,i}^{(Q)}\right]^T$ and $\mathbf{N} = \left[\mathbf{N}^{(1)T} \ldots \mathbf{N}^{(Q)T}\right]^T$.

B. SC-CP

Fig.2 shows the system model considered in this paper for the case of the SC-CP transmission. With the same assumptions as in the OFDMA case, the received signal vector at the q-th antenna is given by

$$\mathbf{r}_{sc}^{(q)} = \mathbf{C}_d^{(q)}\mathbf{s}_d + \sum_{i=1}^{U} \mathbf{C}_{u,i}^{(q)}\mathbf{s}_{u,i} + \mathbf{n}^{(q)}. \tag{8}$$

Also, stacking Q received signal vectors $\mathbf{r}_{sc}^{(q)}$, $(q = 1 \ldots Q)$, we obtain a received signal vector as

$$\begin{aligned} \mathbf{r}_{sc} &\overset{\text{def}}{=} \left[\mathbf{r}_{sc}^{(1)T} \quad \ldots \quad \mathbf{r}_{sc}^{(Q)T}\right]^T \\ &= \begin{bmatrix} \mathbf{C}_d^{(1)} & \mathbf{C}_{u,1}^{(1)} & \ldots & \mathbf{C}_{u,U}^{(1)} \\ \vdots & \vdots & & \vdots \\ \mathbf{C}_d^{(Q)} & \mathbf{C}_{u,1}^{(Q)} & \ldots & \mathbf{C}_{u,U}^{(Q)} \end{bmatrix} \begin{bmatrix} \mathbf{s}_d \\ \mathbf{s}_{u,1} \\ \vdots \\ \mathbf{s}_{u,U} \end{bmatrix} + \begin{bmatrix} \mathbf{n}^{(1)} \\ \vdots \\ \mathbf{n}^{(Q)} \end{bmatrix} \\ &= \mathbf{C}_d\mathbf{s}_d + \sum_{i=1}^{U} \mathbf{C}_{u,i}\mathbf{s}_{u,i} + \mathbf{n}, \end{aligned} \tag{9}$$

where $\mathbf{C}_d = \left[\mathbf{C}_d^{(1)T} \ldots \mathbf{C}_d^{(Q)T}\right]^T$, $\mathbf{C}_{u,i} = \left[\mathbf{C}_{u,i}^{(1)T} \ldots \mathbf{C}_{u,i}^{(Q)T}\right]^T$ and $\mathbf{n} = \left[\mathbf{n}^{(1)T} \ldots \mathbf{n}^{(Q)T}\right]^T$.

III. CO-CHANNEL INTERFERENCE CANCELLATION / EQUALIZATION

In this section, we consider MMSE based and iterative LS based co-channel interference cancellation and equalization methods for both the multi-carrier (OFDMA) and the single-carrier (SC-CP) cases.

A. OFDMA

1) MMSE approach: Fig.3 shows the configuration of the MMSE base approach for OFDMA system. Denoting $M \times QM$ matrix of the MMSE interference canceller / equalizer as $\mathbf{F}_{of}^H$, the estimate of $\mathbf{s}_d$ is obtained as

$$\begin{aligned} \hat{\mathbf{s}}_d^{mmse} &= \mathbf{F}_{of}^H\mathbf{R}_{of}, \\ &= \mathbf{F}_{of}^H\mathbf{\Lambda}_d\mathbf{s}_d + \mathbf{F}_{of}^H \sum_{i=1}^{U} \mathbf{\Lambda}_{u,i}\mathbf{s}_{u,i} + \mathbf{F}_{of}^H\mathbf{N}. \end{aligned} \tag{10}$$

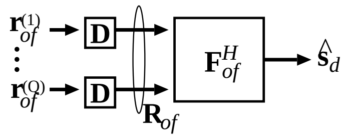

Fig. 3. MMSE interference canceller / equalizer: OFDMA

By solving a minimization problem of cost function

$$J = E\left[||\mathbf{s}_d - \mathbf{F}_{of}^H \mathbf{R}_{of}||^2\right], \tag{11}$$

MMSE canceller / equalizer matrix $\mathbf{F}_{of}^H$ is obtained as

$$\mathbf{F}_{of}^H = \mathbf{\Lambda}_d^H \left(\mathbf{\Lambda}_d \mathbf{\Lambda}_d^H + \sum_{i=1}^{U} \mathbf{\Lambda}_{u,i} \mathbf{\Lambda}_{u,i}^H + \frac{\sigma_n^2}{\sigma_s^2} \mathbf{I}_{QM}\right)^{-1}, \tag{12}$$

where σ_s^2 denotes the variance of $\mathbf{s}_d$ and $\mathbf{s}_{u,i}$. Note that the inverse of large non-diagonal matrix, which results in high computational complexity in general, is required in order to obtain $\mathbf{F}_{of}$.

2) Iterative LS approach: If we look on the second and the third terms in (7) as noise, we can simply obtain the LS estimate of $\mathbf{s}_d$ as

$$\hat{\mathbf{s}}_d^0 = \mathbf{\Lambda}_d^\dagger \mathbf{R}_{of}, \tag{13}$$

where $\mathbf{\Lambda}_d^\dagger$ is a pseudo inverse matrix of $\mathbf{\Lambda}_d$, which is calculated as

$$\begin{aligned} \mathbf{\Lambda}_d^\dagger &= \left(\mathbf{\Lambda}_d^H \mathbf{\Lambda}_d\right)^{-1} \mathbf{\Lambda}_d^H \\ &= \left[\sum_{q=1}^{Q} \mathbf{\Lambda}_d^{(q)H} \mathbf{\Lambda}_d^{(q)}\right]^{-1} \mathbf{\Lambda}_d^H \\ &= \left[\mathbf{\Gamma}_d^{(1)} \quad \ldots \quad \mathbf{\Gamma}_d^{(Q)}\right], \end{aligned} \tag{14}$$

where

$$\mathbf{\Gamma}_d^{(q)} = \text{diag}\left[\frac{\lambda_d^{(q,0)*}}{\sum_{p=1}^{Q} |\lambda_d^{(p,0)}|^2} \quad \cdots \quad \frac{\lambda_d^{(q,M-1)*}}{\sum_{p=1}^{Q} |\lambda_d^{(p,M-1)}|^2}\right]. \tag{15}$$

In the same way, the LS estimate of $\mathbf{s}_{u,i}$ is obtained as

$$\hat{\mathbf{s}}_{u,i}^0 = \mathbf{\Lambda}_{u,i}^\dagger \mathbf{R}_{of}, \tag{16}$$

where

$$\begin{aligned} \mathbf{\Lambda}_{u,i}^\dagger &= \left(\mathbf{\Lambda}_{u,i}^H \mathbf{\Lambda}_{u,i}\right)^{-1} \mathbf{\Lambda}_{u,i}^H \\ &= \left[\sum_{q=1}^{Q} \mathbf{\Lambda}_{u,i}^{(q)H} \mathbf{\Lambda}_{u,i}^{(q)}\right]^{-1} \mathbf{\Lambda}_{u,i}^H \\ &= \left[\mathbf{\Gamma}_{u,i}^{(1)} \quad \ldots \quad \mathbf{\Gamma}_{u,i}^{(Q)}\right], \end{aligned} \tag{17}$$

and

$$\mathbf{\Gamma}_u^{(q)} = \text{diag}\left[\frac{\lambda_{u,i}^{(q,0)*}}{\sum_{p=1}^{Q} |\lambda_{u,i}^{(p,0)}|^2} \quad \cdots \quad \frac{\lambda_{u,i}^{(q,M-1)*}}{\sum_{q=1}^{Q} |\lambda_{u,i}^{(p,M-1)}|^2}\right]. \tag{18}$$

Note that, although the estimation accuracy of $\hat{\mathbf{s}}_d^0$ is worse than $\hat{\mathbf{s}}_d^{mmse}$, the pseudoinverse matrices $\mathbf{\Lambda}_d^\dagger$ and $\mathbf{\Lambda}_{u,i}^\dagger$ can be obtained without any high computational complexity and the multiplication to the received signal is also achieved efficiently due to the diagonal structure.

Moreover, using $\hat{\mathbf{s}}_{u,i}^0$, the estimation of $\mathbf{s}_d$ can be improved as

$$\hat{\mathbf{s}}_d^1 = \mathbf{\Lambda}_d^\dagger \left(\mathbf{R}_{of} - \sum_{i=1}^{U} \mathbf{\Lambda}_{u,i} \hat{\mathbf{s}}_{u,i}^0\right), \tag{19}$$

and the estimation of $\mathbf{s}_{u,i}$ can be also improved as

$$\hat{\mathbf{s}}_{u,i}^1 = \mathbf{\Lambda}_{u,i}^\dagger \left(\mathbf{R}_{of} - \mathbf{\Lambda}_d \hat{\mathbf{s}}_d^0 - \sum_{\substack{j=1 \\ j \neq i}}^{U} \mathbf{\Lambda}_{u,j} \hat{\mathbf{s}}_{u,j}^0\right). \tag{20}$$

Denoting the estimate of $\mathbf{s}_d$ and $\mathbf{s}_{u,i}$ at the k-th iteration of this process as $\hat{\mathbf{s}}_d^k$ and $\hat{\mathbf{s}}_{u,i}^k$, respectively, the iterative LS interference cancellation and equalization scheme is given as follows:

1) $k = 0$
2) Initial estimation of $\mathbf{s}_d$:
$$\hat{\mathbf{s}}_d^k = \mathbf{\Lambda}_d^\dagger \mathbf{R}_{of}$$
3) Initial estimation of $\mathbf{s}_{u,i}$, $(i = 1, \ldots, U)$:
$$\hat{\mathbf{s}}_{u,i}^k = \mathbf{\Lambda}_{u,i}^\dagger \mathbf{R}_{of}$$
4) $k = k + 1$
5) Estimation of $\mathbf{s}_d$:
$$\hat{\mathbf{s}}_d^k = \mathbf{\Lambda}_d^\dagger \left(\mathbf{R}_{of} - \sum_{i=1}^{U} \mathbf{\Lambda}_{u,i} \hat{\mathbf{s}}_{u,i}^{k-1}\right)$$
6) Estimation of $\mathbf{s}_{u,i}$, $(i = 1, \ldots, U)$:
$$\hat{\mathbf{s}}_{u,i}^k = \mathbf{\Lambda}_{u,i}^\dagger \left(\mathbf{R}_{of} - \mathbf{\Lambda}_d \hat{\mathbf{s}}_d^{k-1} - \sum_{\substack{j=1 \\ j \neq i}}^{U} \mathbf{\Lambda}_{u,j} \hat{\mathbf{s}}_{u,j}^{k-1}\right)$$
7) Go to step 4

The configuration of the iterative LS method is shown in Fig.4, where the number of interfering base stations is set to be $U = 1$ for the simplicity. Since the matrices $\mathbf{\Gamma}_{\mathbf{d}}^{(\mathbf{q})}$, $\mathbf{\Lambda}_{\mathbf{d}}^{(\mathbf{q})}$, $\mathbf{\Gamma}_{\mathbf{u,i}}^{(\mathbf{q})}$, and $\mathbf{\Lambda}_{\mathbf{u,i}}^{(\mathbf{q})}$ are all diagonal matrix, the iterative LS method can be efficiently realized by using the same configuration as one-tap frequency domain equalizers.

B. SC-CP

1) MMSE approach: For the case of the SC-CP transmission, the output of the MMSE interference canceller / equalizer is given by

$$\hat{\mathbf{s}}_d = \mathbf{F}_{sc}^H \mathbf{r}_{sc}, \tag{21}$$

where the $M \times QM$ matrix $\mathbf{F}_{sc}^H$ is obtained as

$$\mathbf{F}_{sc}^H = \mathbf{C}_d^H \left(\mathbf{C}_d \mathbf{C}_d^H + \sum_{i=1}^{U} \mathbf{C}_{u,i} \mathbf{C}_{u,i}^H + \frac{\sigma_n^2}{\sigma_s^2} \mathbf{I}_{QM}\right)^{-1}. \tag{22}$$

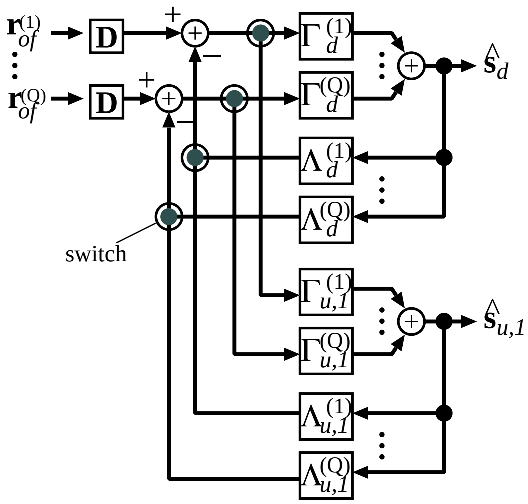

Fig. 4. Iterative LS interference canceller / equalizer: OFDMA

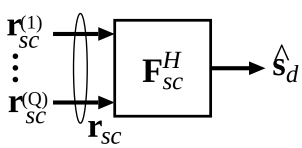

Fig. 5. MMSE interference canceller / equalizer: SC-CP

2) Iterative LS approach: Comparing (7) with (9), we can easily obtain the iterative LS interference cancellation and equalization method for the SC-CP transmission. The iterative algorithm is given as follows:

1) $k = 0$
2) Initial estimation of $\mathbf{s}_d$:

$$\hat{\mathbf{s}}_d^k = \mathbf{C}_d^{\dagger}\mathbf{r}_{sc}$$

3) Initial estimation of $\mathbf{s}_{u,i}$, $(i = 1, \ldots, U)$:

$$\hat{\mathbf{s}}_{u,i}^k = \mathbf{C}_{u,i}^{\dagger}\mathbf{r}_{sc}$$

4) $k = k + 1$
5) Estimation of $\mathbf{s}_d$:

$$\hat{\mathbf{s}}_d^k = \mathbf{C}_d^{\dagger}\left(\mathbf{r}_{sc} - \sum_{i=1}^{U}\mathbf{C}_{u,i}\hat{\mathbf{s}}_{u,i}^{k-1}\right)$$

6) Estimation of $\mathbf{s}_{u,i}$, $(i = 1, \ldots, U)$:

$$\hat{\mathbf{s}}_{u,i}^k = \mathbf{C}_{u,i}^{\dagger}\left(\mathbf{r}_{sc} - \mathbf{C}_d\hat{\mathbf{s}}_d^{k-1} - \sum_{\substack{j=1\\ j\neq i}}^{U}\mathbf{C}_{u,j}\hat{\mathbf{s}}_{u,j}^{k-1}\right)$$

7) Go to step 4

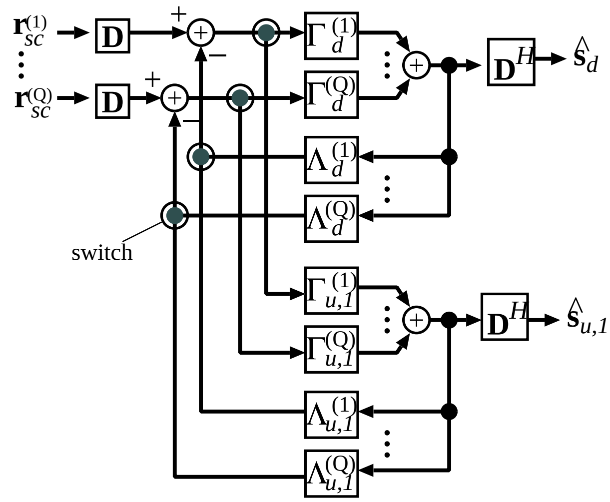

Fig. 6. Iterative LS interference canceller / equalizer: SC-CP

The pseudo inverse $\mathbf{C}_d^{\dagger}$ and $\mathbf{C}_{u,i}^{\dagger}$ can be simplified as

$$\begin{aligned}
\mathbf{C}_d^{\dagger} &= \left(\mathbf{C}_d^H\mathbf{C}_d\right)^{-1}\mathbf{C}_d^H \\
&= \left[\textstyle\sum_{q=1}^{Q}\mathbf{C}_d^{(q)H}\mathbf{C}_d^{(q)}\right]^{-1}\mathbf{C}_d^H \\
&= \left[\mathbf{D}^H\left(\textstyle\sum_{q=1}^{Q}\mathbf{\Lambda}_d^{(q)H}\mathbf{\Lambda}_d^{(q)}\right)\mathbf{D}\right]^{-1}\mathbf{D}^H\mathbf{\Lambda}_d^H\mathbf{D} \\
&= \mathbf{D}^H\left(\sum_{q=1}^{Q}\mathbf{\Lambda}_d^{(q)H}\mathbf{\Lambda}_d^{(q)}\right)^{-1}\mathbf{\Lambda}_d^H\mathbf{D} \\
&= \mathbf{D}^H\left[\mathbf{\Gamma}_d^{(1)} \quad \ldots \quad \mathbf{\Gamma}_d^{(Q)}\right]\mathbf{D},
\end{aligned} \tag{23}$$

and

$$\begin{aligned}
\mathbf{C}_{u,i}^{\dagger} &= \left(\mathbf{C}_{u,i}^H\mathbf{C}_{u,i}\right)^{-1}\mathbf{C}_{u,i}^H \\
&= \mathbf{D}^H\left[\mathbf{\Gamma}_{u,i}^{(1)} \quad \ldots \quad \mathbf{\Gamma}_{u,i}^{(Q)}\right]\mathbf{D},
\end{aligned} \tag{24}$$

where $\mathbf{\Gamma}_d^{(q)}$ and $\mathbf{\Gamma}_{u,i}^{(q)}$ are defined in (15) and (18), respectively. Using the simplified expressions of the pseudo inverse matrices, the iterative LS interference canceller / equalizer for the SC-CP also has a simple configuration as shown in Fig. 6. Here, note that no DFT or inverse DFT operation is required in the iterative process.

IV. Computer Simulation

In order to confirm the performance of the proposed methods, computer simulations are conducted with the following system parameters; Mod./Demod. scheme: QPSK coherent detection, symbols per block: $M = 128$, length of cyclic prefix: $K = 16$, number of antennas: $Q = 2$ or 4, channel order: $L = 16$, channel model: 17-path frequency selective Rayleigh fading channel, number of interfering base stations: $U = 1$, channel noise: additive white Gaussian noise (AWGN). The signal to noise power ratio (SNR) per antenna is set to be 10[dB]. Also, all the subcarriers or time slots of the signals

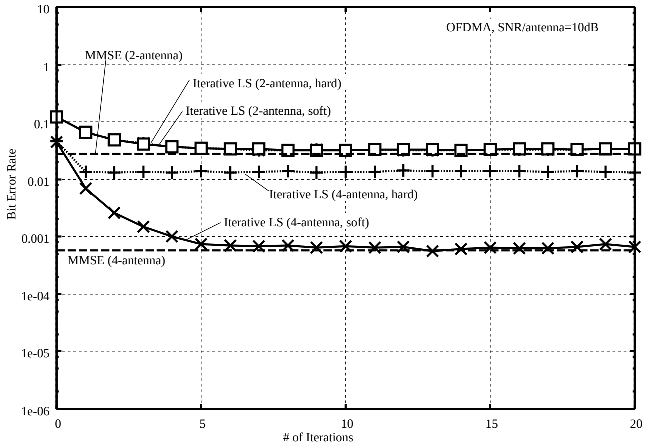

Fig. 7. Bit Error Rate Performance: OFDMA

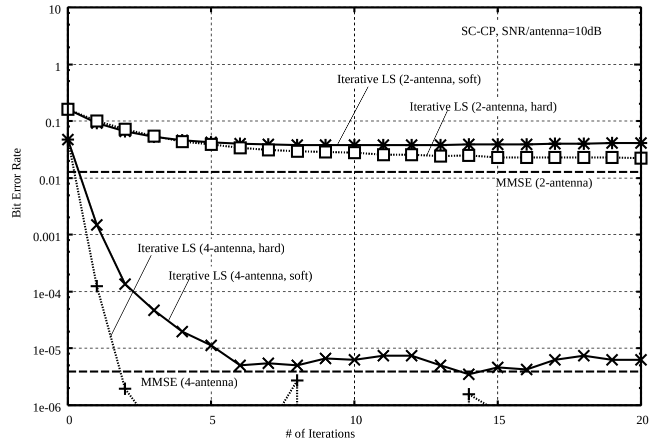

Fig. 8. Bit Error Rate Performance: SC-CP

from the desired base station are assumed to be used for the data transmission of the mobile terminal.

Fig. 7 shows the BER versus the number of iterations of the proposed iterative LS interference cancellation and equalization method for the OFDMA system. In the figure, “soft” means soft decision, which corresponds to the algorithm shown in Sec. III-A.2, and “hard” means an introduction of hard decision for each estimate of $\mathbf{s}_d$ or $\mathbf{s}_{u,i}$. The performance of the MMSE approach is also plotted in the same figure. From the figure, we can see that iterative LS approach with soft decision can achieve comparable performance as the MMSE approach with around 5 iterations, while the performance of the MMSE approach is the best among the three methods. We can also see that the introduction of the hard decision degrades the performance for the case of the OFDMA system.

Fig. 8 shows the BER versus the number of iterations of the proposed iterative LS interference cancellation and equalization method for the SC-CP system. We can also see that the iterative LS approach with soft decision can achieve almost the same performance as the MMSE method with only 5 iterations. For the case of SC-CP system, the introduction of the hard decision could improve the BER performance, while it increases the computational complexity due to the FFT and IFFT operation in the iterative process. So far, we don’t have analytical explanations for this phenomenon, however, the principle of turbo signal processing used in the MMSE-based turbo equalizer[7],[8] may give us an insight into the phenomenon.

V. Conclusion

We have proposed co-channel interference cancellation and equalization methods for downlink block transmission with cyclic prefix in multi-cell environments, assuming the availability of the channel state information of not only the desired base station and the mobile terminal, but also the interfering base stations and the terminal. We have considered the MMSE based method as a performance benchmark and the simple iterative LS based approach for both the OFDMA and the SC-CP schemes. Computer simulation results show that the iterative LS based approach can achieve comparable performance to that of the MMSE approach with only 5 iterations. Moreover, for the case of the SC-CP transmission, the iterative LS with the hard decision could outperform the MMSE method.

Future study will include the analytical reasoning of the difference in performance between the hard and soft decisions, and the introduction of error correcting code or belief propagation based iterative signal processing.

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