A Study of Frequency-Domain Signal Detection for Single-Carrier Transmission

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Abstract—**One-tap frequency-domain equalization (FDE) 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 FDE. In this paper, we point out that the frequency-domain received SC signal can be represented similar to the multiple-input multiple-output (MIMO) multiplexing case and then, we propose a new frequency-domain block signal detection, which combines FDE with MIMO signal detection. We apply well-known minimum mean square error (MMSE) detection and Vertical-Bell Laboratories layered space-time architecture (V-BLAST) detection and evaluate, by computer simulation, the achievable BER performance of SC transmission.**

Keywords-component; SC, MIMO, MMSE, V-BLAST

I. 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 as the transmission data rate increases. Simple one-tap frequency-domain equalization (FDE) based on the minimum mean square error criterion (MMSE) can significantly improve the bit error rate (BER) performance of single-carrier (SC) transmission in a frequency-selective fading channel [1, 2]. However, a big performance gap from the theoretical lower bound still exists due to the presence of residual inter-symbol interference (ISI) after FDE [3].

In this paper, we first 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. Then, we propose a new signal detection, called frequency-domain block signal detection, for the SC signal transmissions. There are various MIMO signal detection schemes; MMSE detection [4], Vertical-Bell Laboratories layered space-time architecture (V-BLAST) detection [5], etc. We apply these MIMO signal detections and evaluate, by computer simulation, the achievable BER performance of SC transmissions.

The remainder of this paper is organized as follows. Sect. II presents transmission system model. In Sect. III, frequency-domain block signal detection with the MMSE detection and V-BLAST detection is presented. In Sect. IV, the achieved BER performance in a frequency-selective fading channel is evaluated by computer simulation. Sect. V offers the conclusion.

II. SC TRASNMISSION MODEL

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 is expressed using the vector form as $\mathbf{d}=[d(0),\ldots,d(N_c-1)]^T$. 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. A sequence of GI-inserted signal blocks is transmitted.

Each transmitted signal block goes through frequency-selective fading channels which is 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 $E[\sum_{l=0}^{L-1}|h_l|^2]=1$ and the time delay of the lth path. The GI-removed received signal block can be expressed using the vector form as

$$\mathbf{r}=[r(0),\ldots r(N_c-1)]^T=\sqrt{\frac{2E_s}{T_s}}\mathbf{hd}+\mathbf{n}\,, \tag{2}$$

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

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

and $\mathbf{n}=[n(0),\ldots,n(N_c-1)]^T$ is the noise vector. The t-th element $n(t)$ of $\mathbf{n}$ is the zero-mean additive white Gaussian noise

(AWGN) having the 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 $\mathbf{R}=[R(0),\ldots,R(N_c-1)]^T$. $\mathbf{R}$ is expressed as

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

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 [6]. 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~N_c–1, and $(.)^H$ is the Hermitian transpose operation. Using Eq. (6), Eq. (4) can be rewritten as

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

where $\overline{\mathbf{H}}=\mathbf{HF}$ and $\mathbf{N}=[N(0),..,N(N_c-1)]$ 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 $\overline{\mathbf{H}}$ (see Fig.2). According to this understanding, a new frequency-domain block signal detection scheme, which combines FDE with MIMO signal detection, can be developed for the reception of the SC signal.

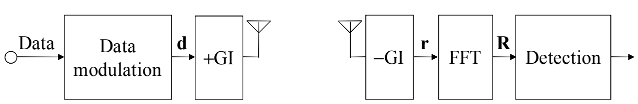

Figure 1. System model of SC transmission.

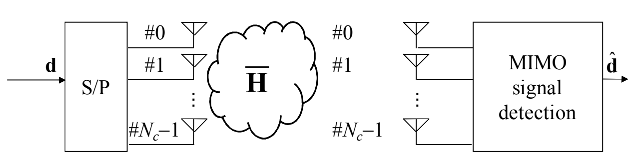

Figure 2. $N_c\times N_c$ MIMO system equivalent of SC trasnmission (Fig.1).

III. Frequency-domain Block Signla Detection

A. One-tap MMSE-FDE

First, we review the conventional one-tap MMSE-FDE [2]. One-tap MMSE-FDE is carried out using the MMSE weight matrix $\mathbf{W}_{MMSE\text{-}FDE}=diag[W(0),...,W(N_c-1)]$, where $W(k)=H^*(k)/(|H(k)|^2+(E_s/N_0)^{-1})$ for k=0~N_c–1. The MMSE weight minimizes the mean square of equalization error $e(k)$ defined as $e(k)=W(k)R(k)–D(k)$, where $D(k)$ is the kth element of the frequency-domain signal vector $\mathbf{D}=[D(0),\ ...,\ D(N_c-1)]^T$. After FDE, N_c-point inverse FFT (IFFT) is applied to obtain the decision variable as

$$\tilde{\mathbf{d}}=\mathbf{F}^H diag[W(0),\ldots,W(N_c-1)]\mathbf{R}\,. \tag{8}$$

B. MMSE Detection

The MMSE detection [4] is carried out using the weight matrix $\mathbf{W}_{MMSED}$ which minimizes the trace $tr[\mathrm{E}(\mathbf{ee}^H)]$ of the covariance matrix of the error vector $\mathbf{e}=\mathbf{W}_{MMSED}\mathbf{R}-\sqrt{2E_s/T_s}\mathbf{d}$. The MMSE weight matrix $\mathbf{W}_{MMSED}$ is given by

$$\mathbf{W}_{MMSED}=\overline{\mathbf{H}}^H\left[\overline{\mathbf{H}\mathbf{H}^H}+\left(\frac{E_s}{N_0}\right)^{-1}\mathbf{I}_{N_c}\right]^{-1}, \tag{9}$$

where $\mathbf{I}_{N_c}$ is the $N_c\times N_c$ identity matrix. The decision valuable vector $\tilde{\mathbf{d}}=[\tilde{d}(0),\cdots,\tilde{d}(N_c-1)]^T$ is given by $\tilde{\mathbf{d}}=\mathbf{W}_{MMSED}\mathbf{R}$. Since $\overline{\mathbf{H}\mathbf{H}^H}=\mathbf{HH}^H$ and $\overline{\mathbf{H}}^H=\mathbf{F}^H\mathbf{H}^H$, $\tilde{\mathbf{d}}$ can be expressed as

$$\tilde{\mathbf{d}}=\mathbf{F}^H diag[W(0),W(1),\ldots,W(N_c-1)]\mathbf{R}\ , \tag{10}$$

where $W(k)$ is the weight for the kth frequency. Eq. (10) shows that the MMSE detection is equivalent to one-tap MMSE-FDE.

C. Hard decision iterative V-BLAST Detection

Frequency-domain block signal detection using iterative V-BLAST detection is illustrated in Fig.3. The V-BLAST detection is composed of i) ordering, ii) ISI cancellation, and iii) signal detection. The transmitted symbol which has the highest signal-to-interference plus noise power ratio (SINR) among undetected symbols is detected by performing MMSE detection [7]. However, in the case of SC transmission, since the SINR is the same for all symbols, no ordering is necessary; the detection can be carried out simply from the first symbol (i.e., $d(0)$) in a block. The replica of the symbol, which has been just detected, is generated and is subtracted from the received signal (i.e., ISI cancellation). A new MMSE weight matrix for the next symbol to be detected is re-computed for detection. Until all of the transmitted symbols are detected, the above symbol detection & ISI cancellation is repeated.

V-BLAST detection cannot suppress the ISI sufficiently, in particular for the symbols with small indices. To further improve the performance, the use of iterative detection is effective [8]. After all of transmitted symbols are detected, V-BLAST detection is carried out again. This is repeated a sufficient number of times. In what follows, the signal detection of the nth layer (n=0~N_c–1) is presented (assuming that the symbols with n'=0~n–1 have been detected).

1) ISI cancellation

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 $\tilde{\mathbf{R}}^{(i,n)}=[\tilde{R}^{(i,n)}(0),\ldots,\tilde{R}^{(i,n)}(N_c-1)]^T$ in the nth layer of the ith iteration stage is given by

$$\tilde{\mathbf{R}}^{(i,n)}=\mathbf{R}-\sqrt{\frac{2E_s}{T_s}}\overline{\mathbf{H}}\hat{\mathbf{s}}^{(i,n)}, \tag{11}$$

where $\hat{\mathbf{s}}^{(i,n)}$ is the hard replica vector. $\hat{\mathbf{s}}^{(i,n)}$ is given by $\hat{\mathbf{s}}^{(i,n)}=[\hat{s}^{(i)}(0),\ldots,\hat{s}^{(i)}(n-1),0,\hat{s}^{(i-1)}(n+1),\ldots,\hat{s}^{(i-1)}(N_c-1)]^T$, where $\{\hat{s}^{(i)}(n');\ n'=0\sim n-1\}$ are generated using the decision in the present iteration stage while $\{\hat{s}^{(i)}(n');\ n'=n+1\sim N_c-1\}$ are generated using the decision results in the $(i-1)$th iteration stage.

2) MMSE detection

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

$$\tilde{d}^{(i)}(n)=\mathbf{W}^{(i,n)}\tilde{\mathbf{R}}^{(i,n)}. \tag{12}$$

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

$$\mathbf{W}^{(n)}=\overline{\mathbf{H}}_n^H\left[\tilde{\mathbf{H}}^{(i,n)}\tilde{\mathbf{H}}^{(i,n)H}+\left(\frac{E_s}{N_0}\right)^{-1}\mathbf{I}_{N_c}\right]^{-1}, \tag{13}$$

where

$$\tilde{\mathbf{H}}^{(i,n)}=\begin{cases}[\overline{\mathbf{H}}_n,\overline{\mathbf{H}}_{n+1},\ldots,\overline{\mathbf{H}}_{N_c-1}] & \text{for } i=0\\ \overline{\mathbf{H}}_n & \text{for } i>0\end{cases} \tag{14}$$

and $\overline{\mathbf{H}}_n$ is the nth column vector of $\overline{\mathbf{H}}$. In the i=0th iteration stage, we remove the column vectors, $\overline{\mathbf{H}}_0$, $\overline{\mathbf{H}}_1$, ..., $\overline{\mathbf{H}}_{n-1}$, associated with already detected symbols from $\overline{\mathbf{H}}$ assuming that these symbols have been cancelled perfectly. On the other hand, in the i(i>0)th iteration stage, we remove all column vectors $\overline{\mathbf{H}}_0$, ..., $\overline{\mathbf{H}}_{n-1}$, $\overline{\mathbf{H}}_{n+1}$, ..., $\overline{\mathbf{H}}_{N_c-1}$ from $\overline{\mathbf{H}}$; therefore, the weight vector given by Eq. (13) becomes the maximal ratio combining FDE (MRC-FDE) weight.

D. Soft Decision Iterative V-BLAST Detection

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 log-likelihood ratio (LLR) [3].

1) Generation of soft replica

Using the soft decision variable $\tilde{d}^{(i)}(n)$ associated with $d(n)$, the LLR for the xth bit, $x=0\sim\log_2M-1$, 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]

$$\tilde{s}^{(i)}(n)=\begin{cases}\frac{1}{\sqrt{2}}\tanh\left(\frac{\lambda_0^{(i)}(n)}{2}\right)+j\frac{1}{\sqrt{2}}\tanh\left(\frac{\lambda_1^{(i)}(n)}{2}\right) & \text{for QPSK}\\ \frac{1}{\sqrt{10}}\tanh\left(\frac{\lambda_0^{(i)}(n)}{2}\right)\left\{2+\tanh\left(\frac{\lambda_1^{(i)}(n)}{2}\right)\right\} & \text{for 16QAM}\\ +j\frac{1}{\sqrt{10}}\tanh\left(\frac{\lambda_2^{(i)}(n)}{2}\right)\left\{2+\tanh\left(\frac{\lambda_3^{(i)}(n)}{2}\right)\right\} & \end{cases}. \tag{15}$$

The LLR can be computed as [9]

$$\lambda_x^{(i)}(n)=\ln\left(\frac{p(b_{n,x}=1)}{p(b_{n,x}=0)}\right)\approx\frac{1}{2\hat{\sigma}_n^{(i)2}}\left\{\left|\tilde{d}^{(i)}(n)-\sqrt{\frac{2E_s}{T_s}}\hat{H}_n d^{\min}_{b_{x,n}=0}\right|^2-\left|\tilde{d}^{(i)}(n)-\sqrt{\frac{2E_s}{T_s}}\hat{H}_n d^{\min}_{b_{x,n}=1}\right|^2\right\}, \tag{16}$$

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. $d^{\min}_{b_{x,n}=0}$ (or $d^{\min}_{b_{x,n}=1}$) 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 $\hat{\mathbf{H}}=\mathbf{W}^{(i,n)}\overline{\mathbf{H}}$. $2\hat{\sigma}_n^{(i)2}$ is the variance of the noise plus residual ISI and is given by [11]

$$2\hat{\sigma}_n^{(i)2}=\frac{2N_0}{T_s}\left[\left\|\mathbf{W}^{(i,n)}\right\|^2+\frac{E_s}{N_0}\left\{\sum_{n'=0}^{n-1}\rho_{n'}^{(i)}\left|\hat{H}_{n'}\right|^2+\rho_{n'}^{(i-1)}\sum_{n'=n+1}^{N_c-1}\left|\hat{H}_{n'}\right|^2\right\}\right], \tag{17}$$

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

$$\rho_n^{(i)}=E\left[\left|d(n)-\tilde{s}^{(i)}(n)\right|^2\right]=\begin{cases}1-\left|\tilde{s}^{(i)}(n)\right|^2 & \text{for QPSK}\\ \frac{4}{10}\tanh\left(\frac{\lambda_1^{(i)}(n)}{2}\right)+\frac{4}{10}\tanh\left(\frac{\lambda_3^{(i)}(n)}{2}\right)+1-\left|\tilde{s}^{(i)}(n)\right|^2 & \text{for 16QAM}\end{cases}. \tag{18}$$

2) MMSE detection

In hard decision iterative V-BLAST detection, 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. However, the residual ISI is still present after ISI cancellation and therefore, the MMSE weight vector needs to be updated taking into account the residual ISI in each layer of each iteration stage. In [11], the MMSE weight vector which takes into account the residual inter-antenna interference is proposed for MIMO multiplexing. In this paper, we apply this MMSE weight vector to SC detection.

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

$$\mathbf{W}^{(i,n)} = \overline{\mathbf{H}}_n^H \left[\overline{\mathbf{H}} \boldsymbol{\rho}^{(i,n)} \overline{\mathbf{H}}^H + \left(\frac{E_s}{N_0} \right)^{-1} \mathbf{I}_{N_c} \right]^{-1}, \quad (19)$$

where $\boldsymbol{\rho}^{(i,n)} = diag[\rho_0^{(i)}, \cdots, \rho_{N_c-1}^{(i)}]$. 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. After ISI cancellation using soft replica as Eq. (11), MMSE detection on the nth symbol is performed by multiplying the frequency-domain received signal vector by the $1\times N_c$ MMSE weight vector $\mathbf{W}^{(i,n)}$ as Eq. (12).

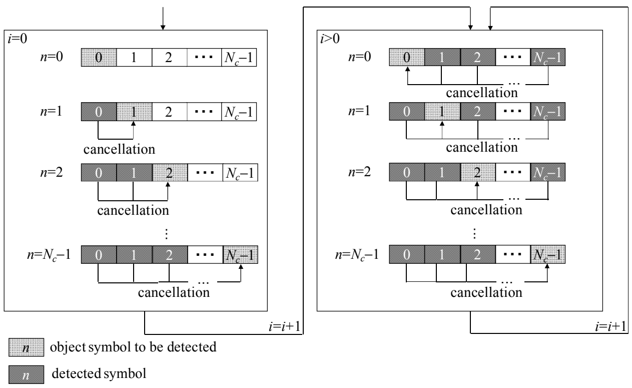

Figure 3. Frequency-domain block signal detection combined with iterative V-BLAST detection.

IV. Computer Simulation

We assume the 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 uniform power delay profile. Ideal channel estimation is assumed.

A. MMSE Detection and Hard Decision Iterative V-BLAST Detection

The average BER performances of SC transmission using frequency-domain block signal detection combined with MMSE detection or hard decision iterative V-BLAST detection are plotted in Fig. 4 as a function of average received bit energy-to-noise power spectrum density ratio $E_b/N_0(=(E_s/N_0)(N_c+N_g)/\log_2 M)$. Also plotted are the BER performances achievable by the one-tap MMSE-FDE and the theoretical lower bound [12].

As discussed in Sect. III, the achievable BER performances with the MMSE detection and the one-tap MMSE-FDE are identical. A big BER performance gap still exists from the lower bound for both detection schemes due to the residual ISI. For QPSK (16QAM) data modulation, the performance gap in E_b/N_0 for the average BER=10^{-4} is 7.2 (10.5) dB (including the GI insertion loss of 0.97dB).

On the other hand, the frequency-domain block signal detection using V-BLAST detection can suppress more the residual ISI and hence, can achieve better BER performance than the one-tap MMSE-FDE. It can be seen from Fig. 4 that hard decision iterative V-BLAST detection can improve the BER performance compared to non iterative V-BLAST detection (i=0). For QPSK (16QAM), the E_b/N_0 gap from the lower bound for the average BER=10^{-4} is reduced 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) does not further improve the BER performance due to the error propagation.

B. Soft Decision Iterative V-BLAST Detection

The average BER performance of SC transmission using frequency-domain block signal detection using 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 residual ISI is sufficiently suppressed by using the MMSE weight which takes into account the residual ISI and because the influence of error propagation can be reduced by using soft replica. For QPSK (16QAM), the E_b/N_0 gap from the lower bound for the average BER=10^{-4} is reduced by 1.5(3.2) dB when using two iterations (i=2).

Frequency-domain block signal detection using soft decision iterative V-BLAST detection always requires $N_c\times N_c$ matrix inversion at all layers of each iteration stage. Therefore, it can be said that significant performance improvement is obtained at the cost of increased complexity.

V. Conclusions

In this paper, we pointed out that the frequency-domain received SC signal can be represented similarly to the MIMO multiplexing case. Then, we proposed frequency-domain block signal detection, which combines FDE and MIMO signal detection such as MMSE detection and V-BLAST detection. We evaluated, by computer simulation, the BER performance of SC transmissions using frequency-domain block signal detection in a frequency-selective block Rayleigh fading channel. We showed that the frequency-domain block signal detection using the MMSE detection is identical to the one-tap MMSE-FDE. V-BLAST detection can achieve better BER performance than the one-tap MMSE-FDE. Iterative V-BLAST detection can further improve the BER performance compared to non-iterative V-BLAST detection, but when hard replica is used, the use of more than two iterations gives almost the same BER performance due to the error propagation. On the other hand, soft decision iterative V-BLAST detection can further improve the BER performance as the number of iterations increases. For QPSK (16QAM), the E_b/N_0 gap from the lower bound for the average BER=10^{-4} can be reduced to 1.5(3.2) dB when using two iterations.

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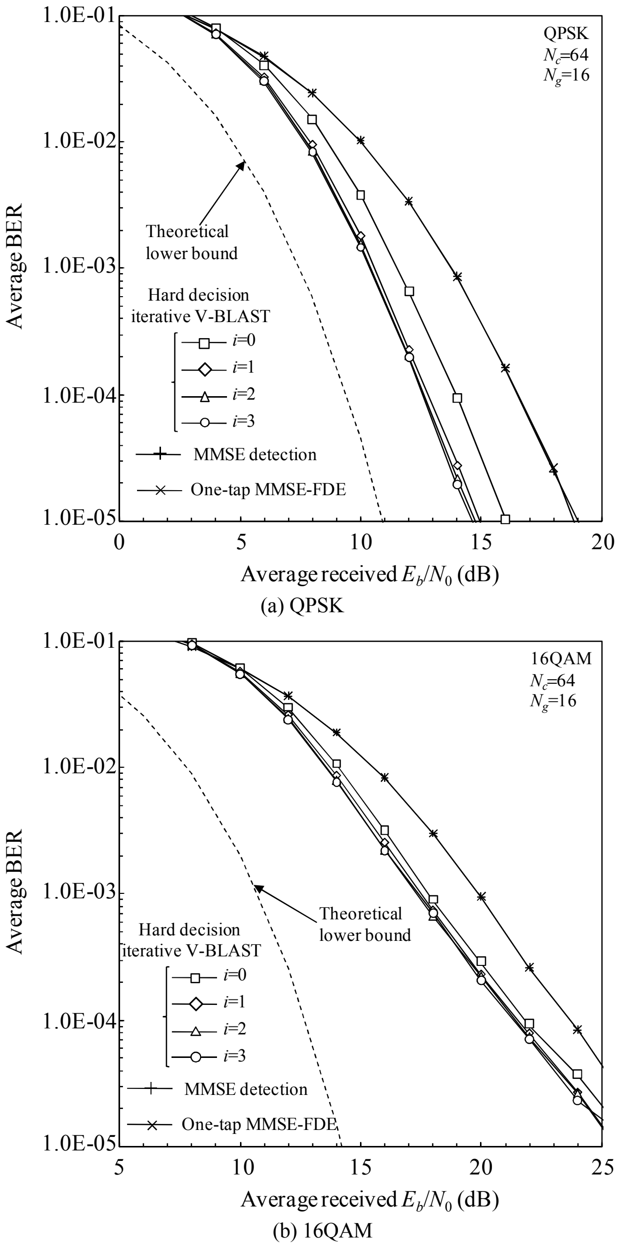

Figure 4. Hard decision iterative V-BLAST detection.

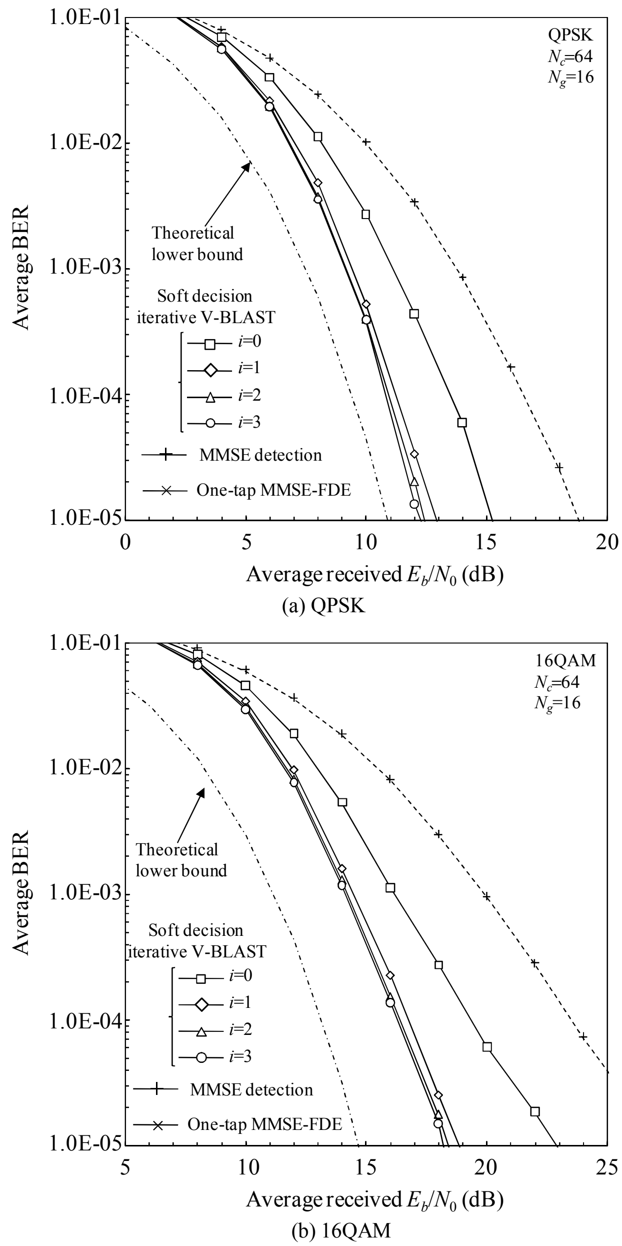

Figure 5. Soft decision iterative V-BLAST detection.