Iterative MMSE Detection and Interference Cancellation for Uplink SC-FDMA MIMO using HARQ

Suguru OKUYAMA† Kazuki TAKEDA† and Fumiyuki ADACHI‡
Dept. of Electrical and Communication Engineering, Graduate School of Engineering, Tohoku University
6-6-05 Aza-Aoba, Aramaki, Aoba-ku, Sendai, 980-8579 Japan
E-mail: †{okuyama, kazuki}@mobile.ecei.tohoku.ac.jp, ‡adachi@ecei.tohoku.ac.jp

***Abstract*—For high-speed packet access, hybrid automatic repeat request (HARQ) and multiple-input multiple-output (MIMO) multiplexing are indispensable techniques. As the multiple-access technique, single carrier (SC)-frequency-division multiple-access (FDMA) using transmit filtering is promising. The use of square-root Nyquist transmit filter reduces peak-to-average power ratio (PAPR) of the transmit SC-FDMA signal and increases the frequency diversity gain. However, if the carrier-frequency separation among multiple-access users is kept the same as in the case of the transmit filter's roll-off factor α=0 (i.e., brick wall filter), the adjacent users' signal spectra overlap and the multi-user interference (MUI) is produced. In this paper, we propose an iterative minimum mean square error signal detection and interference cancellation (MMSED-IC) for turbo coded uplink SC-FDMA MIMO using HARQ. In the proposed MMSED-IC, not only the inter-antenna interference (IAI) and residual inter-symbol interference (ISI) but also MUI are cancelled in an iterative fashion. We evaluate the HARQ throughput performances by computer simulation, and show that the use of roll-off factor α=1 achieves the highest throughput performance.**

Keywords-component; SC-FDMA, MIMO, HARQ, MMSE, Iterative Interference cancellation, frequency-domain filtering

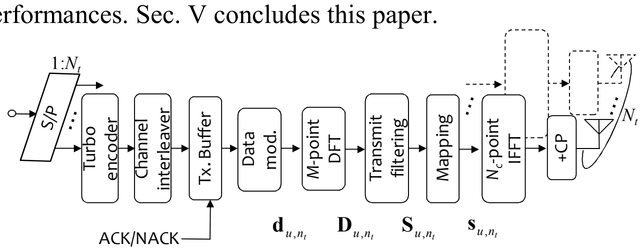
I. INTRODUCTION

In the next generation mobile communication systems, high speed, high quality and high efficient packet services are demanded. For broadband wireless single-carrier (SC) transmission, the use of minimum mean square error frequency-domain equalization (MMSE-FDE) can exploit the channel frequency-selectivity [1],[2] to improve the bit error rate (BER) performance [3-5]. One of the promising techniques for achieving highly frequency efficient packet transmissions is the multi-input multi-output (MIMO) multiplexing which can increase the transmission rate without bandwidth expansion. Hybrid automatic repeat request (HARQ) with error correcting coding (e.g., turbo code) is an indispensable error control technique for packet access. In Ref [6], iterative inter-antenna interference (IAI) and residual inter-symbol interference (ISI) cancellation using MMSE-FDE was proposed, for SC-MIMO multiplexing with turbo coded HARQ.

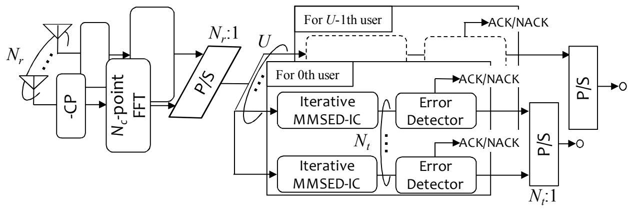
To limit the signal bandwidth in the wireless communication systems, the square-root Nyquist filter can be used as transmit/receive filters [2],[7]. As the filter's roll-off factor α increases, the peak-to-average power ratio (PAPR) decreases and furthermore, an increased frequency diversity gain can be obtained by making use of the excess bandwidth introduced by the transmit filtering [8]. However, if the carrier-frequency separation among multiple-access users is kept the same as in the case of α=0 to prevent degradation of spectrum use efficiency, the adjacent users' signal spectra overlap and the multi-user interference (MUI) is produced.

In this paper, we propose an iterative MMSE detection and interference cancellation (MMSED-IC) for turbo coded uplink SC-FDMA MIMO using HARQ with frequency-domain transmit filtering. First, adjacent users' overlapped signals and multiplexed signals from different transmit antennas are separated using MMSED and then, spectrum combining [8] is done to restore the desired user's signal spectrum. Finally, not only the IAI and ISI but also the MUI cancellation is carried out. The above series of operations is repeated sufficient times. By computer simulation, we evaluate the PAPR and HARQ throughput performances of uplink SC-FDMA MIMO using MMSED-IC and show that SC-FDMA with α=1 can achieve the highest throughput performance compare to SC-FDMA with α=0 and OFDMA using IAI cancellation [9].

The remainder of this paper is organized as follows. The frequency-domain filtered uplink SC-FDMA MIMO using HARQ is presented in Sec. II. In Sec. III, we describe the proposed iterative MMSED-IC. In Sec. IV, we evaluate the simulation results for PAPR and HARQ throughput performances. Sec. V concludes this paper.



(a)Transmitter of the *u*th user



(b)BS Receiver

Fig. 1. Uplink SC-FDMA MIMO transmitter/receiver structure.

II. FREQUENCY-DOMAIN FILTERED UPLINK SC-FDMA MIMO SIGNAL TRANSMISSION

The transmitter/receiver structure of frequency-domain filtered uplink SC-FDMA MIMO is illustrated in Fig. 1. Throughout the paper, fast Fourier transform (FFT) sample-spaced discrete-time signal representation is used. At the *u*th (u=0,…,U–1) user's transmitter, an information bit sequence to be transmitted is serial-parallel (S/P) converted into N_t bit streams. Each bit stream is turbo encoded, interleaved, and data-modulated. The resulting data-modulated symbol stream is divided into a sequence of blocks of M-symbol each. The uth user's symbol block of the n_tth (n_t=0,…,N_t–1) symbol stream is expressed as $\mathbf{d}_{u,n_t}=[d_{u,n_t}(0),\ldots,d_{u,n_t}(M-1)]^T$ using the vector form. $\mathbf{d}_{u,n_t}$ is transformed by M-point discrete Fourier transform (DFT) into the frequency-domain signal vector $\mathbf{D}_{u,n_t}$. To limit the signal bandwidth, the square root-raised cosine Nyquist filter with the roll-off factor α is applied. The transmit filter transfer function is denoted by $\{f(k);k=-M\sim M-1\}$. The frequency-domain signal after transmit filtering is mapped over N_c (>>M) subcarriers, to obtain the transmit frequency-domain signal. We assume the full loaded condition, in which U (=N_c/M) users access the same base station (BS) simultaneously.

In this paper, we consider the localized spectrum mapping illustrated in Fig. 2. The carrier-frequency separation is kept the same as in the case of α=0 to accommodate the same number of users irrespective of α. The frequency-domain signal after spectrum mapping is transformed back to the time-domain signal by applying N_c-point inverse FFT (IFFT). Last N_g samples of each block are copied as a cyclic prefix (CP) and inserted into the guard interval (GI) placed at the beginning of each block.

The CP-inserted signal vector is transmitted over a frequency-selective fading channel and received at the n_rth (n_r=0,…,N_r–1) BS receive antenna. The received signal vector after the CP removal is transformed by applying N_c-point FFT into the frequency-domain signal. Iterative signal detection is carried out using the entire received frequency-domain signal. The number of iterations is denoted by I.

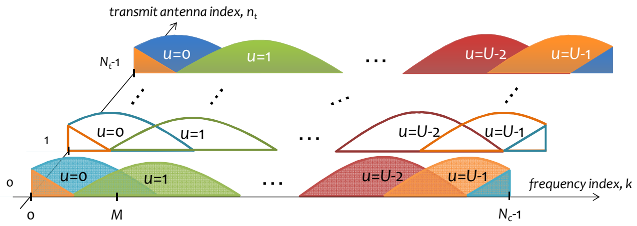


Fig. 2. Spectrum mapping

A. Transmit Signal

The data symbol vector $\mathbf{d}_{u,n_t}=[d_{u,n_t}(0),\ldots,d_{u,n_t}(M-1)]^T$ is first transformed by M-point DFT into the frequency-domain signal vector $\mathbf{D}_{u,n_t}=[D_{u,n_t}(0),\ldots,D_{u,n_t}(M-1)]^T$, to which the frequency-domain transmit filtering and spectrum mapping are applied. The frequency-domain filtered signal $\mathbf{S}_{u,n_t}=[S_{u,n_t}(0),\ldots,S_{u,n_t}(N_c-1)]^T$ is expressed as

$$\mathbf{S}_{u,n_t}=\mathbf{\Phi}_u\mathbf{H}_T\mathbf{D}_{u,n_t}=\mathbf{\Phi}_u\mathbf{H}_T\mathbf{F}_M\mathbf{d}_{u,n_t}, \tag{1}$$

where $\mathbf{F}_K$ is a K×K DFT matrix, given as

$$\mathbf{F}_K=\frac{1}{\sqrt{K}}\begin{bmatrix}1 & 1 & \cdots & 1\\ 1 & e^{\frac{-j2\pi\cdot1\cdot1}{K}} & \cdots & e^{\frac{-j2\pi\cdot1\cdot(K-1)}{K}}\\ \vdots & \vdots & \ddots & \vdots\\ 1 & e^{\frac{-j2\pi\cdot(K-1)\cdot1}{K}} & \cdots & e^{\frac{-j2\pi\cdot(K-1)\cdot(K-1)}{K}}\end{bmatrix}, \tag{2}$$

and $\mathbf{H}_T$ is a $2M$×M transmit filter matrix. The (x,y)th (x=0,…,2M–1;y=0,..,M–1) element, $H_T(x,y)$, of $\mathbf{H}_T$ is expressed as

$$H_T(x,y)=\begin{cases}f(x-M) & y=x \bmod M\\ 0 & otherwise\end{cases}, \tag{3}$$

where $f(k)$ is the transmit filter transfer function. In this paper, the square root-raised cosine Nyquist filter with roll-off factor α is assumed, i.e.,

$$f(k)=\begin{cases}1, & 0\le|k|<\frac{1-\alpha}{2}M\\ \cos\left[\frac{\pi}{2\alpha}\left(\frac{|k|}{M}-\frac{1-\alpha}{2}\right)\right], & \frac{1-\alpha}{2}M\le|k|<\frac{1+\alpha}{2}M\\ 0, & otherwise\end{cases}. \tag{4}$$

$\mathbf{\Phi}_u$ is an N_c×2M spectrum mapping matrix. The (x,y)th (x=0,…,N_c–1;y=0,…,2M–1) element, $\Phi_u(x,y)$, of $\mathbf{\Phi}_u$ is expressed as

$$\Phi_u(x,y)=\begin{cases}1 & x=(y+(u-1/2)M)\bmod N_c\\ 0 & otherwise\end{cases}. \tag{5}$$

An N_c-point IFFT is applied to $\mathbf{S}_{u,n_t}=[S_{u,n_t}(0),\ldots,S_{u,n_t}(N_c-1)]^T$ to obtain the transmit time-domain signal $\mathbf{s}_{u,n_t}=[s_{u,n_t}(0),\ldots,s_{u,n_t}(N_c-1)]^T$, which is given by

$$\mathbf{s}_{u,n_t}=\sqrt{\frac{2E_{s,u}}{T_s}}\mathbf{F}_{N_c}^H\mathbf{S}_{u,n_t}, \tag{6}$$

where $E_{s,u}$ and T_s are the symbol energy and symbol duration, respectively, and $[.]^H$ denotes the Hermitian transpose operation.

B. Received Signal

In this paper, we consider Chase combining as a packet combining and assume that the same packet has been transmitted Q times (i.e., the number of retransmissions is Q–1). The propagation channel is assumed to be an FFT sample-spaced L-path block fading channel, each path being subjected to independent fading. Let $h^{(q)}_{u,n_t,n_r,l}$ and $\tau_{u,l}$ be respectively the complex-valued path gain of lth path for uth user between the n_tth transmit antenna and the n_rth receive antenna at the qth (q=1,…,Q) packet transmission and delay time of the lth path for uth user (l=0,…,L–1). The channel impulse response is expressed as

$$h^{(q)}_{u,n_t,n_r}(\tau)=\sum_{l=0}^{L-1}h^{(q)}_{u,n_t,n_r,l}\delta(\tau-\tau_{u,l}), \tag{7}$$

where δ(τ) is the delta function.

The received signal vector $\mathbf{r}^{(q)}_{n_r}=[r^{(q)}_{n_r}(0),\ldots,r^{(q)}_{n_r}(N_c-1)]^T$ on the n_rth receive antenna can be expressed as

$$\mathbf{r}^{(q)}_{n_r}=\sum_{u=0}^{U-1}\sum_{n_t=0}^{N_t-1}\mathbf{h}^{(q)}_{u,n_t,n_r}\mathbf{s}_{u,n_t}+\mathbf{n}^{(q)}_{n_r}, \tag{8}$$

where $\mathbf{n}^{(q)}_{n_r}=[n^{(q)}_{n_r}(0),\ldots,n^{(q)}_{n_r}(N_c-1)]^T$ is the noise vector with each element is an independent zero-mean complex Gaussian noise with variance $2N_0/T_s$ (N_0 is the single-sided power spectrum density of the additive white Gaussian noise (AWGN)). $\mathbf{h}^{(q)}_{u,n_t,n_r}$ is an N_c×N_c channel impulse response matrix for uth user between the n_tth transmit antenna and the n_rth receive antenna at the qth packet transmission given as

$$\mathbf{h}^{(q)}_{u,n_t,n_r}=\begin{bmatrix}h^{(q)}_{u,n_t,n_r,0} & & & h^{(q)}_{u,n_t,n_r,L-1} & \cdots & h^{(q)}_{u,n_t,n_r,1}\\ h^{(q)}_{u,n_t,n_r,1} & \ddots & & & \ddots & \vdots\\ \vdots & & h^{(q)}_{u,n_t,n_r,0} & \mathbf{0} & & h^{(q)}_{u,n_t,n_r,L-1}\\ h^{(q)}_{u,n_t,n_r,L-1} & & h^{(q)}_{u,n_t,n_r,1} & \ddots & & \\ & \ddots & \vdots & & \ddots & \\ \mathbf{0} & & h^{(q)}_{u,n_t,n_r,L-1} & \cdots & \cdots & h^{(q)}_{u,n_t,n_r,0}\end{bmatrix}. \tag{9}$$

N_c-point FFT is applied to $\mathbf{r}_{n_r}^{(q)}$ to transform it into the frequency-domain signal vector $\mathbf{R}_{n_r}^{(q)}=[R_{n_r}^{(q)}(0),\ldots,R_{n_r}^{(q)}(N_c-1)]^T$. $\mathbf{R}_{n_r}^{(q)}$ is given by

$$\mathbf{R}_{n_r}^{(q)}=\mathbf{F}_{N_c}\mathbf{r}_{n_r}^{(q)}=\mathbf{F}_{N_c}\sum_{u=0}^{U-1}\sum_{n_t=0}^{N_t-1}\mathbf{h}_{u,n_t,n_r}^{(q)}\mathbf{s}_{u,n_t}+\mathbf{F}_{N_c}\mathbf{n}_{n_r}^{(q)}. \quad (10)$$

Since the channel impulse matrix $\mathbf{h}_{u,n_t,n_r}^{(q)}$ is a circulant matrix, we have

$$\begin{aligned}&\mathbf{F}_{N_c}\mathbf{h}_{u,n_t,n_r}^{(q)}\mathbf{F}_{N_c}^H\\&=diag[H_{u,n_t,n_r}^{(q)}(0),\ldots,H_{u,n_t,n_r}^{(q)}(N_c-1)]\equiv\mathbf{H}_{u,n_t,n_r}^{(q)}\end{aligned}. \quad (11)$$

Using Eq. (11), Eq. (10) can be rewritten as

$$\mathbf{R}_{n_r}^{(q)}=\sum_{u=0}^{U-1}\sum_{n_t=0}^{N_t-1}\sqrt{\frac{2E_{s,u}}{T_s}}\overline{\mathbf{H}}_{u,n_t,n_r}^{(q)}\mathbf{D}_{u,n_t}+\mathbf{N}_{n_r}^{(q)}, \quad (12)$$

where $\overline{\mathbf{H}}_{u,n_t,n_r}^{(q)}$ is equivalent channel matrix for uth user between the n_tth transmit antenna and n_rth receive antenna given by $\overline{\mathbf{H}}_{u,n_t,n_r}^{(q)}=\mathbf{H}_{u,n_t,n_r}^{(q)}\mathbf{\Phi}_u\mathbf{H}_T$ and $\mathbf{N}_{n_r}^{(q)}=[N_{n_r}^{(q)}(0),\ldots,N_{n_r}^{(q)}(N_c-1)]^T$ is frequency-domain noise vector at the n_rth receive antenna. The qth receive signal vector $\mathbf{R}^{(q)}$ is given as

$$\mathbf{R}^{(q)}=\left[(\mathbf{R}_0^{(q)})^T\ldots(\mathbf{R}_{N_r-1}^{(q)})^T\right]^T=\sum_{u=0}^{U-1}\sum_{n_t=0}^{N_t-1}\overline{\mathbf{H}}_{u,n_t}^{(q)}\mathbf{D}_{u,n_t}+\mathbf{N}^{(q)} \quad (13)$$

where $\overline{\mathbf{H}}_{u,n_t}^{(q)}=[\overline{\mathbf{H}}_{u,n_t,0}^{(q)}\overline{\mathbf{H}}_{u,n_t,1}^{(q)}\ldots\overline{\mathbf{H}}_{u,n_t,N_r-1}^{(q)}]^T$.

III. Iterative MMSE Detection and Packet Combining / Interference Cancellartion

Iterative MMSED-IC separates overlapped spectra from adjacent users and different transmit antennas using MMSED and restores the desired user's signal spectrum by using the spectrum combining [8]. After MMSED and packet combining, the cancellation of MUI, IAI and ISI is performed. The frequency-domain signal vector $\hat{\mathbf{D}}_{u,n_t}^{(i)}=[\hat{D}_{u,n_t}^{(i)}(0),\ldots,\hat{D}_{u,n_t}^{(i)}(M-1)]^T$ at the uth user's n_tth transmit antenna after MMSED at the ith (i=0,…I–1) stage is given as

$$\begin{aligned}\hat{\mathbf{D}}_{u,n_t}^{(i)}&=\sum_{q=1}^{Q}\mathbf{W}_{u,n_t}^{(q,i)}\mathbf{R}^{(q)}\\&=\sqrt{\frac{2E_{s,u}}{T_s}}\mathbf{D}_{u,n_t}+\sqrt{\frac{2E_{s,u}}{T_s}}\left(\sum_{q=1}^{Q}\mathbf{W}_{u,n_t}^{(q,i)}\overline{\mathbf{H}}_{u,n_t}^{(q)}-\mathbf{I}\right)\mathbf{D}_{u,n_t}\\&\quad+\sum_{q=1}^{Q}\mathbf{W}_{u,n_t}^{(q,i)}\left(\sum_{u'=0\neq u}^{U-1}\sum_{n_t'=0}^{N_t-1}\sqrt{\frac{2E_{s,u'}}{T_s}}\overline{\mathbf{H}}_{u',n_t'}^{(q)}\mathbf{D}_{u',n_t'}\right)\\&\quad+\sqrt{\frac{2E_{s,u}}{T_s}}\sum_{q=1}^{Q}\mathbf{W}_{u,n_t}^{(q,i)}\left(\sum_{n_t'=0\neq n_t}^{N_t-1}\overline{\mathbf{H}}_{u,n_t'}^{(q)}\mathbf{D}_{u,n_t'}\right)+\sum_{q=1}^{Q}\mathbf{W}_{u,n_t}^{(q,i)}\mathbf{N}^{(q)}\end{aligned}, \quad (14)$$

and $\mathbf{W}_{u,n_t}^{(q,i)}$ is the MMSE weight matrix of size $M\times N_rN_c$ which minimizes the trace $trE[\{\mathbf{e}_{u,n_t}^{(i)}\}\{\mathbf{e}_{u,n_t}^{(i)}\}^H]$ of the covariance matrix of the error vector between $\mathbf{D}_{u,n_t}$ (the transmitted signal vector) and $\breve{\mathbf{D}}_{u,n_t}^{(i)}$ (the soft decision signal vector after interference cancellation).

Frequency-domain interference cancellation is carried out to simultaneously suppress the MUI, IAI and ISI, as

$$\breve{\mathbf{D}}_{u,n_t}^{(i)}=\hat{\mathbf{D}}_{u,n_t}^{(i)}-\widetilde{\mathbf{M}}_{u,n_t}^{(i)}-\widetilde{\mathbf{A}}_{u,n_t}^{(i)}-\widetilde{\mathbf{I}}_{u,n_t}^{(i)}. \quad (15)$$

where $\widetilde{\mathbf{M}}_{u,n_t}^{(i)}$, $\widetilde{\mathbf{A}}_{u,n_t}^{(i)}$ and $\widetilde{\mathbf{I}}_{u,n_t}^{(i)}$ are respectively the MUI, IAI and ISI replicas which will be shown in Sect. III. A.

Finally, the soft decision signal vector $\breve{\mathbf{d}}_{u,n_t}^{(i)}=[\breve{d}_{u,n_t}^{(i)}(0),\ldots,\breve{d}_{u,n_t}^{(i)}(M-1)]^T$ is obtained by applying M-point IDFT to $\breve{\mathbf{D}}_{u,n_t}^{(i)}=[\breve{D}_{u,n_t}^{(i)}(0),\ldots,\breve{D}_{u,n_t}^{(i)}(M-1)]^T$ as

$$\breve{\mathbf{d}}_{u,n_t}^{(i)}=\mathbf{F}_M^H\breve{\mathbf{D}}_{u,n_t}^{(i)}. \quad (16)$$

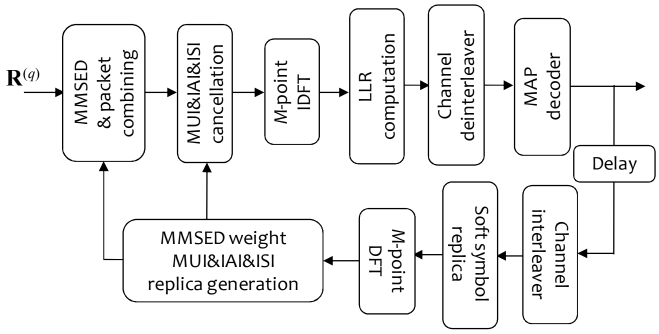


Fig. 3. Iterative MMSE detection MUI&IAI&ISI cancellation.

A. Interference Replica Generation

How to generate MUI, IAI and ISI replica of the uth user's n_tth transmit antenna at the ith stage is presented below. The receiver structure of interference cancellation is illustrated in Fig.3. The log-likelihood ratio (LLR) of the xth (x=0,…,N–1) bit in the mth symbols in a block (N is the number of bits per symbol and m=0,…,M–1) is computed using the decision variable $\{\breve{d}_{u,n_t}^{(i-1)}(m)\}$ [10] as

$$\begin{aligned}\lambda_{u,n_t,x}^{(i-1)}(m)&=\ln\left(\frac{p_{u,n_t}^{(i-1)}(b_{m,x}=1)}{p_{u,n_t}^{(i-1)}(b_{m,x}=0)}\right)\\&\approx\frac{\left|\breve{d}_{u,n_t}^{(i-1)}(m)-\sqrt{2E_{s,u}/T_s}\,d_{b_{m,x}=0}^{\min}\right|^2}{2\left(\hat{\sigma}_{u,n_t}^{(i-1)}\right)^2}-\frac{\left|\breve{d}_{u,n_t}^{(i-1)}(m)-\sqrt{2E_{s,u}/T_s}\,d_{b_{m,x}=1}^{\min}\right|^2}{2\left(\hat{\sigma}_{u,n_t}^{(i-1)}\right)^2}\end{aligned}, \quad (17)$$

where $p_{u,n_t}^{(i-1)}(b_{m,x}=0)$ and $p_{u,n_t}^{(i-1)}(b_{m,x}=1)$ are the *a posteriori* probabilities of the transmitted bit $b_{m,x}$ being $b_{m,x}$=0 and $b_{m,x}$=1, respectively, at the (i–1)th iteration stage, $d_{b_{m,x}=0}^{\min}$ (or $d_{b_{m,x}=1}^{\min}$) is the symbol whose xth bit is 0 (or 1) and which has the shortest Euclidean distance from $\breve{d}_{u,n_t}^{(i-1)}(m)$. $2\left(\hat{\sigma}_{u,n_t}^{(i-1)}\right)^2$ is sum of the variance of the MUI, IAI, ISI plus noise, given as

$$\begin{aligned}&2\left(\hat{\sigma}_{u,n_t}^{(i)}\right)^2\\&=2\left(\hat{\sigma}_{u,n_t,MUI}^{(i)}\right)^2+2\left(\hat{\sigma}_{u,n_t,IAI}^{(i)}\right)^2+2\left(\hat{\sigma}_{u,n_t,ISI}^{(i)}\right)^2+2\left(\hat{\sigma}_{u,n_t,noise}^{(i)}\right)^2\end{aligned}, \quad (18)$$

where $2\left(\hat{\sigma}_{u,n_t,MUI}^{(i)}\right)^2$, $2\left(\hat{\sigma}_{u,n_t,IAI}^{(i)}\right)^2$, $2\left(\hat{\sigma}_{u,n_t,ISI}^{(i)}\right)^2$, and $2\left(\hat{\sigma}_{u,n_t,noise}^{(i)}\right)^2$ are the MUI, IAI, ISI, and noise variances, respectively, and are given as

$$\begin{aligned}&2\left(\hat{\sigma}_{u,n_t,MUI}^{(i)}\right)^2\\&=\sum_{q=1}^{Q}\sum_{u'=0\neq u}^{U-1}\sum_{n_t'=0}^{N_t-1}\frac{2E_{s,u'}}{T_s}\frac{\rho_{u',n_t'}^{(i-1)}}{M}tr[(\mathbf{W}_{u,n_t}^{(q,i)}\overline{\mathbf{H}}_{u',n_t'}^{(q)})(\mathbf{W}_{u,n_t}^{(q,i)}\overline{\mathbf{H}}_{u',n_t'}^{(q)})^H]\end{aligned}, \quad (18a)$$

$$\begin{aligned}&2\left(\hat{\sigma}_{u,n_t,IAI}^{(i)}\right)^2\\&=\sum_{q=1}^{Q}\sum_{n_t'=0\neq n_t}^{N_t-1}\frac{2E_{s,u}}{T_s}\frac{\rho_{u,n_t'}^{(i-1)}}{M}tr[(\mathbf{W}_{u,n_t}^{(q,i)}\overline{\mathbf{H}}_{u,n_t'}^{(q)})(\mathbf{W}_{u,n_t}^{(q,i)}\overline{\mathbf{H}}_{u,n_t'}^{(q)})^H]\end{aligned}, \quad (18b)$$

$$2\left(\hat{\sigma}^{(i)}_{u,n_t,ISI}\right)^2 = \frac{2E_{s,u}}{T_s}\frac{\rho^{(i-1)}_{u,n_t}}{M}\left[\sum_{q=1}^{Q} tr[(\mathbf{W}^{(q,i)}_{u,n_t}\overline{\mathbf{H}}^{(q)}_{u,n_t})(\mathbf{W}^{(q,i)}_{u,n_t}\overline{\mathbf{H}}^{(q)}_{u,n_t})^H] - tr[\mathbf{I}]\right], \quad (18c)$$

$$2\left(\hat{\sigma}^{(i)}_{u,n_t,noise}\right)^2 = \frac{2N_0}{T_s}\frac{1}{M}\sum_{q=1}^{Q} tr[\mathbf{W}^{(q,i)}_{u,n_t}(\mathbf{W}^{(q,i)}_{u,n_t})^H], \quad (18d)$$

where [10]

$$\rho^{(i)}_{u,n_t} = E[|D_{u,n_t}(k) - \tilde{D}^{(i)}_{u,n_t}(k)|^2] \approx \frac{1}{M}\sum_{m=0}^{M-1}\left(E[|d_{u,n_t}(m)|^2] - |\tilde{d}^{(i)}_{u,n_t}(m)|^2\right), \quad (19)$$

with $E[.]$ being the expectation operation using the *a posteriori* probabilities, $p^{(i-1)}_{u,n_t}(b_{m,x}=0)$ and $p^{(i-1)}_{u,n_t}(b_{m,x}=1)$, for the given received signal vector.

The soft symbol replica $\tilde{d}^{(i)}_{u,n_t}(m)$ is generated, following to [10], as

$$\begin{cases} \tilde{d}^{(i-1)}_{u,n_t}(m) = \frac{1}{\sqrt{2}}\tanh\left(\frac{\lambda^{(i-1)}_{u,n_t,0}(m)}{2}\right) + j\frac{1}{\sqrt{2}}\tanh\left(\frac{\lambda^{(i-1)}_{u,n_t,1}(m)}{2}\right) \quad \text{for QPSK} \\ \tilde{d}^{(i-1)}_{u,n_t}(m) = \frac{1}{\sqrt{10}}\tanh\left(\frac{\lambda^{(i-1)}_{u,n_t,0}(m)}{2}\right)\left\{2+\tanh\left(\frac{\lambda^{(i-1)}_{u,n_t,1}(m)}{2}\right)\right\} \\ \quad + j\frac{1}{\sqrt{10}}\tanh\left(\frac{\lambda^{(i-1)}_{u,n_t,2}(m)}{2}\right)\left\{2+\tanh\left(\frac{\lambda^{(i-1)}_{u,n_t,3}(m)}{2}\right)\right\} \quad \text{for 16QAM} \end{cases}. \quad (20)$$

The soft decision signal replica vector $\tilde{\mathbf{d}}^{(i-1)}_{u,n_t} = [\tilde{d}^{(i-1)}_{u,n_t}(0),...,\tilde{d}^{(i-1)}_{u,n_t}(M-1)]^T$ is transformed into the frequency-domain soft signal replica vector $\tilde{\mathbf{D}}^{(i-1)}_{u,n_t} = [\tilde{D}^{(i-1)}_{u,n_t}(0),...,\tilde{D}^{(i-1)}_{u,n_t}(M-1)]^T$ by applying M-point DFT as

$$\tilde{\mathbf{D}}^{(i-1)}_{u,n_t} = \mathbf{F}_M\tilde{\mathbf{d}}^{(i-1)}_{u,n_t}, \quad (21)$$

where $\tilde{d}^{(-1)}_{u,n_t}(m)=0$. Interference cancellation is carried out according to Eq. (15). The frequency-domain MUI replica vector $\tilde{\mathbf{M}}^{(i)}_{u,n_t} = [\tilde{M}^{(i)}_{u,n_t}(0),...,\tilde{M}^{(i)}_{u,n_t}(M-1)]^T$, IAI replica vector $\tilde{\mathbf{A}}^{(i)}_{u,n_t} = [\tilde{A}^{(i)}_{u,n_t}(0),...,\tilde{A}^{(i)}_{u,n_t}(M-1)]^T$ and ISI replica vector $\tilde{\mathbf{I}}^{(i)}_{u,n_t} = [\tilde{I}^{(i)}_{u,n_t}(0),...,\tilde{I}^{(i)}_{u,n_t}(M-1)]^T$ are generated as

$$\begin{cases} \tilde{\mathbf{M}}^{(i)}_{u,n_t} = \sum_{q=1}^{Q}\mathbf{W}^{(q,i)}_{u,n_t}\left(\sum_{u'=0\neq u}^{U-1}\sum_{n_t'=0}^{N_t-1}\sqrt{\frac{2E_{s,u'}}{T_s}}\overline{\mathbf{H}}^{(q)}_{u',n_t'}\tilde{\mathbf{D}}^{(i-1)}_{u',n_t'}\right) \\ \tilde{\mathbf{A}}^{(i)}_{u,n_t} = \sqrt{\frac{2E_{s,u}}{T_s}}\sum_{q=1}^{Q}\mathbf{W}^{(q,i)}_{u,n_t}\left(\sum_{n_t'=0\neq n_t}^{N_t-1}\overline{\mathbf{H}}^{(q)}_{u,n_t'}\tilde{\mathbf{D}}^{(i-1)}_{u,n_t'}\right) \\ \tilde{\mathbf{I}}^{(i)}_{u,n_t} = \sqrt{\frac{2E_{s,u}}{T_s}}\left(\sum_{q=1}^{Q}\mathbf{W}^{(q,i)}_{u,n_t}\overline{\mathbf{H}}^{(q)}_{u,n_t} - \mathbf{I}\right)\tilde{\mathbf{D}}^{(i-1)}_{u,n_t} \end{cases}. \quad (22)$$

B. MMSE Weight Matrix

The error vector between the frequency-domain signal vector $\breve{\mathbf{D}}^{(i)}_{u,n_t}$ after interference cancellation and the transmit frequency-domain signal vector $\mathbf{D}_{u,n_t}$ is defined as

$$\mathbf{e}^{(i)}_{u,n_t} = \breve{\mathbf{D}}^{(i)}_{u,n_t} - \sqrt{\frac{2E_{s,u}}{T_s}}\mathbf{D}_{u,n_t}. \quad (23)$$

The MMSE weight matrix is derived which minimizes the trace $trE[\{\mathbf{e}^{(i)}_{u,n_t}\}\{\mathbf{e}^{(i)}_{u,n_t}\}^H]$. The MMSE weight matrix of uth user's n_tth transmit antenna at the ith stage is given by

$$\mathbf{W}^{(q,i)}_{u,n_t} = \arg\min trE[\{\mathbf{e}^{(i)}_{u,n_t}\}\{\mathbf{e}^{(i)}_{u,n_t}\}^H] = \left(\overline{\mathbf{H}}^{q}_{u,n_t}\right)^H\left[\sum_{u'=0}^{U-1}E_{s,u'}\sum_{n_t'=0}^{N_t-1}\rho^{(i-1)}_{u',n_t'}\sum_{q=1}^{Q}\overline{\mathbf{H}}^{q}_{u',n_t'}\left(\overline{\mathbf{H}}^{q}_{u',n_t'}\right)^H + N_0\mathbf{I}\right]^{-1}. \quad (24)$$

IV. COMPUTER SIMULATION

The simulation condition is summarized in Table 1. We consider HARQ with Chase combining. A turbo encoder with two (13,15) RSC encoders and decoders with Log-MAP algorithm are used. The length of the code bit sequence is 2048 bits. The code rate is set to R=1/2.

We assume QPSK or 16QAM data modulation, M=16, N_c=64 and N_g=32 samples GI. Channel is assumed to be an L=16-path frequency-selective block Rayleigh fading channel having uniform power delay profile (i.e., $E[|h^{(q)}_{u,n_t,n_r,l}|^2]=1/L$ for all q, u, n_t, n_r, l). It is assumed that the sum of maximum transmit timing offset among users and channel maximum delay time is less than the GI length. Ideal channel estimation and ideal slow transmit power control ($E_{s,u}$=E_s for all user) are also assumed.

Table. 1 Simulation condition

Channel coding	No. of coded bits	2048
	Encoder	(13,15) RSC
	Coding rate	R=1/2
	Channel interleaver	Block
	Packet combining	Chase Combining
	Decoder	Log-MAP with 10 iterations
Number of Tx. Rx. antennas		N_t=N_r=2
Transmitter	Data modulation	QPSK, 16QAM
	Number of symbols per block	M=16
	FFT/IFFT block size	N_c=64
	Number of users	U=4 (=N_c/M)
	GI length	N_g=32
Transmit/receive filters	Transfer function	Square-root raised cosine
	Roll off factor	α=0~1
Channel	Fading type	Frequency-selective block Rayleigh
	Power delay profile	L=16-path uniform power delay profile
	Time delay	$\tau_{u,l}$=l, l=0,…,L−1
Receiver	Signal detection	MMSE detection
	Channel estimation	Ideal

A. PAPR

PAPR is defined as [11]

$$\text{PAPR} = \frac{\max\{|s_{u,n_t}(t)|^2\}_{t=0\sim N_c-1}}{E[|s_{u,n_t}(t)|^2]}. \quad (25)$$

The PAPR level at the complementary cumulative distribution function (CCDF)=10^{-3} is shown in Table 2 for various values of the roll-off factor α [12]. As α increases, the $\text{PAPR}_{0.1\%}$ level decreases, but it becomes almost the same beyond α=0.5.

Table. 2 PAPR level at CCDF=10^{-3}

		SC-FDMA					OFDMA
α		0	0.25	0.5	0.75	1	
$\text{PAPR}_{0.1\%}$ (dB)	QPSK	7.1	5.06	3.37	3.37	3.59	9.98
	16QAM	7.9	6.76	6.46	6.86	7.2	9.77

B. Throughput performance

Throughput η[bps/Hz] is defined as

$$\eta = N \times N_t \times R \times \frac{1}{\overline{Q}} \times \frac{N_c}{N_c + N_g}, \tag{26}$$

where $\overline{Q} - 1$ is the average number of packet retransmissions.

The throughput performance with IC and without IC for α=0, 1 is plotted as a function of the average received symbol energy-to-noise power spectrum density ratio E_s/N_0 in Fig. 4. For comparison, the throughput performances of OFDMA with IAI cancellation [9] and without IAI cancellation are also plotted.

MMSED-IC improves the throughput performance compared with w/o IC since it can suppress the interference (MUI&IAI&ISI for SC, IAI for OFDMA) while achieving a higher coding gain. By use of MMSED-IC, the throughput of α=1 is higher than that of α=0. While sufficiently suppressing MUI&IAI&ISI, the additional frequency diversity gain can be obtained by exploiting the excess bandwidth introduced by transmit filtering. The throughput of SC with α=1 is superior to that of OFDMA in the region of E_s/N_0>11dB and E_s/N_0<10dB since there is a larger frequency diversity gain for SC transmission.

The peak E_s/N_0 is an important parameter which determines the peak transmit power required for the mobile terminal transmitter power amplifiers. In this paper, peak E_s/N_0 is defined as peak E_s/N_0=average received E_s/N_0 + $\mathrm{PAPR}_{0.1\%}$ [13]. The throughput performance is plotted as a function of peak E_s/N_0 in Fig. 5. It can be seen from Fig. 5 that the throughput performance improvement is more pronounced due to the reduced PAPR.

V. CONCLUSION

In this paper, we proposed an iterative MMSED and IC (MMSED-IC), which performs iteratively a series of MMSE detection and interference (MUI&IAI&ISI) cancellation, for turbo coded uplink SC-FDMA MIMO using HARQ. The PAPR and HARQ throughput performances using the proposed scheme were evaluated by computer simulation. As the filter roll-off factor α increases, the PAPR decreases, and the throughput performance with IC improves since larger frequency diversity gain is obtained. It was shown that SC-FDMA with α=1 using MMSED-IC gives higher or almost same throughput as OFDMA. The throughput improvement is more pronounced when the throughput performance vs. peak E_s/N_0 is considered.

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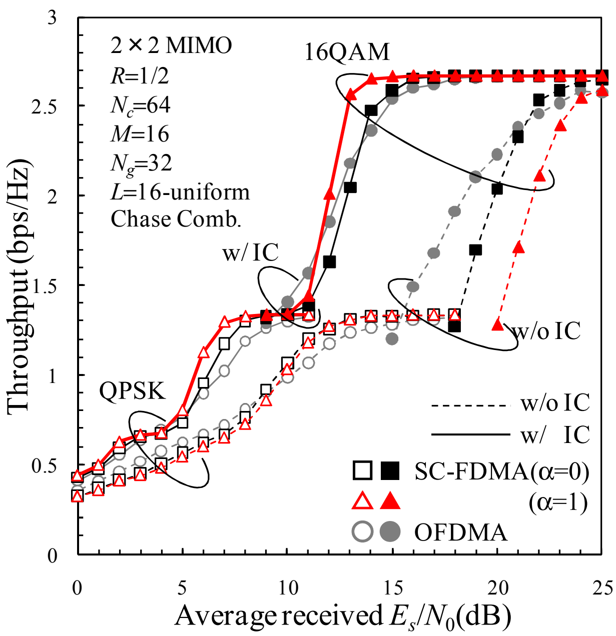

Fig. 4 Throughput vs. average received E_s/N_0

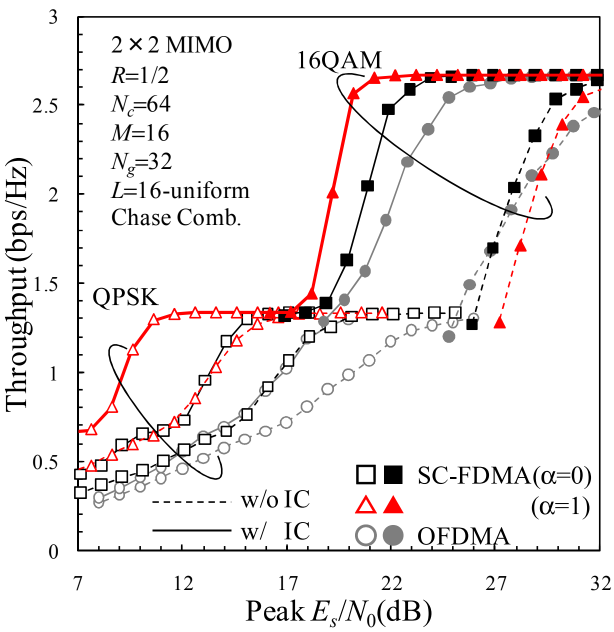

Fig. 5 Throughput vs. peak E_s/N_0