Noncoherent Decorrelating Multiuser Detector over a Frequency-Selective Rayleigh Fading Channel

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Abstract: In this paper, a novel noncoherent multiuser detector with diversity reception for a differentially encoded DS/CDMA signal in a frequency-selective Rayleigh fading channel is proposed. The proposed receiver employs multipath decorrelator and decision feedback differential detector (DFDD) with diversity reception. DFDD performs differntial encoding and diversity combining. The performance is obtained by analysis and simulation and it is sown that the proposed receiver outperforms conventioanl differential receiver with slight increas of complexity.

1. Introduction

When the accurate channel phase information is not available at the receiver, the use of noncoherent system is considered. In the context of CDMA systems, there have been considerable researches on noncohenrent multiuser detection schemes. For DPSK, it was shown that the decorrelator followed by differential detector is optimally near-far resistance [1]. Decorrelating differntial detector with diversity combiner is also studied in [2]. Adaptive MMSE multiuser detector for DPSK modulation is studied in [3]. For nonlinear modulations (e.g. FSK), noncoherent decorrelator [4] and noncoherent adaptive MMSE multiuser detector is investigated [5].

In this paper, a decorrelating multiuser detector for DPSK over a frequency-selective Rayleigh fading channel is considered. Instead of using equal gain combiner and conventional differential detector, decision feedback differential detection (DFDD) with diversity reception is employed. DFDD was proposed to improve conventional differential detector and modified for diversity channel. DFDD with diversity reception consists of MMSE channel estimator and optimum diversity combiner. By analysis and computer simulation it is shown that DF-DD shows better performance than that of conventional differential detector with slight increase of computational complexity. The rest of this paper is organized as follows. In section 2, signal and channel model is presented. The proposed receiver is introduced in section 3 and its performance is analyzed in section 4. Numerical results are shown in section 5 and conclusions are drawn in section 6.

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2. Signal and System Model

Consider a synchronous DS/CDMA system with user K in a frequency-selective fading channel having L resolvable multipaths. Assum that the BPSK symbol of user k transmitted at the n th symbol time, $b^{(k)}(n) \in \{+1,-1\}$, n=1,2,..., is independent, identically distributed (i.i.d). The differentially encoded BPSK symbol of user k at the n th symbol time is given by $d^{(k)}(n) = b^{(k)}(n)b^{(k)}(n-1)$. The signature waveform of user k is given by $u^{(k)}(t) = \sum_{p=0}^{P-1} u_p^{(k)} \psi(t - T_c p)$ where P is the spreading factor, $\{u_p^{(k)}\}_{p=0}^{P-1} \in \{+1,-1\}$, are the element of the signature sequence of use k and $\psi(t)$ stnads for a chip waveform with duration T_c. It is assumed that the energy of the signature waveform is normalized to 1. The channel impuse response of user k is given by

$$h^{(k)}(t) = \sum_{l=1}^{L} \sqrt{w_l^{(k)}} c_l^{(k)}(n) \delta(t-(l-1)T_c), \quad nT_s < t < (n+1)T_s,$$

where $w_l^{(k)}$ is the average power of path gain, $c_l(n)$ is the normalized complex Gaussian channel gain of the l th path and T_s stands for symbol duration, respectively. Assume that each path has same average power w. The baseband representation of the received signal in the n th symbol time is given by

$$r(t) = \sum_{k=1}^{K} \sum_{l=1}^{L} \sqrt{w} d^{(k)}(n) c_l^{(k)}(n) u^{(k)}(t - nT_s - (l-1)T_c) + n(t),$$

where $n(t)$ is the complex Gaussian random process with zero mean and two sided power spectral density with N_0. Letting $\mathbf{c}_k(n) \triangleq \left[c_1^{(k)}(n) \cdots c_L^{(k)}(n)\right]^T$, the correlation matrix between channel gain vector at the i th and j th symbol time is given by

$$\begin{aligned}\Sigma_k(i,j) &\triangleq E[\mathbf{c}_k(i)\mathbf{c}_k^H(j)] \\ &= J_0(2\pi f_d |i-j| T_s)\Sigma_k(0)\end{aligned}$$

where H stands for Hermittian transposition, $J_0(\cdot)$ is the zero-th order Bessel function of the first kind, f_d is maximum Doppler frequency, and $\Sigma_k(0) \triangleq E[\mathbf{c}_k(n)\mathbf{c}_k^H(n)]$.

Since $\Sigma_k(i,j)$ depends on time difference, $\Sigma_k(i,j)$ is represented by $\Sigma_k(|i-j|)$. Let $\mathbf{d}_k(n) \triangleq d^{(k)}(n)\mathbf{I}_L$ and $\mathbf{u}_k(t) \triangleq \left[u^{(k)}(t),\cdots,u^{(k)}(t-(L-1)T_c)\right]^T$ where $\mathbf{I}_L$ stands for the $L\times L$ identity matrix. The received signal in the n th symbol time is given by

$$r(t) = \sqrt{w}\mathbf{U}^T(t)\mathbf{D}(n)\mathbf{C}(n) + n(t), \tag{1}$$

where

$\mathbf{U}^T(t) \triangleq \left[\mathbf{u}_1^T(t)\cdots\mathbf{u}_K^T(t)\right]$, $\mathbf{D}(n) \triangleq \mathrm{diag}\left\{\mathbf{d}_1^T(n),\cdots\mathbf{d}_K^T(n)\right\}$

and $\mathbf{C}(n) \triangleq \left[\mathbf{c}_1^T(n)\cdots\mathbf{c}_K^T(n)\right]^T$.

3. Noncoherent Decorrelating Detector

The proposed noncoherent decorrelating multiuser detector consists of multipath decorrlator and decision feedback differential detector (DFDD) with diversity reception. The received signal is passed through the bank of KL filters matched to the delayed signature waveforms of each user. The multipath decorrelator eliminates multiple access interference (MAI) of matched filter output. Since L signals are avilable for each user after decorrleating, $b_k(n)$ is detected by employing DFDD with diversity recption.

The matched filter output is given by

$$\begin{aligned}\mathbf{y}(n) &\triangleq \int_{nT_s}^{(n+1)T_s}\mathbf{U}(t)r(t)dt \\ &= w\mathbf{R}\mathbf{D}(n)\mathbf{C}(n) + \mathbf{n}(n),\end{aligned} \tag{2}$$

where $\mathbf{R} = \int_{nT_s}^{(n+1)T_s}\mathbf{U}(t)\mathbf{U}^T(t)dt$ and $\mathbf{n}(n) = \int_{nT_s}^{(n+1)T_s}\mathbf{U}(t)n(t)dt$.

The noise correlation matrix is given by $N_0\mathbf{R}$. The multipath decorrelator output $\mathbf{q}(n) \triangleq \left[\mathbf{q}_1^T(n),\cdots,\mathbf{q}_K^T(n)\right]^T$ is given by

$$\begin{aligned}\mathbf{q}(n) &= \mathbf{R}^{-1}\mathbf{y}(n) \\ &= \sqrt{w}\mathbf{D}(n)\mathbf{C}(n) + \mathbf{v}(n),\end{aligned} \tag{3}$$

where $\mathbf{v}(n) \triangleq \mathbf{R}^{-1}\mathbf{n}(n) = \left[\mathbf{v}_1^T(n) \;\cdots\; \mathbf{v}_K^T(n)\right]^T$.

Decorrelator output of the k th user is given by

$$\mathbf{q}_k(n) = \sqrt{w}d^{(k)}(n)\mathbf{c}_k(n) + \mathbf{v}_k(n). \tag{4}$$

Though decorrelator perfectly eliminated MAI, it enhences output noise power. The correlation matrix of decorrelator output nois is given by $N_0\bar{\mathbf{R}}_k$ where $\bar{\mathbf{R}}_k$ is the k , k th $L\times L$ block matrix of $\mathbf{R}^{-1}$. To whitening noise, consider the Cholesky decomposition of $\bar{\mathbf{R}}_k$, i.e., $\bar{\mathbf{R}}_k = \bar{\mathbf{W}}_k\bar{\mathbf{W}}_k^H$. If the matched filter output $\mathbf{q}_k(n)$ is multipied by the matrix $\bar{\mathbf{W}}_k^{-1}$, an equivalent matched filter output vector is obtained as

$$\mathbf{q}_k'(n) = \sqrt{w}d^{(k)}(n)\bar{\mathbf{W}}_k^{-1}\mathbf{c}_k(n) + \mathbf{v}_k'(n), \tag{5}$$

where $\mathbf{v}_k'(n) \triangleq \bar{\mathbf{W}}_k^H\mathbf{v}_k(n)$ and its covariance matrix is $N_0\mathbf{I}_L$. Due to the independence of $\mathbf{v}_k'(n)$, $\mathbf{q}_k'(n)$ is called the whitened matched filter output vector. In what follows, $\mathbf{q}_k(n)$ and $\mathbf{v}_k(n)$ are used instead of $\mathbf{q}_k'(n)$ and $\mathbf{v}_k'(n)$ for the simplicity of notation.

In [6], DFDD based on MMSE channel estimator was proposed for a frequency-flat Rayleigh fading channel. In this paper, DFDD with diversity reception employing optimum diversity combiner is considered. Differentially coherent diversity combiner for DPSK [7] is the special case of DFDD with diversity reception whose channel estimator order is 1.

Letting the estimated channel gain vector be denoted by $\hat{\mathbf{c}}_k(n)$, the whitened matched filter output vector can be rewritten as

$$\begin{aligned}\mathbf{q}_k(n) &= b^{(k)}(n)d^{(k)}(n-1)\sqrt{w}\bar{\mathbf{W}}_k^{-1}\hat{\mathbf{c}}_k(n) + \sqrt{w}\bar{\mathbf{W}}_k^{-1}d^{(k)}(n-1) \\ &\quad\times\left(\mathbf{c}_k(n)-\hat{\mathbf{c}}_k(n)\right)b^{(k)}(n) + \mathbf{v}_k(n) \\ &= b^{(k)}(n)\mathbf{q}_{k,\mathrm{ref}}(n) + \sqrt{w}\bar{\mathbf{W}}_k^{-1}d^{(k)}(n-1)b^{(k)}(n) \\ &\quad\times\left(\mathbf{c}_k(n)-\hat{\mathbf{c}}_k(n)\right) + \mathbf{v}_k(n),\end{aligned} \tag{6}$$

where the reference signal vector

$$\mathbf{q}_{k,\mathrm{ref}}(n) \triangleq d^{(k)}(n-1)\sqrt{w}\bar{\mathbf{W}}_k^{-1}\hat{\mathbf{c}}_k(n) \tag{7}$$

The 2nd and 3rd terms of (6) corresponds to additive noise and its covariance matrix is given by $\mathbf{N}_k$. For differantial decoding and optimum diversity combining, $\mathbf{q}_{k,\mathrm{ref}}^H(n)\mathbf{N}_k^{-1}$ must be multiplied to (6).

First consider MMSE channel estimator of to produce $\mathbf{q}_{k,\mathrm{ref}}^H(n)$. From $N-1$ received signals, $\mathbf{q}_{k,\mathrm{ref}}^H(n)$ is obtained.

$$\begin{aligned}\mathbf{q}_{k,\mathrm{ref}}^H(n) &= d^{(k)}(n-1)\sqrt{w}\hat{\mathbf{c}}_k^H(n)\bar{\mathbf{W}}_k^{-H} \\ &= \sum_{v=1}^{N-1}\mathbf{q}_k^H(n-N+v)\mathbf{A}_v \\ &= \sum_{v=1}^{N-1}\prod_{p=v+1}^{N-1}\hat{b}^{(k)}(n-N+p)\mathbf{q}_k^H(n-N+v)\mathbf{A}_v',\end{aligned} \tag{8}$$

where $\mathbf{A}_v' \triangleq \prod_{p=v+1}^{N-1}\hat{b}^{(k)}(n-N+p)\mathbf{A}_v$ and $\hat{b}^{(k)}(n-N+p)$, $p = 2,3,\cdots,N-1$, are previously decided feedback symbols.

To determine filter coefficients $\mathbf{A}_v'$, multiply both sides of (8) by $d^{(k)}(n-1)$ and plug (5) into (8), it becomes

$$\begin{aligned}&\mathbf{q}_{k,\mathrm{ref}}^H(n)d^{(k)}(n-1) \\ &= \sqrt{w}\hat{\mathbf{c}}_k^H(n)\bar{\mathbf{W}}_k^{-H} \\ &= \sum_{v=1}^{N-1}\left(\sqrt{w}\mathbf{c}_k^H(n-N+v)\bar{\mathbf{W}}_k^{-H} + d^{(k)}(n-N+v)\mathbf{v}_k^H(n)\right)\mathbf{A}_v'.\end{aligned} \tag{9}$$

Filter coefficients that minimize the mean squared error between $\sqrt{w}\hat{\mathbf{c}}_k^H(n)\bar{\mathbf{W}}_k^{-H}$ and $\sqrt{w}\mathbf{c}_k^H(n)\bar{\mathbf{W}}_k^{-H}$ are obtained by Wiener equation [8]

$$(\mathbf{\Phi}_k + N_0\mathbf{I}_{(N-1)L})\mathbf{A}' = \mathbf{\Gamma}_k \tag{10}$$

where

$\mathbf{\Gamma}_k \triangleq$

$w\left[\bar{\mathbf{W}}_k^{-H}\Sigma_k(1)\bar{\mathbf{W}}_k^{-1} \quad \bar{\mathbf{W}}_k^{-H}\Sigma_k(2)\bar{\mathbf{W}}_k^{-1} \quad\cdots\quad \bar{\mathbf{W}}_k^{-H}\Sigma_k(N-1)\bar{\mathbf{W}}_k^{-1}\right]^H$

$\mathbf{\Phi}_k \triangleq$

$$w\begin{bmatrix} \bar{\mathbf{W}}_k^{-1}\Sigma_k(0)\bar{\mathbf{W}}_k^{-H} & \bar{\mathbf{W}}_k^{-1}\Sigma_k(1)\bar{\mathbf{W}}_k^{-H} & \cdots & \bar{\mathbf{W}}_k^{-1}\Sigma_k(N-2)\bar{\mathbf{W}}_k^{-H} \\ \bar{\mathbf{W}}_k^{-1}\Sigma_k(1)\bar{\mathbf{W}}_k^{-H} & \ddots & \ddots & \vdots \\ \vdots & \ddots & \ddots & \bar{\mathbf{W}}_k^{-1}\Sigma_k(1)\bar{\mathbf{W}}_k^{-H} \\ \bar{\mathbf{W}}_k^{-1}\Sigma_k(N-2)\bar{\mathbf{W}}_k^{-H} & \cdots & \bar{\mathbf{W}}_k^{-1}\Sigma_k(1)\bar{\mathbf{W}}_k^{-H} & \bar{\mathbf{W}}_k^{-1}\Sigma_k(0)\bar{\mathbf{W}}_k^{-H} \end{bmatrix}$$

and

$\mathbf{A}' \triangleq [\mathbf{A}_1'^H \cdots \mathbf{A}_{N-1}'^H]^H$.

Next, consider noise covariance matrix $\mathbf{N}_k$. In (6), noise consists of channel estimation error and AWGN and its correlation matrix is given by $\mathbf{N}_k = \mathbf{E}_k + N_0\mathbf{I}_L$ where MMSE channel estimation error correlation matrix,

$$\mathbf{E}_k \triangleq w\bar{\mathbf{W}}_k^{-1}\Sigma_k(0)\bar{\mathbf{W}}_k^{-H} - \Gamma_k^H(\Phi_k + N_0\mathbf{I}_{(N-1)L})^{-1}\Gamma_k \text{ [8]}. \quad (11)$$

Combining the results of (9) and (11), the decision variable becomes

$$\mathbf{q}_{k,\mathrm{ref}}^H(n)(\mathbf{E}_k + N_0\mathbf{I}_L)^{-1}\mathbf{q}_k(n). \quad (12)$$

If there is only AWGN, (12) becomes well-known maximal ration combiner (MRC). The decision rule for $b(n)$ is given by

$$\begin{aligned}\hat{b}(n) &= \operatorname{sgn}\operatorname{Re}\left\{\mathbf{q}_{k,\mathrm{ref}}^H(n)\left(\mathbf{E}_k + N_0\mathbf{I}_L\right)^{-1}\mathbf{q}_k(n)\right\} \\ &= \operatorname{sgn}\operatorname{Re}\left\{\left(\sum_{v=1}^{N-1}\prod_{p=v+1}^{N-1}\hat{b}^{(k)}(n-N+p)\mathbf{q}_k(n-N+j)\mathbf{P}_v\right)\mathbf{q}_k(n)\right\},\end{aligned} \quad (13)$$

where

$$\begin{aligned}&\begin{bmatrix}\mathbf{P}_1^H & \cdots & \mathbf{P}_{N-1}^H\end{bmatrix}^H \\ &\triangleq \mathbf{A}'\left(\mathbf{E}_k + N_0\mathbf{I}_L\right)^{-1} \\ &= (\Phi_k + N_0\mathbf{I}_{(N-1)L})^{-1}\Gamma_k \\ &\quad\times\left(N_0\mathbf{I}_L + \bar{\mathbf{W}}_k^{-1}\Sigma(0)\bar{\mathbf{W}}_k^{-H} - \Gamma_k^H(\Phi_k + N_0\mathbf{I}_{(N-1)L})^{-1}\Gamma_k\right)^{-1}\end{aligned}$$

4. Performance Analysis

4.1 Bit Error Rate

To simplify analysis, genie-aided DFDD with diversity reception is considered. In genie-aided DFDD, it is assumed that all previously decided feedback symbols are error free. Replacing $\hat{b}^{(k)}(n-N+p)$ with $b^{(k)}(n-N+p)$, the decision rule of genie-aided DFDD for binary DPSK is given by

$$\hat{b}(n) = \operatorname{sgn}\operatorname{Re}\left\{g^{(k)}(n)\right\},$$

where

$$\begin{aligned}g^{(k)}(n) &= d^{(k)}(n-1)\sqrt{w}\hat{\mathbf{c}}_k^H(n)\bar{\mathbf{W}}_k^{-H}\left(\mathbf{E}_k + N_0\mathbf{I}_L\right)^{-1}\mathbf{q}(n) \\ &= d^{(k)}(n-1)\sqrt{w}\hat{\mathbf{c}}_k^H(n)\bar{\mathbf{W}}_k^{-H}\left(\mathbf{E}_k + N_0\mathbf{I}_L\right)^{-1} \\ &\quad\times\{b^{(k)}(n)d^{(k)}(n-1)\sqrt{w}\bar{\mathbf{W}}_k^{-1}\hat{\mathbf{c}}_k(n) \\ &\quad+\sqrt{w}\bar{\mathbf{W}}_k^{-1}d^{(k)}(n-1)\left(\mathbf{c}_k(n)-\hat{\mathbf{c}}_k(n)\right)b^{(k)}(n)+\mathbf{v}_k(n)\}.\end{aligned}$$

Conditioned on the channel estimate $\hat{\mathbf{c}}_k(n)$, the mean and variance of $g^{(k)}(n)$ is given by

$$E[g^{(k)}(n)\,|\,\hat{\mathbf{c}}_k(n)] = w\mathbf{c}_k^H(n)\left(\bar{\mathbf{W}}_k\mathbf{E}_k\bar{\mathbf{W}}_k^H + N_0\bar{\mathbf{R}}_k\right)^{-1}\mathbf{c}_k(n)b^{(k)}(n)$$

and

$$Var[g^{(k)}(n)\,|\,\hat{\mathbf{c}}_k(n)] = w\mathbf{c}_k^H(n)\left(\bar{\mathbf{W}}_k\mathbf{E}_k\bar{\mathbf{W}}_k^H + N_0\bar{\mathbf{R}}_k\right)^{-1}\mathbf{c}_k(n)$$

The conditional bit error rate ig given by

$$P_{e|\hat{\mathbf{c}}(n)}^{(k)} = Q\left(\sqrt{\frac{2w\hat{\mathbf{c}}_k^H(n)\left(\bar{\mathbf{W}}_k\mathbf{E}_k\bar{\mathbf{W}}_k^H N_0^{-1} + \bar{\mathbf{R}}_k\right)^{-1}\mathbf{c}_k(n)}{N_0}}\right), \quad (14)$$

where $Q(x) \triangleq \frac{1}{\sqrt{2\pi}}\int_x^\infty \exp\left(-\frac{z^2}{2\pi}\right)dz$. Averaging (14) over a zero-mean complex Gaussian random vector $\hat{\mathbf{c}}_k(n)$, the avergage bit error rate is given by [7]

$$P_e^{(k)} = \sum_{l=1}^{L}\frac{\alpha_l^{(k)}}{2}\left[1-\sqrt{\frac{w\psi_l^{(k)}}{N_0 + w\psi_l^{(k)}}}\right] \quad (15)$$

where $\alpha_l^{(k)} = \prod_{i=1,i\neq l}^{L}\frac{\psi_i^{(k)}}{\psi_i^{(k)} - \psi_l^{(k)}}$ and ψ_i, $i = 1,2,\cdots,L$, is distinct eigenvalues of the product of $\left(\Sigma_k(0) - \bar{\mathbf{W}}_k\mathbf{E}_k\bar{\mathbf{W}}_k^H / w\right)$ and $\left(\bar{\mathbf{W}}_k\mathbf{E}_k\bar{\mathbf{W}}_k^H N_0^{-1} + \bar{\mathbf{R}}_k\right)^{-1}$.

4.2 Asymptotic Multiuser Efficiency

As another performance measure of multiuser detector, AME (asymptotic multiuser efficiency) is considered. AME is defined as the ratio of SNR with and without the presence of multiple access interference. AME characterizes performance degradation due to MAI in compared with performance of single user receiver. Letting the SNR of single user receiver γ_1 and the SNR of multiuser detector $\gamma_{k,mul}$, AME is defiened as $\eta_k \triangleq \lim_{N_0\to 0}\gamma_{k,mul}/\gamma_k$.

If the output of multiuser detector sufferes from MAI then bit error rate shows an error floor in high SNR region due to MAI and AME becomes to 0. If the AME of a multiuser detector are between 0 and 1, a multiuser detector is call near-far resistance. Decrasing AME means that the performance of a multiuser detector is more degraded in comparision with a single user receiver. The average SNR for sigle user receiver is given by [2]

$$\gamma_1 = \left(\prod_{l=1}^{L}\frac{w\psi_l^{(1)}}{N_0}\right)^{1/L} \quad (16)$$

As N_0 approach to 0, $\psi_l^{(1)}$ becomes the half of the l th eigen value of $\Sigma_1(0)\mathbf{R}_1$, where $\mathbf{R}_1 \triangleq \int_{nT_s}^{(n+1)T_s}\mathbf{u}_1(t)\mathbf{u}_1^T(t)dt$ [7]. The average SNR of the proposed multiuser detector is given by

$$\gamma_{k,mul} = \left(\prod_{l=1}^{L}\frac{w\psi_l^{(k)}}{N_0}\right)^{1/L} \quad (17)$$

As N_0 approach to 0, $\psi_l^{(k)}$ becomes the half of the l th eigen value of $\Sigma(0)\bar{\mathbf{R}}_k^{-1}$. Using the fact that the product of eigenvalues are determinant the matrix, the AME is given by

$$\eta_k = \left(\frac{\det\left(\Sigma(0)\bar{\mathbf{R}}_k^{-1}\right)}{\det\left(\Sigma(0)\mathbf{R}_1\right)}\right)^{1/L} = \left(\frac{1}{\det\left(\mathbf{R}_1\right)\det\left(\bar{\mathbf{R}}_k\right)}\right)^{1/L}. \quad (18)$$

5. Numerical Results

Suppose that BPSK is used as a modulation scheme and the Gold code of length 31 is used as a signature sequence for each user. Fig. 2 shows that the BER in a slow Rayleigh fading channel.

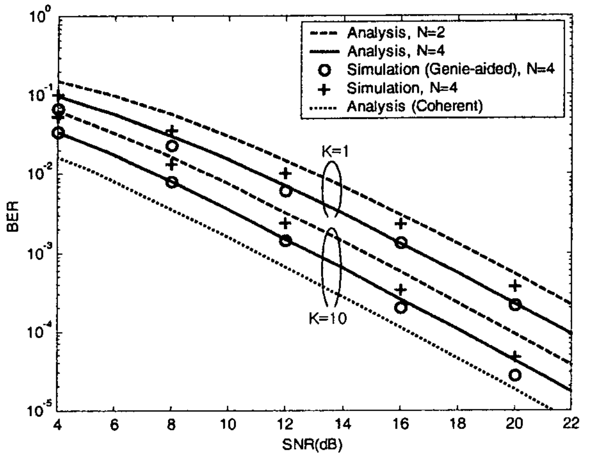

Fig. 1. BER in a slow Ralyeigh fading channel, $f_d T_s = 0.001$.

It is shown that for both $K = 1$ and 10, BER for $N = 4$ is smaller than that for $N = 2$. Proposed detector with genie-aided DFDD for $N = 4$ achieves smaller BER than DFDD for $N = 4$. This is because erroneously decided feedback symbols cause errors in DFDD.

Fig. 2 shows that the BER in a fast Rayleigh fading channel.

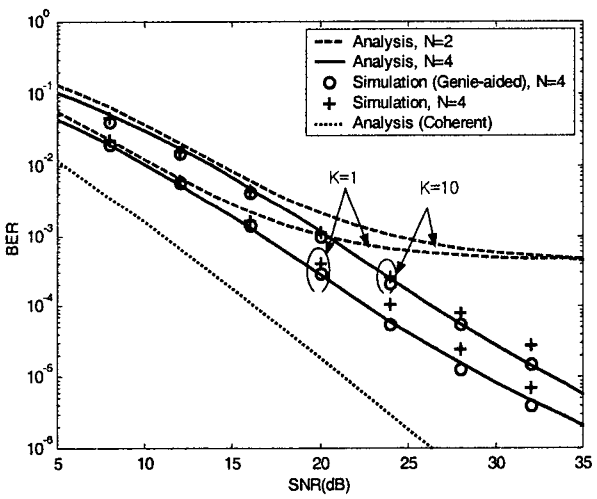

Fig. 2. BER in a fast Rayleigh fading channel, $f_d T_s = 0.05$.

It is shown that when $N = 2$, BER for $K = 1$ becomes the same as BER for $K = 10$ as SNR increases. Although the BER for the proposed detector with DFDD for $N = 4$ is larger than BER for genie-aided DFDD for $N = 4$, the performance of DFDD for $N = 4$ is improved significantly in high SNR region in compared with DFDD with $N = 2$.

Fig. 3 shows the AME for the proposed receiver. In Fig. 4, it is shown that AME decreases as the number of users and the number of multipaths increases. This is because the multipath decorrelator treats K users signal with L multipaths as KL equivalent users.

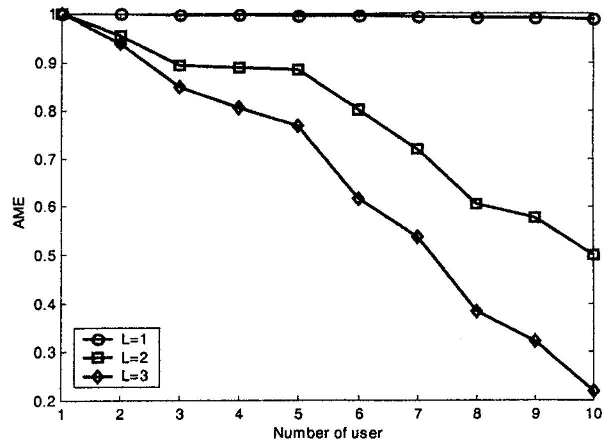

Fig. 3. Asymptotic multiuser efficiency

6. Conclusions

In this paper, the noncoherent decorrelating multiuser for a frequency-selective Rayleigh fading channel is proposed. The proposed receiver is a multipath decorrelator employing DF-DD with diversity reception. DFDD with diversity reception consists of an LMMSE channel estimator of order $N-1$ and a differentially coherent optimum diversity combiner. BER and AME of the proposed detector are analyzed. From numerical results, it is shown that the proposed receiver outperforms the conventional differential detector with the slight increase of computational complexity. As the number of observed symbol increases, the performance of proposed receiver approaches that of coherent receiver in a slow fading channel, and error floor is lowered in a fast fading channel.

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