IEICE TRANS. COMMUN., VOL.E93–B, NO.8 AUGUST 2010

PAPER

Frequency-Domain Block Signal Detection for Single-Carrier Transmission

Tetsuya YAMAMOTO†a), Kazuki TAKEDA†, *Student Members*, *and* Fumiyuki ADACHI†, *Fellow*

SUMMARY One-tap frequency-domain equalization (FDE) based on the minimum mean square error (MMSE) criterion can significantly improve the bit error rate (BER) performance of single-carrier (SC) transmission in a frequency-selective fading channel. However, a big performance gap from the theoretical lower bound still exists due to the presence of residual inter-symbol interference (ISI) after MMSE-FDE. In this paper, we point out that the frequency-domain received SC signal can be expressed using the matrix representation similar to the multiple-input multiple-output (MIMO) multiplexing and therefore, signal detection schemes developed for MIMO multiplexing, other than simple one-tap MMSE-FDE, can be applied to SC transmission. Then, for the reception of SC signals, we propose a new signal detection scheme, which combines FDE with MIMO signal detection, such as MMSE detection and Vertical-Bell Laboratories layered space-time architecture (V-BLAST) detection (we call this frequency-domain block signal detection). The achievable average BER performance using the proposed frequency-domain block signal detection is evaluated by computer simulation.

key words: *single-carrier, frequency-domain equalization, MMSE, V-BLAST*

1. Introduction

In next generation mobile communication systems, broadband data services are demanded. Since the mobile wireless channel is composed of many propagation paths with different time delays, the channel becomes severely frequency-selective. In a severe frequency-selective fading channel, the bit error rate (BER) performance significantly degrades due to inter-symbol interference (ISI) when single carrier (SC) transmission without equalization is used [1]. A simple one-tap frequency-domain equalization (FDE) based on the minimum mean square error (MMSE) criterion can take advantage of the channel frequency-selectivity and significantly improve the BER performance of SC transmission in a strong frequency-selective fading channel [2], [3]. However, a big performance gap from the theoretical lower bound still exists due to the presence of residual ISI after FDE [4]. Some advanced signal detection schemes combined with equalization are thus necessary. A frequency-domain iterative ISI cancellation technique combined with MMSE-FDE was proposed in [4]. However, the achievable BER performance is still a few dB away from the theoretical lower bound. In this paper, we take another approach to improve the BER performance in a strong frequency-selective fading channel.

We first point out that the frequency-domain received SC signal is expressed using the matrix representation similar to the multiple-input multiple-output (MIMO) multiplexing and therefore, signal detection schemes developed for MIMO multiplexing, other than simple one-tap MMSE-FDE, can be applied to the SC transmissions. Among the well-known signal detection schemes for MIMO multiplexing are the MMSE detection [1] and the Vertical-Bell Laboratories layered space-time architecture (V-BLAST) detection [5]. Therefore, we combine equalization with both MMSE detection and V-BLAST detection.

We propose a new signal detection scheme, which combines FDE with MIMO signal detection, for the reception of SC signals. In the proposed signal detection combined with V-BLAST detection, a combination of ISI cancellation and MMSE detection taking into account by the residual ISI is carried out successively symbol by symbol within a block unlike the conventional MMSE-FDE. To clearly indicate this difference, we call our proposed signal detection method as the frequency-domain block signal detection. We use iterative detection [6] to further improve the performance of V-BLAST detection and two types of iterative V-BLAST detection are considered. The first (called the hard decision iterative V-BLAST in this paper) uses the hard symbol replica which is obtained from the hard decision result. The second (called the soft decision iterative V-BLAST in this paper) uses the soft symbol replica which is generated based on the log-likelihood ratio (LLR). In Ref. [6], iterative parallel interference cancellation (PIC) using soft symbol replica was proposed. On the other hand, our proposed block signal detection used iterative successive interference cancellation (SIC) based on V-BLAST. In Ref. [7], iterative SIC was proposed for the signal transmissions using SC-MIMO spatial multiplexing. The MMSE weight taking into account the residual inter-antenna interference (IAI) after interference cancellation was derived. However, the residual ISI from the own antenna was not considered in the MMSE weight. In this paper, we derive the MMSE weight taking into account the residual ISI, after interference cancellation.

The remainder of this paper is organized as follows. Sect. 2 presents the SC signal representation. In Sect. 3, our proposed frequency-domain block signal detection which combines equalization and iterative V-BLAST detection is described. In Sect. 4, the achievable average BER performance using a new frequency-domain block signal detection, which combines equalization and iterative V-BLAST

Manuscript received June 23, 2009.
Manuscript revised March 8, 2010.

†The authors are with the Department of Electrical and Communication Engineering, Graduate School of Engineering, Tohoku University, Sendai-shi, 980-8579 Japan.
a) E-mail: yamamoto@mobile.ecei.tohoku.ac.jp

DOI: 10.1587/transcom.E93.B.2104

detection in a frequency-selective fading channel is evaluated by computer simulation. Sect. 5 offers the conclusion.

2. Signal Representation

The SC transmission model is illustrated in Fig. 1. Throughout the paper, the symbol-spaced discrete time representation is used. At the transmitter, a binary information sequence is data-modulated and then, the data-modulated symbol sequence is divided into a sequence of signal blocks of N_c symbols each, where N_c is the size of fast Fourier transform (FFT). The data symbol block can be expressed using the vector form as $\mathbf{d} = [d(0), \ldots, d(N_c-1)]^T$. Then, the last N_g symbols of each block are copied as a cyclic prefix and inserted into the guard interval (GI) placed at the beginning of each block and a sequence of signal blocks is transmitted.

Each transmitted signal block is assumed to pass through a frequency-selective fading channel composed of L distinct paths. The channel impulse response $h(\tau)$ is given by

$$h(\tau) = \sum_{l=0}^{L-1} h_l \delta(\tau - \tau_l), \tag{1}$$

where h_l and τ_l are respectively the complex-valued path gain with $E\left[\sum_{l=0}^{L-1}|h_l|^2\right] = 1$ and the time delay of the lth path. The GI-removed received signal block $\mathbf{r} = [r(0), \ldots, r(N_c-1)]^T$ can be expressed using the vector form as

$$\mathbf{r} = \sqrt{\frac{2E_s}{T_s}}\mathbf{hd} + \mathbf{n}, \tag{2}$$

where E_s and T_s are respectively the symbol energy and the symbol duration, $\mathbf{h}$ is the $N_c \times N_c$ channel impulse response matrix given as

$$\mathbf{h} = \begin{bmatrix} h_0 & & & & h_{L-1} & & \\ \vdots & h_0 & & & & \ddots & \\ & \vdots & h_0 & \mathbf{0} & & & h_{L-1} \\ h_{L-1} & & \vdots & \ddots & & & \\ & h_{L-1} & & & h_0 & & \\ & & h_{L-1} & & \vdots & \ddots & \\ \mathbf{0} & & & \ddots & & & h_0 \end{bmatrix}, \tag{3}$$

and $\mathbf{n} = [n(0), \ldots, n(N_c-1)]^T$ is the noise vector. The tth element, $n(t)$, of $\mathbf{n}$ is the zero-mean additive white Gaussian noise (AWGN) having variance $2N_0/T_s$ with N_0 being the one-sided noise power spectrum density.

At the receiver, N_c-point FFT is applied to transform the received signal block into the frequency-domain signal vector $\mathbf{R} = [R(0), \ldots, R(N_c-1)]^T$. $\mathbf{R}$ is expressed as

$$\mathbf{R} = \mathbf{Fr} = \sqrt{\frac{2E_s}{T_s}}\mathbf{Fhd} + \mathbf{Fn}, \tag{4}$$

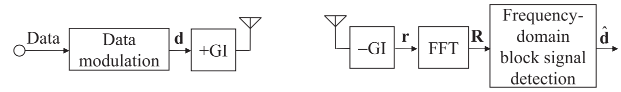


Fig. 1 SC transmission system model.

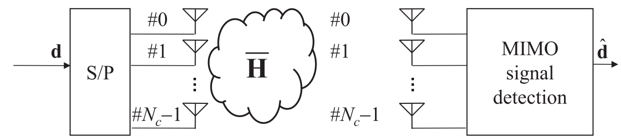


Fig. 2 Equivalent system model.

where $\mathbf{F}$ is the FFT matrix of size $N_c \times N_c$ given by

$$\mathbf{F} = \frac{1}{\sqrt{N_c}} \begin{bmatrix} e^{-j2\pi\frac{0\times 0}{N_c}} & e^{-j2\pi\frac{0\times 1}{N_c}} & \cdots & e^{-j2\pi\frac{0\times(N_c-1)}{N_c}} \\ e^{-j2\pi\frac{1\times 0}{N_c}} & e^{-j2\pi\frac{1\times 1}{N_c}} & \cdots & e^{-j2\pi\frac{1\times(N_c-1)}{N_c}} \\ \vdots & \vdots & \ddots & \vdots \\ e^{-j2\pi\frac{(N_c-1)\times 0}{N_c}} & e^{-j2\pi\frac{(N_c-1)\times 1}{N_c}} & \cdots & e^{-j2\pi\frac{(N_c-1)\times(N_c-1)}{N_c}} \end{bmatrix}. \tag{5}$$

Since the channel impulse response matrix $\mathbf{h}$ is a circulant matrix, the eigenvalue decomposition using $\mathbf{F}$ can be applied [7]. We have

$$\mathbf{FhF}^H = diag\,[H(0), \ldots, H(N_c-1)] \equiv \mathbf{H}, \tag{6}$$

where $H(k) = \sum_{l=0}^{L-1} h_l \exp(-j2\pi k\tau_l/N_c)$, $k = 0 \sim N_c - 1$, and $(.)^H$ is the Hermitian transpose operation. Using Eq. (6), Eq. (4) can be rewritten as

$$\mathbf{R} = \sqrt{\frac{2E_s}{T_s}}\mathbf{HFd} + \mathbf{N} = \sqrt{\frac{2E_s}{T_s}}\bar{\mathbf{H}}\mathbf{d} + \mathbf{N}, \tag{7}$$

where $\bar{\mathbf{H}} = \mathbf{HF}$ and $\mathbf{N} = [N(0), \ldots, N(N_c-1)]^T$ are respectively the equivalent channel matrix and the frequency-domain noise vector. From Eq. (7), it can be understood that the frequency-domain received SC signal can be treated as a received signal in MIMO multiplexing using N_c transmit antennas and N_c receive antennas with the channel matrix $\bar{\mathbf{H}}$ (see Fig. 2). According to this understanding, a new frequency-domain block signal detection scheme, which combines equalization and MIMO signal detection such as such as MMSE detection and V-BLAST detection, can be developed for the reception of the SC signals. We can show that the frequency-domain block signal detection combined with MMSE detection using the equivalent channel matrix $\mathbf{HF}$ is identical to the conventional one-tap MMSE-FDE. In the next section, therefore, we consider the V-BLAST detection only.

3. Frequency-Domain Block Signal Detection

Frequency-domain block signal detection combined with iterative V-BLAST detection is illustrated in Fig. 3. V-BLAST [5] uses interference cancellation and is composed of i) ordering, ii) interference cancellation, and iii) signal detection. The transmitted symbol which has the highest signal-to-interference plus noise power ratio (SINR) among

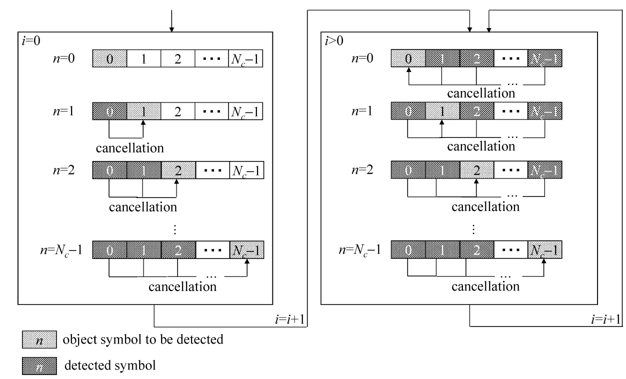


Fig. 3 Frequency-domain block signal detection combined with iterative V-BLAST detection.

undetected symbols is detected by performing MMSE detection [9]. However, in the case of SC transmission, since the SINR is the same for all symbols, signal detection can be carried out simply from the first symbol (i.e., $d(0)$). Then, the replica of the symbol, which has been just detected, is generated and is subtracted from $\mathbf{R}$. A new MMSE weight matrix for the undetected symbols is computed again and one of these symbols is detected.

The above symbol detection and interference cancellation for the reception of the nth symbol $d(n)$ is called the nth layer and is repeated until all of the transmitted symbols are detected. However, since V-BLAST cannot suppress sufficiently the ISI, V-BLAST is iterated a sufficient number of times. In what follows, the signal detection in the nth layer ($n = 0 \sim N_c - 1$) of the ith iteration stage is presented.

3.1 Frequency-Domain Block Signal Detection Combined with Hard Decision Iterative V-BLAST

In the nth layer of the ith iteration stage, the hard replica $\hat{s}^{(i)}(n-1)$ of the symbol $d(n-1)$, which has been just detected in the previous layer, is generated. The frequency-domain received signal vector $\tilde{\mathbf{R}}^{(i,n)} = \left[\tilde{R}^{(i,n)}(0), \ldots, \tilde{R}^{(i,n)}(N_c - 1)\right]^T$ in the nth layer of the ith iteration stage is given by

$$\tilde{\mathbf{R}}^{(i,n)} = \mathbf{R} - \sqrt{\frac{2E_s}{T_s}}\bar{\mathbf{H}}\hat{\mathbf{s}}^{(i,n)}, \tag{8}$$

where $\hat{\mathbf{s}}^{(i,n)}$ is the hard replica vector. $\hat{\mathbf{s}}^{(i,n)}$ is given by $\hat{\mathbf{s}}^{(i,n)} = \left[\hat{s}^{(i)}(0), \ldots, \hat{s}^{(i)}(n-1), 0, \hat{s}^{(i-1)}(n+1), \ldots, \hat{s}^{(i-1)}(N_c - 1)\right]^T$, where $\{\hat{s}^{(i)}(n'); n' = 0 \sim n-1\}$ are generated using the decision in the present iteration stage while $\{\hat{s}^{(i-1)}(n'); n' = n+1 \sim N_c - 1\}$ are generated using the decision results in the $(i-1)$th iteration stage. After interference cancellation, MMSE detection on the nth symbol is performed by multiplying $\tilde{\mathbf{R}}^{(i,n)}$ by an $1 \times N_c$ MMSE weight vector $\mathbf{W}^{(i,n)}$ as

$$\tilde{d}^{(i)}(n) = \mathbf{W}^{(i,n)}\tilde{\mathbf{R}}^{(i,n)}. \tag{9}$$

The MMSE weight vector $\mathbf{W}^{(i,n)}$ is given as

$$\mathbf{W}^{(i,n)} = \bar{\mathbf{H}}^H\left[\tilde{\mathbf{H}}\tilde{\mathbf{H}}^H + \left(\frac{E_s}{N_0}\right)^{-1}\mathbf{I}_{N_c}\right]^{-1}, \tag{10}$$

where

$$\tilde{\mathbf{H}} = \begin{cases} \left[\bar{\mathbf{H}}_n, \bar{\mathbf{H}}_{n+1}, \ldots, \bar{\mathbf{H}}_{N_c-1}\right] & \text{for } i = 0 \\ \bar{\mathbf{H}}_n & \text{for } i > 0 \end{cases} \tag{11}$$

and $\bar{\mathbf{H}}_n$ is the nth column vector of $\bar{\mathbf{H}}$. In the $i = 0$th iteration stage, we remove the column vectors, $\bar{\mathbf{H}}_0$, $\bar{\mathbf{H}}_1$, …, $\bar{\mathbf{H}}_{n-1}$, associated with already detected symbols from $\bar{\mathbf{H}}$ assuming that these symbols have been cancelled perfectly. On the other hand, in the $i(i > 0)$th iteration stage, assuming that all of symbols have been cancelled perfectly, we remove all column vectors $\bar{\mathbf{H}}_0$, …, $\bar{\mathbf{H}}_{n-1}$, $\bar{\mathbf{H}}_{n+1}$, …, $\bar{\mathbf{H}}_{N_c-1}$ from $\bar{\mathbf{H}}$; therefore, the weight vector described by Eq. (15) become the maximal ratio combining FDE (MRC-FDE) weight.

Until all of the transmitted symbols are detected, interference cancellation and MMSE detection are repeated. After all the transmitted symbols are detected, the next iteration is carried out. The above iteration is done a sufficient number of times.

3.2 Frequency-Domain Block Signal Detection Combined with Soft Decision Iterative V-BLAST

So far, we have assumed hard cancellation. The use of soft replica reduces the influence of error propagation. The soft replica can be generated using the LLR [4]. Using the soft decision variable $\tilde{d}^{(i)}(n)$ associated with $d(n)$, the LLR for the xth bit, $x = 0 \sim \log_2 M - 1$, in the nth symbol, $n = 0 \sim N_c - 1$, is obtained where M is the modulation level. The soft replica $\tilde{s}^{(i)}(n)$ is generated using the LLR as [10]

$$\tilde{s}^{(i)}(n) = \begin{cases} \dfrac{1}{\sqrt{2}}\tanh\left(\dfrac{\lambda_0^{(i)}(n)}{2}\right) + j\dfrac{1}{\sqrt{2}}\tanh\left(\dfrac{\lambda_1^{(i)}(n)}{2}\right) \\ \hfill \text{for QPSK} \\ \dfrac{1}{\sqrt{10}}\tanh\left(\dfrac{\lambda_0^{(i)}(n)}{2}\right)\left\{2 + \tanh\left(\dfrac{\lambda_1^{(i)}(n)}{2}\right)\right\} \\ +j\dfrac{1}{\sqrt{10}}\tanh\left(\dfrac{\lambda_2^{(i)}(n)}{2}\right)\left\{2 + \tanh\left(\dfrac{\lambda_3^{(i)}(n)}{2}\right)\right\} \\ \hfill \text{for 16QAM.} \end{cases} \tag{12}$$

The LLR can be computed as [11]

$$\begin{aligned} \lambda_x^{(i)}(n) &= \ln\left(\frac{p(b_{n,x} = 1)}{p(b_{n,x} = 0)}\right) \\ &\approx \frac{1}{2\hat{\sigma}_n^{(i)^2}}\left\{\begin{aligned} &\left|\tilde{d}^{(i)}(n) - \sqrt{\frac{2E_s}{T_s}}\hat{H}_n d^{\text{mim}}_{b_{x,n}=0}\right|^2 \\ &-\left|\tilde{d}^{(i)}(n) - \sqrt{\frac{2E_s}{T_s}}\hat{H}_n d^{\text{mim}}_{b_{x,n}=1}\right|^2 \end{aligned}\right\}, \end{aligned} \tag{13}$$

where $p(b_{n,x} = 1)$ and $p(b_{n,x} = 0)$ are *a posteriori* probabilities of the transmitted bit $b_{n,x} = 1$ and 0, respectively. $d^{\text{mim}}_{b_{x,n}=0}$ (or $d^{\text{mim}}_{b_{x,n}=1}$) is the most probable symbol that gives the minimum Euclidean distance from $\tilde{d}^{(i)}(n)$ among all candidate symbols with $b_{n,x} = 0$ (or 1). $\hat{H}_n$ is the nth element of $\hat{\mathbf{H}} = \mathbf{W}^{(i,n)}\bar{\mathbf{H}}$. $2\hat{\sigma}_n^{(i)^2}$ is the variance of the noise plus residual ISI and is given by [12]

$$2\hat{\sigma}_n^{(i)^2} = \frac{2N_0}{T_s}\left[\begin{array}{l}\left\|\mathbf{W}^{(i,n)}\right\|^2 \\ +\dfrac{E_s}{N_0}\left\{\sum_{n'=0}^{n-1}\rho_{n'}^{(i)}\left|\hat{H}_{n'}\right|^2 + \sum_{n'=n+1}^{N_c-1}\rho_{n'}^{(i-1)}\left|\hat{H}_{n'}\right|^2\right\}\end{array}\right], \quad (14)$$

where $\rho_n^{(i)}$ indicates the extent to which the residual ISI remains and is given by [10]

$$\begin{aligned}\rho_n^{(i)} &= E\left[\left|d(n) - \tilde{s}^{(i)}(n)\right|^2\right] \\ &= \begin{cases} 1 - \left|\tilde{s}^{(i)}(n)\right|^2 & \text{for QPSK} \\ \dfrac{4}{10}\tanh\left(\dfrac{\lambda_1^{(i)}(n)}{2}\right) + \dfrac{4}{10}\tanh\left(\dfrac{\lambda_3^{(i)}(n)}{2}\right) & \\ +1 - \left|\tilde{s}^{(i)}(n)\right|^2 & \text{for 16QAM}\end{cases} \quad (15)\end{aligned}$$

After the replica of the symbol, which has been just detected, is generated, interference cancellation using soft replica vector $\tilde{\mathbf{s}}^{(i,n)} = \left[\tilde{s}^{(i)}(0), \ldots, \tilde{s}^{(i)}(n-1), 0, \tilde{s}^{(i-1)}(n+1), \ldots, \tilde{s}^{(i-1)}(N_c-1)\right]^T$ in the nth layer of the ith iteration stage is carried out as in Eq. (13). After interference cancellation, MMSE detection on the nth symbol is performed by multiplying $\tilde{\mathbf{R}}^{(i,n)}$ by an $1 \times N_c$ MMSE weight vector $\mathbf{W}^{(i,n)}$ as in Eq. (14).

When hard replica is used, by assuming the interference from the symbols which have been detected is cancelled perfectly, the weight vector is generated by using $\tilde{\mathbf{H}}^{(i,n)}$ in the nth layer of the ith iteration stage. On the other hand, when soft replica is used, the MMSE weight vector can be updated by taking into account the residual ISI in each layer. In [7], iterative SIC was proposed for the signal transmissions using SC-MIMO spatial multiplexing. The MMSE weight taking into account the residual IAI after interference cancellation was derived, but, the residual ISI from the own antenna was not considered. In this paper, we derive the MMSE weight taking account the residual ISI after interference cancellation for our proposed frequency-domain block signal detection.

The MMSE weight vector taking into account the residual ISI in the nth layer of the ith iteration stage is given by

$$\mathbf{W}^{(i,n)} = \bar{\mathbf{H}}^H\left[\bar{\mathbf{H}}\boldsymbol{\rho}^{(i,n)}\bar{\mathbf{H}}^H + \left(\frac{E_s}{N_0}\right)^{-1}\mathbf{I}_{N_c}\right]^{-1}, \quad (16)$$

where $\boldsymbol{\rho}^{(i,n)} = diag\left[\rho_0^{(i)}, \ldots, \rho_{N_c-1}^{(i)}\right]$ is an $N_c \times N_c$ diagonal matrix which represents the impact of the residual ISI. $\rho_n^{(i)} = 1$ means the residual ISI is kept intact and $\rho_n^{(i)} = 0$ means the residual ISI is sufficiently reduced. Setting $\rho_n^{(i)}$ to 1, the nth symbol is detected. When $i = 0$, the symbols with $n' = n \sim (N_c - 1)$ are undetected and hence, $\rho_{n'}^{(i)}$ becomes 1.

By using soft replica, the influence of error propagation can be reduced and the MMSE weight which takes into account the impact of the residual ISI can be used. Therefore, soft decision iterative V-BLAST can much improve the BER performance compared to the hard decision iterative V-BLAST. However, soft decision iterative V-BLAST always requires $N_c \times N_c$ matrix inversion in all layers of each iteration stage. On the other hand, hard decision iterative V-BLAST requires matrix inversion only in the first iteration stage (much smaller size of $(N_c - n) \times (N_c - n)$ matrix inversion in the nth layer of first iteration stage). Therefore, the hard decision iterative V-BLAST has reduced computational complexity compared to the soft decision iterative V-BLAST.

4. Computer Simulation

The condition for the computer simulation is shown in Table 1. We assume an FFT block size of $N_c = 64$ symbols and a GI of $N_g = 16$ symbols. The channel is assumed to be a symbol-spaced $L = 16$-path frequency-selective block Rayleigh fading channel having an exponential power delay profile with the decay factor α. Ideal channel estimation is assumed.

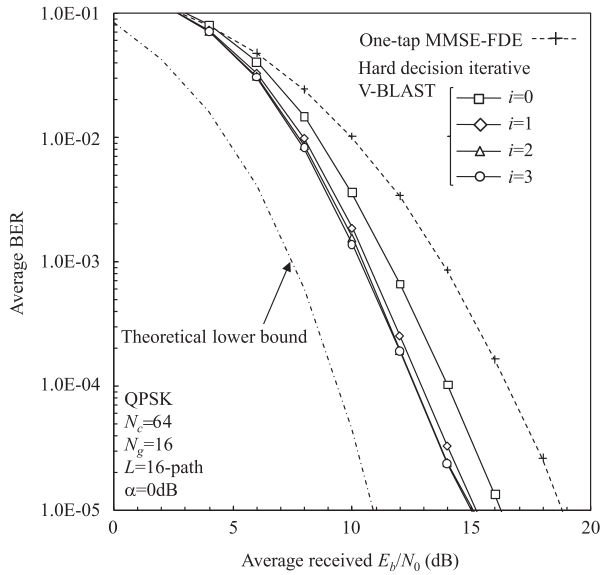
4.1 Frequency-Domain Block Signal Detection Combined with Hard Decision Iterative V-BLAST

The average BER performance of SC transmission using frequency-domain block signal detection combined with hard decision iterative V-BLAST detection is plotted in Fig. 4 as a function of average received bit energy-to-noise power spectrum density ratio $E_b/N_0(= (E_s/N_0)(1 + N_g/N_c)/\log_2 M)$. Also plotted for comparison are the BER performances achievable by the one-tap MMSE-FDE and the theoretical lower bound [13].

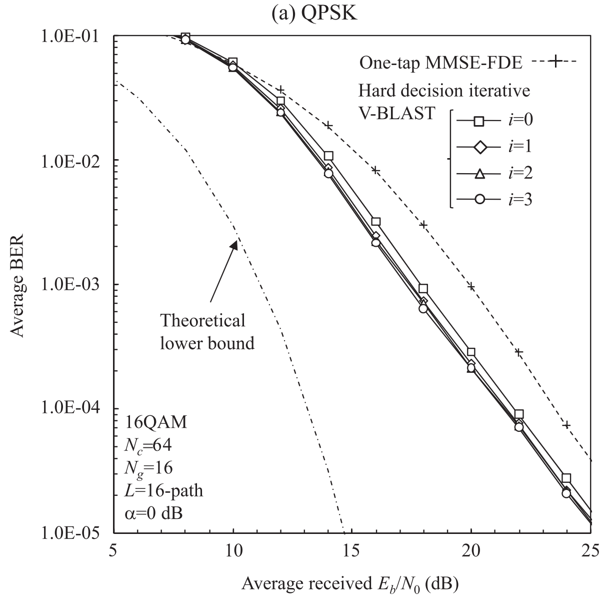
When MMSE-FDE is used, a big BER performance gap from the lower bound still exists due to the residual ISI. For QPSK (16QAM) data modulation, the gap in the required E_b/N_0 for the average BER $= 10^{-4}$ is 7.2 (10.5) dB including the GI insertion loss of 0.97 dB. On the other hand, frequency-domain block signal detection combined with V-BLAST detection can achieve better BER performance than the one-tap MMSE-FDE. It can be seen from Fig. 4 that frequency-domain block signal detection combined with hard decision iterative V-BLAST detection can improve the BER performance compared to non iterative V-BLAST detection ($i = 0$). The reason for this is discussed

Table 1 Computer simulation condition.

Transmitter	Modulation	QPSK, 16QAM
	Block size	$N_c = 64$
	GI	$N_g = 16$
Channel	Fading type	Frequency-selective block Rayleigh
	Power delay profile	$L = 16$-path exponential power delay profile
	Decay factor	$\alpha = 0 \sim 12$dB
	Time delay	$\tau_l = l(l = 0 \sim L-1)$
Receiver	Channel estimation	Ideal

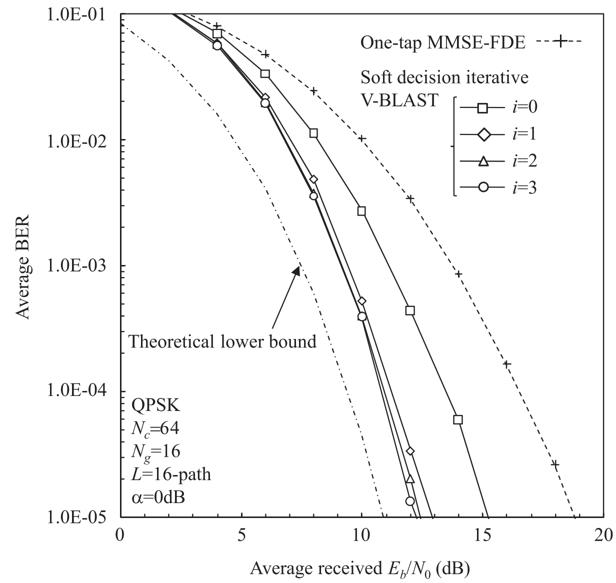


(a) QPSK

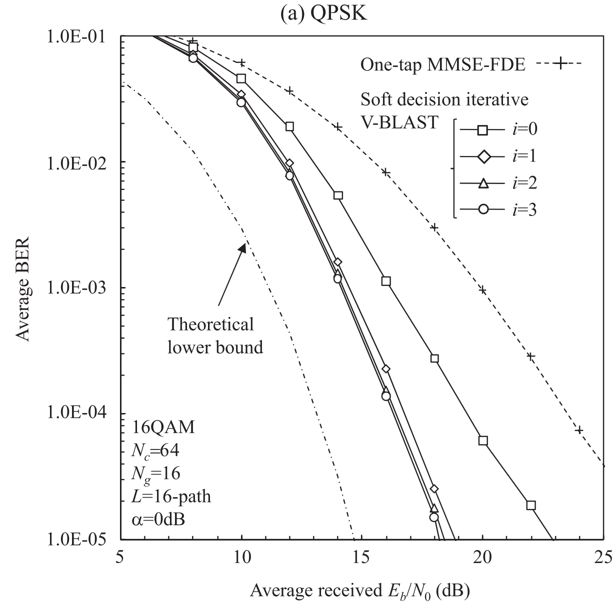


(b) 16QAM

Fig. 4 BER performances using frequency-domain block signal detection combined with MMSE detection or hard decision iterative V-BLAST.



(a) QPSK



(b) 16QAM

Fig. 5 BER performances using frequency-domain block signal detection combined with MMSE detection or soft decision iterative V-BLAST.

below. We focus our attention on the detection of $d(n)$. When $i = 0$, the interference from $d(n+1) \sim d(N_c - 1)$ still remain. On the other hand, when $i > 0$, the interference from $d(n+1) \sim d(N_c - 1)$ can be cancelled. Therefore, iterative detection can suppress the ISI sufficiently for the symbols with small indices. For QPSK (16QAM), the E_b/N_0 gap from the lower bound for the average BER $= 10^{-4}$ reduces by 4.5(8.5) and 3.7(8.4) dB when $i = 0$ and 1, respectively. However, the use of more than two iterations ($i > 1$) improve the BER performance only slightly due to the error propagation.

4.2 Frequency-Domain Block Signal Detection Combined with Soft Decision Iterative V-BLAST

The average BER performance of SC transmission using frequency-domain block signal detection combined with soft decision iterative V-BLAST detection is plotted in Fig. 5 as a function of average received E_b/N_0. Also plotted are the BER performances achievable by the one-tap MMSE-FDE and the theoretical lower bound.

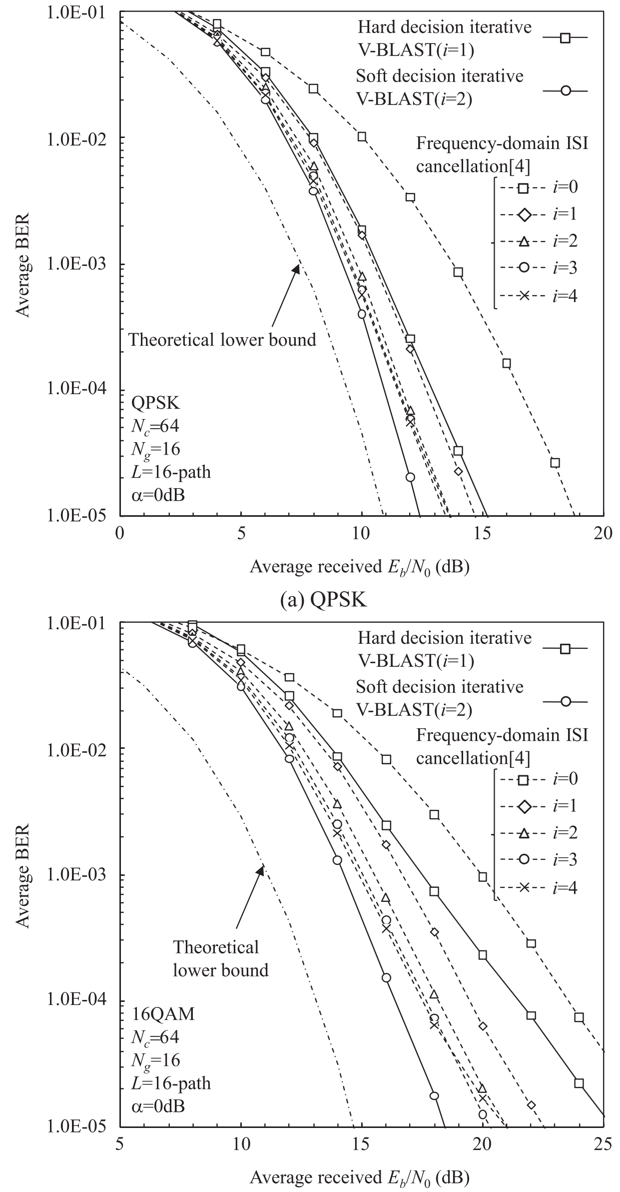
It can be seen from Fig. 5 that as the number of iterations increases, the BER performance improves and approaches that of the lower bound. This is because the influence of error propagation can be reduced by using soft replica and the MMSE weight which takes into account the residual ISI. For QPSK (16QAM), the E_b/N_0 gap from the lower bound for the average BER $= 10^{-4}$ reduces by 4.0(6.2), 1.8(3.6) and 1.5(3.2) dB when $i = 0$, 1 and 2, respectively. It can be seen that two iteration ($i = 2$) provides sufficiently improved BER performance. Compared to the hard decision iterative V-BLAST in Sect. 4.1, the soft decision iterative V-BLAST achieves smaller required E_b/N_0 by 0.5(3.3) and 2.0(5.2) dB when $i = 0$ and 2, respectively, for QPSK (16QAM).

4.3 Performance Comparison

Figure 6 shows the BER performance comparison between the frequency-domain block signal detection combined with iterative V-BLAST. For comparison, the BER performance of the frequency-domain ISI cancellation [4] is plotted for $i = 0$ (MMSE-FDE), 1, 2, 3, and 4; the use of three iteration ($i = 3$) provides sufficiently improved BER performance. For iterative V-BLAST detection, the number of iterations which provide sufficiently improved BER performance is used. $i = 1$ is used for the hard decision iterative V-BLAST, $i = 2$ is used for the soft decision iterative V-BLAST. It can be seen from Fig. 6 that the proposed frequency-domain block signal detection combined with soft decision iterative V-BLAST provides the best BER performance. In the case of QPSK (16QAM), the frequency-domain block signal detection combined with soft decision iterative V-BLAST can reduce the required average E_b/N_0 for an average BER $= 10^{-4}$ by 0.4(1.3) dB compared to the conventional frequency-domain ISI cancellation with $i = 3$.

Figure 7 shows the influence of the channel frequency-selectivity (represented by the decay factor α) on the achievable BER for the proposed frequency-domain block signal detection combined with iterative V-BLAST detection ($i = 1$ for hard decision and $i = 2$ for soft decision), MMSE-FDE, and frequency-domain ISI cancellation ($i = 3$). The sufficient number i of iterations is used for each signal detection method. It can be seen from Fig. 7 that the BER decreases with α for all signal detection method because larger frequency diversity gain can be obtained. However, the proposed frequency-domain block signal detection combined with soft decision iterative V-BLAST provides the smallest BER regardless of the value of α.

Figure 8 shows the influence of block size N_c on the achievable BER for the proposed frequency-domain block signal detection combined with iterative V-BLAST detection, MMSE-FDE, and frequency-domain ISI cancellation. As the block size N_c increases, the number of orthogonal frequency-components increases with the block size N_c. As the number of orthogonal frequency-components increases, the achievable BER performance improves because larger frequency diversity gain is obtained, but the performance improvement becomes saturated for a large block size because the achievable maximum diversity order of the channel is equal to the number L of paths. Consequently, the achievable BER performance and the computational complexity depend on the block size. Below, we discuss the computational complexity.



(a) QPSK
(b) 16QAM

Fig. 6 BER performance comparison between frequency-domain block signal detection combined with iterative V-BLAST and conventional frequency-domain ISI cancellation [4].

4.4 Complexity

Figure 9 compares the computational complexities for the frequency-domain block signal detection combined with iterative V-BLAST and conventional frequency-domain ISI cancellation in case of 16QAM. The complexity here is defined as the number of complex multiply operations required in the signal detection. The number of iterations is set to $i = 1$, 2, and 3 for the hard decision iterative V-BLAST, for the soft decision iterative V-BLAST, and frequency-domain

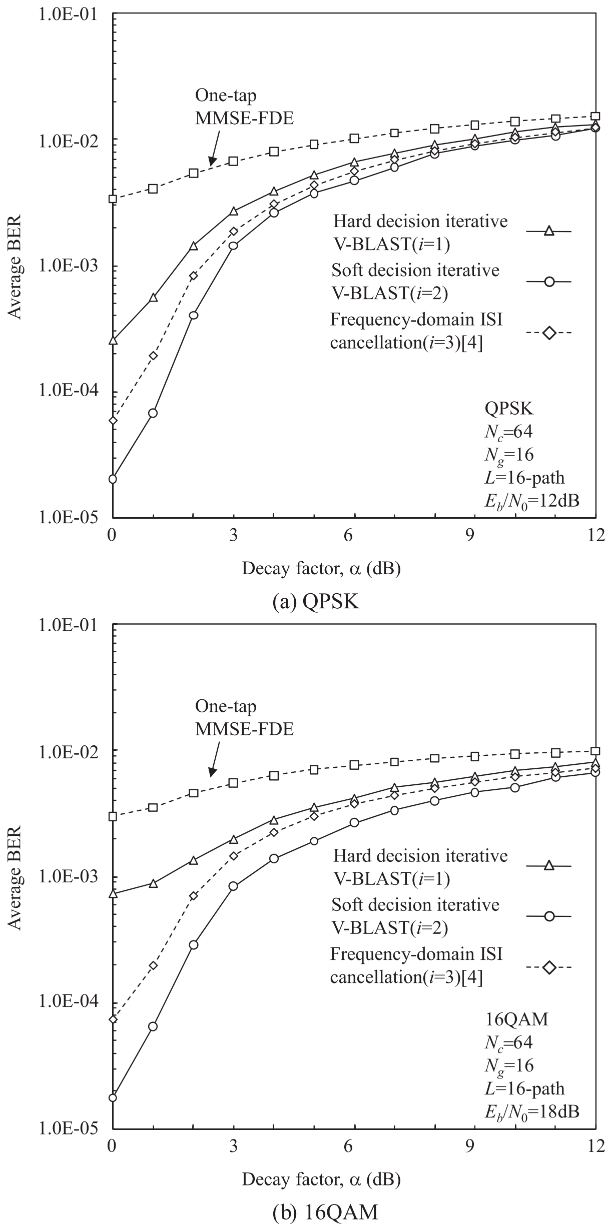


Fig. 7 Influence of frequency-selectivity.

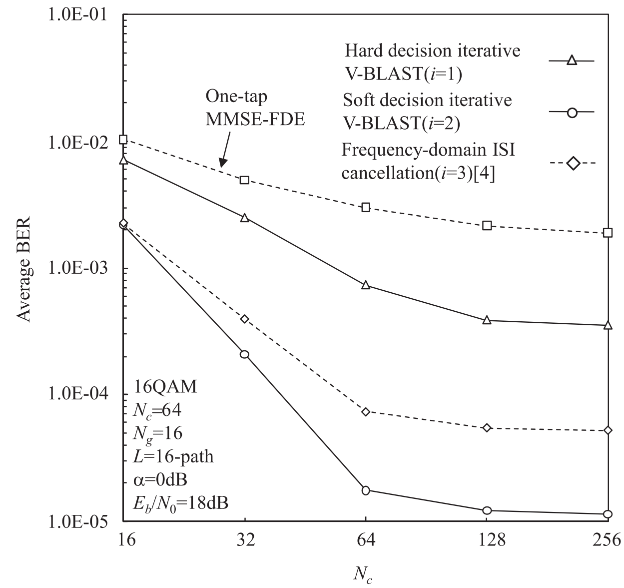


Fig. 8 Influence of block size, N_c.

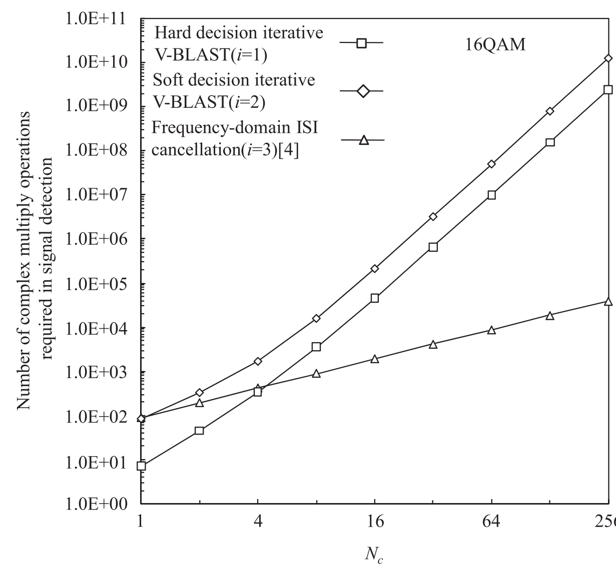


Fig. 9 Complexity comparison.

ISI cancellation, respectively. The number of multiply operations per FFT block is given in Table 2. The soft decision iterative V-BLAST requires $(i + 1) \times N_c$ times matrix inversion of an $N_c \times N_c$ matrix and the total number of multiply operations is $(i + 1)N_c^4$. On the other hand, hard decision iterative V-BLAST requires $(N_c - n) \times (N_c - n)$ times matrix inversion in the nth layer of first iteration stage only and therefore, requires much less computational complexity. The computational complexity of frequency-domain block signal detection combined with hard decision iterative V-BLAST is about 20% of frequency-domain block signal detection combined with soft decision iterative V-BLAST when $N_c = 64$. The computational complexity of frequency-domain block signal detection combined with soft decision iterative V-BLAST is about 6000 times higher than that of the conventional ISI cancellation when $N_c = 64$. The frequency-domain block signal detection combined with soft decision iterative V-BLAST can improve the BER performance at the cost of increased complexity compared to the conventional ISI cancellation.

In this paper, we did not consider any computation reduction algorithm for the matrix inversion required for iterative V-BLAST. Various complexity reduction algorithms have been studied [13]–[15]. Reducing the complexity of the proposed frequency-domain block signal detection is left as an important future study item.

Table 2 Number of multiply operations per block.

	Hard decision iterative V-BLAST	Soft decision iterative V-BLAST	Frequency-domain ISI cancellation [4]
FFT/IFFT	$N_c \log_2 N_c$	$N_c \log_2 N_c$	$2(i+1)N_c \log_2 N_c$
Weight computation	$\sum_{n=0}^{N_c-1} \left[(N_c-n)^3 + N_c(N_c-n)^2 + N_c(N_c-n)\right]$	$(i+1)N_c^4 + (i+2)N_c^3 + (i+1)N_c^2$	$(2i+1)N_c$
Wight multiplication	$(i+1)N_c^2$	$(i+1)N_c^2$	$(i+1)N_c$
Replica generation	$(i+1)N_c^2 - N_c$	$(i+1)N_c^3 + (i+4)N_c^2 + 3(i+15)N_c$	$i(25N_c+17) + N_c$

5. Conclusion

In this paper, we pointed out that the frequency-domain received SC signal can be expressed using the matrix representation similar to the MIMO multiplexing and therefore, signal detection schemes developed for MIMO multiplexing can be applied to SC transmission. Accordingly, we proposed a new frequency-domain block signal detection, which combines FDE with MIMO signal detection, for the reception of SC signals. We evaluated, by computer simulation, the BER performance of SC transmissions using frequency-domain block signal detection combined with iterative V-BLAST detection in a frequency-selective block Rayleigh fading channel.

The frequency-domain block signal detection combined with iterative V-BLAST detection can reduce the residual ISI and hence, achieve better BER performance than the one-tap MMSE-FDE. Further-more, the use of soft replica for the ISI cancellation can reduce the influence of error propagation and therefore, improve the BER performance compared to the use of hard replica. We also showed that the frequency-domain block signal detection combined with soft decision iterative V-BLAST provides better BER performance than the conventional frequency-domain ISI cancellation combined with MMSE-FDE, but, at the cost of increased complexity. The complexity reduction of the proposed signal detection is left as an important future study item.

References

[1] J.G. Proakis and M. Salehi, Digital Communications, 5th ed., McGraw-Hill, 2008.

[2] D. Falconer, S.L. Ariyavisitakul, A. Benyamin-Seeyar, and B. Edison, "Frequency domain equalization for single-carrier broadband wireless systems," IEEE Commun. Mag., vol.40, no.4, pp.58–66, April 2002.

[3] K. Takeda, T. Itagaki, and F. Adachi, "Joint use of frequency-domain equalization and transmit/receive antenna diversity for single-carrier transmissions," IEICE Trans. Commun., vol.E87-B, no.7, pp.1946–1953, July 2004.

[4] K. Takeda, K. Ishihara, and F. Adachi, "Frequency-domain ICI cancellation with MMSE equalization for DS-CDMA downlink," IEICE Trans. Commun., vol.E89-B, no.12, pp.3335–3343, Dec. 2006.

[5] P.W. Wolniansky, G.J. Foschini, G.D. Golden, and R.A. Valenzuela, "V-BLAST: An architecture for realizing very high data rates over the rich-scattering wireless channel," Proc. 1998 URSI International Symposium on Singals, Systems, and Electronics (ISSSE'98), pp.295–300, Sept. 1998.

[6] C. Shen, H. Zhuang, L. Dai, S. Zhou, and Y. Yao, "Performance improvement of V-BLAST through an iterative approach," Proc. IEEE Giga bit Wireless for Real Personal, Indoor and Mobile Radio Communications (PIMRC) 2003, vol.3, pp.2553–2557, Sept. 2003.

[7] A. Nakajima and F. Adachi, "Throughput performance of iterative frequency-domain SIC with 2D MMSE-FDE for SC-MIMO multiplexing," Proc. IEEE 64th Vehicular Technology Conference (VTC-Fall), pp.25–28, Canada, Sept. 2006.

[8] G.H. Golub and C.F. van Loan, Matrix Computations, 3rd ed., Johns Hopkins Univ. Press, Baltimore, MD, 1996.

[9] R. Bohnke and K. Kammeyer, "SINR analysis for V-BLAST with ordered MMSE-SIC detection," Proc. International Wireless Communications and Mobile Computing Conference, pp.623–628, July 2006.

[10] K. Takeda and F. Adachi, "HARQ throughput performance of multicode DS-CDMA with MMSE turbo equalization," Proc. IEEE 65th Vehicular Technology Conference (VTC-Spring), pp.1603–1607, Ireland, April 2007.

[11] A. Stefanov and T. Duman, "Turbo coded modulation for wireless communications with antenna diversity," Proc. IEEE 49th Vehicular Technology Conference (VTC-Fall), pp.1565–1569, Netherlands, Sept. 1999.

[12] F. Adachi and K. Takeda, "Bit error rate analysis of DS-CDMA with joint frequency-domain equalization and antenna diversity combining," IEICE Trans. Commun., vol.E87-B, no.10, pp.2991–3002, Oct. 2004.

[13] C.R. Glassey, "An orthogonalization method of computing the generalized inverse of matrix," Operations Research Center, Univ. of California, Berkeley, 1966.

[14] T.N.E. Grevill, "Some applications of the pseudo inverse of a matrix," SIAM Review, vol.2, no.1, pp.15–22, Jan. 1960.

[15] A. Ben-Israel and S.J. Wersasn, "An elimination method for computing the generalized inverse of an arbitrary complex matrix," J. ACM, vol.10, pp.532–537, Oct. 1963.

Tetsuya Yamamoto received his B.S. degree in Electrical, Information and Physics Engineering from Tohoku University, Sendai, Japan, in 2008. Currently he is a graduate student at the Department of Electrical and Communications Engineering, Tohoku University. His research interests include frequency-domain equalization and signal detection techniques for mobile communication systems. He was a recipient of the 2008 IEICE RCS (Radio Communication Systems) Active Research Award.

Kazuki Takeda received his B.S. and M.S. degree in communications engineering from Tohoku University, Sendai, Japan, in 2006 and 2008. Currently he is a Japan Society for the Promotion of Science (JSPS) research fellow, studying toward his Ph.D. degree at the Department of Electrical and Communications Engineering, Graduate School of Engineering, Tohoku University. His research interests include precoding and channel equalization techniques for mobile communication systems. He was a recipient of the 2006 IEICE RCS (Radio Communication Systems) Active Research Award.

Fumiyuki Adachi received the B.S. and Dr.Eng. degrees in electrical engineering from Tohoku University, Sendai, Japan, in 1973 and 1984, respectively. In April 1973, he joined the Electrical Communications Laboratories of Nippon Telegraph & Telephone Corporation (now NTT) and conducted various types of research related to digital cellular mobile communications. From July 1992 to December 1999, he was with NTT Mobile Communications Network, Inc. (now NTT DoCoMo, Inc.), where he led a research group on wideband/broadband CDMA wireless access for IMT-2000 and beyond. Since January 2000, he has been with Tohoku University, Sendai, Japan, where he is a Professor of Electrical and Communication Engineering at the Graduate School of Engineering. His research interests are in CDMA wireless access techniques, equalization, transmit/receive antenna diversity, MIMO, adaptive transmission, and channel coding, with particular application to broadband wireless communications systems. From October 1984 to September 1985, he was a United Kingdom SERC Visiting Research Fellow in the Department of Electrical Engineering and Electronics at Liverpool University. He was a was a co-recipient of the IEICE Transactions best paper of the year award 1996 and again 1998 and, also a recipient of Achievement award 2003. He is an IEEE Fellow and was a co-recipient of the IEEE Vehicular Technology Transactions best paper of the year award 1980 and again 1990 and also a recipient of Avant Garde award 2000. He was a recipient of Thomson Scientific Research Front Award 2004 and the Ericsson Telecommunications Award 2008.