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周波数領域遅延検波と等化を用いる差動符号化 DS-CDMA の BER の理論検討

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あらまし これまで筆者らは周波数領域遅延検波と等化(FDDDE)を用いる差動符号化 DS-CDMA を提案し，フェージングの時間変動の激しい環境下では，IIR フィルタを用いる FDDED はパイロットを用いてチャネル推定するコヒーレント周波数領域等化(FDE)よりも優れたビット誤り率(BER)特性が得られることを計算機シミュレーションにより明らかにしてきた．本報告では，FDDDE を用いる差動符号化 DS-CDMA の BER を理論検討し，遅延スプレッドおよびドップラー周波数の関数として導出している．導出した BER 表現を用いて最適な IIR フィルタ係数を求め，そのときの BER 特性を計算機シミュレーションにより確認している．

キーワード　DS-CDMA，差動符号等化，周波数領域等化．

A Closed-form BER Expression for Differentially Modulated DS-CDMA Signals with Frequency-domain Differential Detection and Equalization

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Abstract Recently, we have proposed frequency-domain differential detection and equalization (FDDDE) for the reception of differentially encoded DS-CDMA signals in a time- and frequency-selective fading channel. It was shown by computer simulation that FDDDE using an IIR filter with decision feedback can provide better BER performance than coherent frequency-domain equalization (FDE) with pilot-assisted channel estimation in a fast fading channel. In this paper, a closed-form BER expression is derived for the FDDDE reception of DQPSK-modulated DS-CDMA signals as a function of the normalized root mean square (rms) delay spread and the normalized Doppler frequency. Using the derived BER expression, an optimum IIR filter coefficient is found under different channel conditions and the theoretical BER performance is confirmed by the computer simulation.

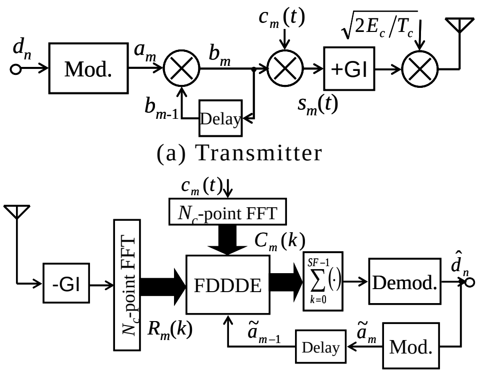
Keyword DS-CDMA, frequency-domain differential detection, frequency-domain equalization.

1. Introduction

Direct sequence-code division multiple access (DS-CDMA) has been adopted as one of multiple access schemes in the 3rd generation (3G) wireless communication systems [1]. However, future broadband wireless mobile systems are required to support a wide range of services and data rates. As the chip rate increases, the frequency-selectivity of the fading channel becomes severer due to the increasing number of resolvable propagation paths, which makes the well known coherent rake combining ineffective due to strong interpath interference (IPI) and too complex to implement [2]. Recently, frequency-domain equalization (FDE) has been proposed for DS-CDMA [3]. It is shown [4] that the FDE based on the minimum mean square error (MMSE) criterion can significantly improve the bit error rate (BER) performance of DS-CDMA downlink transmissions in a severe frequency-selective fading channel.

However, coherent FDE requires accurate channel estimation and the imperfect channel estimation results in the performance degradation. For pilot-assisted channel estimation, the known pilot block needs to be periodically transmitted. In order to track against fast fading, pilot block transmission rate must be increased. This reduces the transmission efficiency. Differential detection that requires no channel estimation is attractive owing to its simplicity and robustness against fast fading, but its BER performance is inferior to that of coherent detection since a delayed version of the received noisy signal is used as the phase reference. Recently, we have proposed frequency-domain differential detection and equalization (FDDDE) [5] for the reception of differentially encoded DS-CDMA signals in a time- and frequency-selective fading channel. In FDDDE, a simple infinite impulse response (IIR) filter with decision feedback is used to provide a noise-reduced reference signal and then narrow the performance gap between differential detection and coherent detection. It was shown by computer simulation [5] that FDDDE with IIR filter can provide better BER performance than coherent FDE with pilot-assisted channel estimation in a fast fading channel.

This paper is intended to provide a theoretical foundation to our proposed FDDDE. A closed-form BER expression is derived for differentially encoded DS-CDMA with FDDDE, taking into account error propagation due to the decision feedback. In [7], a general quadratic form in [6, Appendix B] is used to evaluate the BER performance of differentially modulated OFDM with antenna-diversity. Here, we use this general quadratic form to derive a closed-form BER expression of differentially encoded DS-CDMA with FDDDE in a time- and frequency-selective fading channel. An optimum IIR filter coefficient is obtained. The theoretical BER performance is confirmed by the computer simulation.



(a) Transmitter

(b) Receiver

Fig. 1. Transmitter/receiver structure.

2. Transmission System Model

The DS-CDMA transmitter/receiver structure is illustrated in Fig.1. Here, we assume the square-root Nyquist chip shaping filter at the transmitter and the same filter at the receiver as the chip-matched filter. Throughout the paper, T_c-spaced discrete time representation is used, where T_c represents the chip duration. In the following, $\lfloor x \rfloor$ is the largest integer smaller than or equal to x, $\lceil x \rceil$ is the smallest integer larger than or equal to x, and $E[\cdot]$ denotes the ensemble average.

2.1. Transmitted signal

At the transmitter, a binary bit sequence $\{d_n\}$ is modulated into a symbol sequence $\{a_m\}$. Then, $\{a_m\}$ is differentially encoded into $\{b_m\}$, where $b_m = a_m b_{m-1}$ with b_0=1 being the initial reference symbol and $|a_m|$=1. Next, b_m is spread by a spreading sequence $\{c_m(t), t=0\sim N_c-1\}$ with $|c_m(t)|=1$ to obtain the data chip sequence as

$$s_m(t) = b_m c_m(t). \quad (1)$$

Since each data symbol is transformed into an N_c-chip sequence, the spreading factor is N_c. Then, an N_g-chip GI is inserted every N_c chips to avoid inter-block interference (IBI). The sum of N_g-GI chip and N_c-data chip constitutes a chip block. The transmitted spread signal belonging to the mth differentially encoded data symbol b_m can be expressed using lowpass representation as

$$\hat{s}_m(t) = \sqrt{2E_c/T_c}\, s_m\left(t \bmod N_c\right) \quad (2)$$

for $t=-N_g\sim N_c-1$, where E_c is the average chip energy.

2.2. Channel

The GI-inserted signal is transmitted over a frequency- and time-selective fading channel. We assume a block fading, which means the path gain $h_l(t)$ remains constant over one block interval, but vary block-by-block. Assuming that the channel has L independent propagation paths, the discrete-time impulse response $h_m(\tau)$ during the reception of the mth chip block is expressed as

$$h_m(\tau) = \sum_{l=0}^{L-1} h_{m,l}\,\delta(\tau - \tau_l), \quad (3)$$

where $h_{m,l}$ and τ_l are respectively the complex-valued path gain with $\sum_{l=0}^{L-1} E[|h_{m,l}|^2] = 1$ and time delay of the lth path, and $\delta(x)$ is the delta function. τ_l is assumed to be a multiple integer of T_c and equal to $\tau_l=l$, $l=0\sim L-1$. The maximum time delay of $\{\tau_l\}$ is assumed to be shorter than the GI. The power delay profile of the channel is represented as

$$\Omega(\tau) = \sum_{l=0}^{L-1} \Omega_l\,\delta(\tau - \tau_l), \quad (4)$$

where $\Omega_l=E[|h_l(mT)|^2]$. Often used model is an exponentially decaying power delay profile. It was confirmed by recent propagation measurements [11], taken at a carrier frequency of 4.6GHz and a distance between transmitter and receiver of around 1km, that the power delay profile under a line-of-sight environment is well approximated by an exponentially decaying power delay profile. For the exponential power delay profile model, Ω_l is expressed as $\Omega_l=A\alpha^{-1}$, where α is the decay factor and $A=(1-\alpha^{-1})(1-\alpha^{-L})$. The frequency selectivity is represented by the normalized root mean square (rms) delay spread τ_{rms}/T_c, defined as [12]

$$\frac{\tau_{rms}}{T_c} = \sqrt{E[\tau^2] - [E(\tau)]^2} = \sqrt{\sum_{l=0}^{L-1} \tau_l^2 \Omega_l - \left[\sum_{l=0}^{L-1} \tau_l \Omega_l\right]^2}. \quad (5)$$

The movement of the mobile terminals causes time-selective fading. Assuming the Jakes' fading model [9], the autocorrelation function of $h_{m,l}$ is given as $E[h_{m,l} h^*_{m',l}] = J_0(2\pi|m-m'| f_D T)$, where $J_0(\cdot)$ is the zeroth-order Bessel function of the first kind and $f_D T$ is the normalized maximum Doppler frequency with $T=(N_c+N_g)T_c$.

2.3. Received signal

The GI is removed from the received signal. The

GI-removed received signal belonging to the mth chip block can be written as

$$r_m(t)=\sum_{l=0}^{L-1}h_{m,l}\hat{s}_m(t-\tau_l)+n_m(t)\,, \qquad (6)$$

where $n_m(t)$ is the additive white Gaussian noise (AWGN) with zero-mean and the variance of $2N_0/T_c$ with N_0 being the one-sided power spectrum density. An N_c-point FFT is applied to decompose $r_m(t)$ into N_c frequency components $\{R_m(k);\ k=0\sim N_c\text{-}1\}$ as

$$\begin{aligned}R_m(k)&=1\big/\sqrt{N_c}\sum_{t=0}^{N_c-1}r_m(t)\exp\left(-j2\pi k\,t/N_c\right)\\&=\sqrt{2E_c/T_c}\,b_mC_m(k)H_m(k)+\Pi_m(k)\end{aligned}, \qquad (7)$$

where $H_m(k)$ is the kth frequency channel gain of the mth block, $\Pi_m(k)$ is the kth frequency component of $n_m(t)$, and $C_m(k)$ is the kth frequency component of the spreading sequence $c_m(t)$. They are given by

$$\begin{cases}H_m(k)=\sum_{l=0}^{L-1}h_{m,l}\exp\left(-j2\pi k\,\tau_l/N_c\right)\\ \Pi_m(k)=1\big/\sqrt{N_c}\sum_{t=0}^{N_c-1}n_m(t)\exp\left(-j2\pi k\,t/N_c\right)\\ C_m(k)=1\big/\sqrt{N_c}\sum_{t=0}^{N_c-1}c_m(t)\exp\left(-j2\pi k\,t/N_c\right)\end{cases} \qquad (8)$$

for $k=0\sim N_c\text{-}1$, where $E[|H_m(k)|^2]=1$, $E[|\Pi_m(k)|^2]=2N_0/T_c$ and $E[|C_m(k)|^2]=1$.

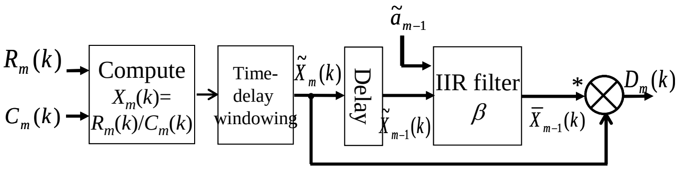


Fig. 2. FDDDE structure.

2.4. FDDDE

The structure for FDDDE is illustrated in Fig. 2. The chip modulation is first removed from the received signal component as

$$\begin{aligned}X_m(k)&=R_m(k)/C_m(k)\\&=\sqrt{2E_c/T_c}\,b_mH_m(k)+\Pi_m(k)/C_m(k)\end{aligned}, \qquad (9)$$

where $\Pi_m(k)/C_m(k)$ is a new zero-mean Gaussian noise with variance $2N_0/T_c$ for the given $C_m(k)$. $X_m(k)$ is noisy and therefore, the delay-time domain windowing technique [10] is applied to $X_m(k)$ to reduce the noise. Firstly, an N_c-point IFFT is applied to $X_m(k)$ to obtain the delay time-domain sequence, which is a noisy instantaneous channel impulse response modulated by differentially encoded data symbol. Since the real channel impulse response is assumed to be present only within the GI length (i.e., $\tau=0\sim N_g\text{-}1$), while the noise is distributed over the entire delay-time range (i.e., $\tau=0\sim N_c\text{-}1$), the noise can be suppressed by zero-padding beyond GI [10]. Then, after applying an N_c-point FFT, $\tilde{X}_m(k)$ is obtained as

$$\tilde{X}_m(k)=\sqrt{2E_c/T_c}\,b_mH_m(k)+\tilde{\Pi}_m(k)\,, \qquad (10)$$

where $\tilde{\Pi}_m(k)$ is a zero-mean Gaussian noise with reduced variance $2\sigma^2=2(N_0/T_c)(N_g/N_c)$ (see details in Sect. 3).

To recover the data symbol a_m, we need the reference signal. Remembering $a_{u,m}=b_{u,m}b^*_{u,m\text{-}1}$, $\tilde{X}_{u,m-1}(k)$ can be used as the reference for FDDDE. However, since $\tilde{X}_{m-1}(k)$ is still noisy, we apply a simple IIR filter to reduce the noise. As shown in Fig. 2, the first-order IIR filter with forgetting factor β ($0\le\beta\le1$) is used to improve the reference signal. The filter output $\bar{X}_{m-1}(k)$ is given by

$$\bar{X}_{m-1}(k)=\beta\bar{X}_{m-2}(k)\tilde{a}_{m-1}+(1-\beta)\tilde{X}_{m-1}(k) \qquad (11)$$

for $m>1$. The initial condition is $\bar{X}_0(k)=\tilde{X}_0(k)$ and $\tilde{a}_{m-1}$ is the decision feedback from the previous block. β is an important design parameter to trade off between the noise reduction and the tracking ability against fading. When $\beta=0$, $\bar{X}_{m-1}(k)$ becomes $\tilde{X}_{m-1}(k)$; a high tracking ability against fading can be achieved, but the noise reduction is insufficient. In a slow fading channel, β should be close to 1 so that the noise can be effectively reduced; while, in a fast fading channel, β should be smaller to achieve better tracking ability. There exists an optimum value in β, which depends on the received signal-to-noise ratio (SNR) and the Doppler spread.

Then FDDDE [5] is performed on each frequency component as follows:

$$D_m(k)=\bar{X}^*_{m-1}(k)\tilde{X}_m(k)\,, \qquad (12)$$

The sum of the FDDDE outputs over all frequencies, given by

$$D_m=\sum_{k=0}^{N_c-1}D_m(k)\,, \qquad (13)$$

is the decision variable to detect d_m. The detected symbol $\tilde{d}_m$ is remodulated to get $\tilde{a}_m$.

3. Analysis

In this section, the BER performance for the FDDDE reception of differentially encoded DS-CDMA signals in a time- and frequency-selective Rayleigh fading channel is developed. In Sect. 3.1, a closed-form BER expression is derived for FDDDE without IIR filtering and then an approximate analysis is presented for the case of FDDDE with IIR filtering, taking into account the error propagation due to decision feedback.

3.1. FDDDE without IIR filtering (β=0)

We assume DQPSK data modulation and assume all "1" transmission (i.e., $a_m=(1+j)/\sqrt{2}$) without loss of generality. The analysis method given in [6, Appendix B] can be applied to derive the bit error rate (BER). The BER is given as

$$P_{b,FDDDE} = \text{Prob}\{\text{Re}[D_m] < 0\}. \tag{14}$$

In the frequency-domain, the channel gains $\{H_m(k)\}$ are correlated each other and their frequency correlation function $E[H_m(k)H_m^*(k+i)]$ depends on the channel power delay profile shape. In this paper, an approximate analysis is presented assuming that the channel gains of N adjacent frequencies are strongly correlated, i.e., $H_m(k) \approx H_m(k'N)$ with $k' = \lfloor k/N \rfloor$. N is related to the ratio of the coherence bandwidth B_c over the bandwidth $1/T_c$. τ_{rms} and B_c are inversely proportional to one another [12], that is $B_c=1/(\varepsilon\tau_{rms})$ [13], where ε depends on the frequency correlation. For instance, ε=4~6 if B_c is defined as the bandwidth over which the frequency correlation is at least 0.5 [12]. In this paper, we approximate that the channel gains over the coherent bandwidths are the same, that is, $H_m(k+i) \approx H_m(k)$ if $i \le N$, where $N=B_cT_cN_c = \lfloor (N_c/\varepsilon)(\tau_{rms}/T_c)^{-1} \rfloor$.

Replacing Eq. (10) into Eq. (13) and neglecting the noise-by-noise component, we have

$$\begin{aligned}\text{Re}[D_m] &= \text{Re}\left[\sum_{k=0}^{N_c-1} \tilde{X}_m(k)\tilde{X}_{m-1}^*(k)\right] \\ &\approx \text{Re}\left[\sum_{k'=0}^{\frac{N_c}{N}-1}\left\{\begin{array}{l}\left(\sqrt{\dfrac{2E_cN}{T_c}}b_mH_m(k'N) + \dfrac{1}{\sqrt{N}}\displaystyle\sum_{i=0}^{N-1}\tilde{\Pi}_m(k'N+i)\right)\times \\ \left(\sqrt{\dfrac{2E_cN}{T_c}}b_{m-1}H_{m-1}(k'N) + \dfrac{1}{\sqrt{N}}\displaystyle\sum_{i=0}^{N-1}\tilde{\Pi}_{m-1}(k'N+i)\right)^*\end{array}\right\}\right]\end{aligned} \tag{15}$$

for a large E_c/N_0 region. Noting that the noise components $\{\tilde{\Pi}_m(k);\ k=0\sim N_c\text{-}1\}$ also correlated, we use the approximation $\tilde{\Pi}_m(i) \approx \tilde{\Pi}_m(pN')$, where $p = \lfloor i/N' \rfloor$. N' depends on N_g and is given as $N' = N_c\sqrt{12}/(\varepsilon N_g)$. Since N_g is assumed to be longer than the maximum time delay, we have $N' < N$. Therefore, we have

$$1/\sqrt{N}\sum_{i=0}^{N-1}\tilde{\Pi}_m(k'N+i) \approx \sqrt{N'}\hat{\Pi}_m(k'N), \tag{16}$$

where $\hat{\Pi}_m(k'N)$ is a new zero-mean Gaussian variable with variance of $2\sigma^2 = 2N_0/T_c(N_g/N_c)$.

Eq. (15) can be further approximated as

$$\text{Re}[D_m] \approx \text{Re}\left[\sum_{k=0}^{N_c/N-1} X_kY_k^*\right], \tag{17}$$

where

$$\begin{cases} X_k = \sqrt{2E_cN/T_c}\,b_mH_m(kN) + \sqrt{N'}\hat{\Pi}_m(kN) \\ Y_k = \sqrt{2E_cN/T_c}\,b_{m-1}H_{m-1}(kN) + \sqrt{N'}\hat{\Pi}_{m-1}(kN) \end{cases}. \tag{18}$$

Eq. (17) is a special case of a general quadratic form defined in [6, Appendix B] as

$$\text{Re}[D_m] = \sum_{k=0}^{G-1}\left[A|X_k|^2 + B|Y_k|^2 + CX_k^*Y_k + C^*X_kY_k^*\right] \tag{19}$$

with A=0, B=0, C=0.5, and $G=N_c/N$. Here, G pairs of $\{X_k, Y_k\}$ are mutually statistically independent and identically distributed Gaussian variables. G is defined as the equivalent frequency diversity order and equal to $G=\varepsilon(\tau_{rms}/T_c)$. If $B_s>B_c$, we have $G>1$. From [6, Appendix B], we obtain the following BER expression as

$$P_{b,FDDDE} \approx \frac{1}{(1+v_2/v_1)^{2G-1}}\sum_{k=0}^{G-1}\binom{2G-1}{k}\left(\frac{v_2}{v_1}\right)^k, \tag{20}$$

where

$$\frac{v_2}{v_1} = \frac{\sqrt{2m_{xx}m_{yy} - |m_{xy}|^2} + |m_{xy}|}{\sqrt{2m_{xx}m_{yy} - |m_{xy}|^2} - |m_{xy}|} \tag{21}$$

and m_{xx}, m_{yy}, and m_{xy} are given by

$$\begin{cases} m_{xx} = E[|X_k|^2] = \left(\dfrac{2E_s}{T_cG}\right)\left[1+\sigma_x^2\right] \\ m_{yy} = E[|Y_k|^2] = \left(\dfrac{2E_s}{T_cG}\right)\left[1+\sigma_y^2\right] \\ m_{xy} = E[X_kY_k^*] = \left(\dfrac{2E_s}{T_cG}\right)a_mE[H_m(k)H_{m-1}^*(k)] \end{cases} \tag{22}$$

with $\sigma_x^2 = \sigma_y^2 = \sqrt{12}(\tau_{rms}/T_c)(E_s/N_0)^{-1}$. Since $E[H_m(k)H^*_{m-1}(k)]=J_0(2\pi f_DT)$ and $a_m = (1+j)/\sqrt{2}$, we have

$$\frac{v_2}{v_1} = \frac{\sqrt{2(1+\sigma_x^2)(1+\sigma_y^2) - J_0^2(2\pi f_DT)} + J_0(2\pi f_DT)}{\sqrt{2(1+\sigma_x^2)(1+\sigma_y^2) - J_0^2(2\pi f_DT)} - J_0(2\pi f_DT)}. \tag{23}$$

Finally, substituting Eq. (23) into Eq. (20), we get the BER performance for FDDDE without IIR filtering. If we set $\sigma_y^2 = 0$ and f_DT=0 in Eq. (23), Eq. (20) gives the BER of coherent FDE with ideal channel estimation.

3.2. FDDDE with IIR filtering ($\beta \neq 0$)

If the IIR filter with decision feedback is applied to improve FDDDE. The IIR filter output $\bar{X}_{m-1}(k)$ is used as the FDDDE weight $w_m(k)$. Since β<1, $\bar{X}_{m-1}(k)$ in Eq. (11) can be well approximated, for a large value of m, as

$$\bar{X}_{m-1}(k) \approx (1-\beta)\sum_{i=0}^{m-2}\beta^i\tilde{X}_{m-i-1}(k)\prod_{j=1}^{i}\tilde{a}_{m-j}. \tag{24}$$

Substituting Eq. (10) into the above, we have

$$\bar{X}_{m-1}(k) \approx \sqrt{\frac{2E_c}{T_c}}(1-\beta)\sum_{i=0}^{m-2}\beta^i b_{m-i-1}H_{m-i-1}(k)\prod_{j=1}^{i}\tilde{a}_{m-j} + (1-\beta)\sum_{i=0}^{m-2}\beta^i\tilde{\Pi}_{m-i-1}(k)\prod_{j=1}^{i}\tilde{a}_{m-j} \quad (25)$$

At first, we do not consider the error propagation and assume perfect decision feedback, i.e., $\tilde{a}_{m-j} = a_{m-j}$. Since

$$b_{m-i-1}\prod_{j=1}^{i}\tilde{a}_{m-j} = b_{m-i-1}\prod_{j=1}^{i}a_{m-j} = b_{m-1}, \quad (26)$$

Eq. (25) becomes

$$\bar{X}_{m-1}(k) \approx \sqrt{\frac{2E_c}{T_c}}b_{m-1}(1-\beta)\sum_{i=0}^{m-2}\beta^i H_{m-i-1}(k) + (1-\beta)\sum_{i=0}^{m-2}\beta^i\tilde{\Pi}_{m-i-1}(k)\prod_{j=1}^{i}a_{m-j} \quad (27)$$

Since IIR filtering is the time-domain process, $\bar{X}_{m-1}(k)$ has the same frequency correlation function as $\tilde{X}_{m-1}(k)$. The BER can be obtained from Eqs. (14) and (17), but, in this case, we use

$$\begin{cases} X_k = \sqrt{\frac{2E_cN}{T_c}}b_m H_m(kN) + \sqrt{N'}\hat{\Pi}_m(kN) \\ Y_k = \sqrt{\frac{2E_cN}{T_c}}(1-\beta)\sum_{i=0}^{m-2}\beta^i H_{m-i-1}(kN) + \sqrt{N'}(1-\beta)\sum_{i=0}^{m-2}\beta^i\hat{\Pi}_{m-i-1}(kN)\prod_{j=1}^{i}a_{m-j} \end{cases} \quad (28)$$

for k=0~G-1. G pairs of $\{X_k, Y_k\}$ are mutually statistically independent and identically distributed. m_{yy} and m_{xy} are given by

$$\begin{cases} m_{yy} = \left(\frac{2E_s}{T_cG}\right)\left(\frac{1-\beta}{1+\beta}\right)\left[1+\sigma_x^2+(1-\beta^2)\sum_{i=0}^{m-2}\sum_{\substack{i'=0 \\ i'\neq i}}^{m-2}\beta^{i+i'}J_0(2\pi f_D T|i-i'|)\right] \\ m_{xy} = b_m\left(\frac{2E_s}{T_cG}\right)(1-\beta)\sum_{i=0}^{m-2}\beta^i J_0[2\pi f_D T(i+1)] \end{cases}. \quad (29)$$

So far, we have assumed the perfect decision feedback. However, once a decision error is produced due to the AWGN, the reference signal is wrongly phase rotated and thus, the next decision is most likely incorrect, but the decision error may not propagate further [11]. As a consequence, one decision error likely results in two symbol errors. Therefore, the approximate BER taking into account the error propagation can be expressed as [11]

$$P_{b,FDDDE-IIR} \approx \frac{2}{(1+v_2/v_1)^{2G-1}}\sum_{k=0}^{G-1}\binom{2G-1}{k}\left(\frac{v_2}{v_1}\right)^k, \quad (30)$$

where $\binom{a}{b} = \frac{a!}{b!(a-b)!}$ is the binomial coefficient and v_2/v_1 is obtained by substituting m_{xx} in Eq. (22) and m_{yy}, m_{xy} in Eq. (29) into Eq. (21).

4. Theoretical and Simulation Results

An L=16-path frequency-selective block Rayleigh fading channel having the exponential power delay profile with decay factor α is assumed. In this paper, the Zadoff-Chu sequence with length SF is used as the spreading sequence $c_m(t)$ in Eq. (2), which is given by $c_m(t) = \exp(j\pi(t^2+2mt)/SF)$ [8]. Since the Zadoff-Chu sequence $c_m(t)$ has a constant amplitude both in the time- and in the frequency-domain [8], that is, $|c_m(t)|=|C_m(k)|=1$, there is no noise enhancement in $X_m(k)$ of Eq. (11). We assume $SF=N_c$=256, N_g=32 and the maximum time delay, max{τ_l}, is less than the GI.

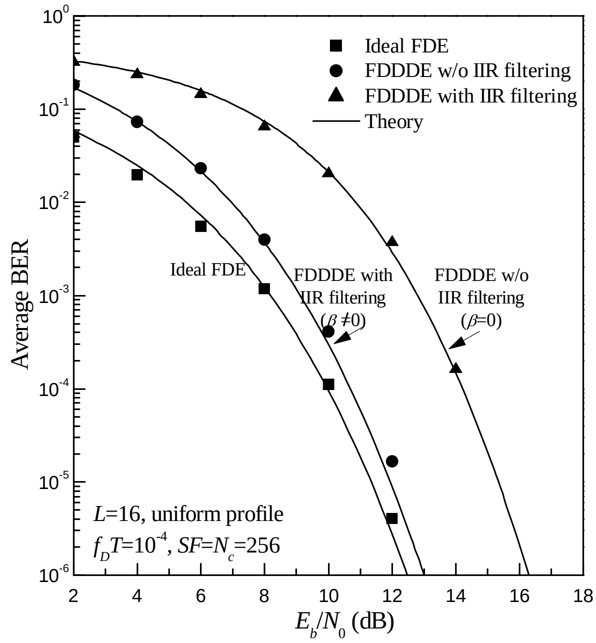


Fig. 4. Comparison between FDE and FDDDE.

Fig. 4 shows the theoretical and computer simulated BER performances as a function of the average received bit energy-to-the AWGN power spectrum density ratio E_b/N_0, defined by $E_b/N_0=0.5(E_c/N_0)SF(1+N_g/N_c)$, in the case of uniform delay profile (α=0dB) when $f_DT=10^{-4}$. The optimum β is used for FDDDE with IIR filtering; the optimal β is found to be β=0.95, 0.95, 0.90 and 0.85 for $f_DT=10^{-4}$, 10^{-3}, 5×10^{-3} and 10^{-2}, respectively. $f_DT=10^{-4}$~10^{-2} corresponds to the vehicle speed of about 7.5~750km/h for a carrier frequency of 5GHz and a chip data rate of 100Mcps. It can be seen that the theoretical BER performance agrees well with the simulated BER performance. The use of an IIR filter significantly improves the BER performance of FDDDE. The E_b/N_0 degradation of FDDDE with IIR filtering from coherent FDE with ideal channel estimation is as small as 0.4dB at BER=10^{-5}.

Fig. 5 plots the BER performance in the case of exponential delay profile with the decay factor α as a parameter when f_DT=10^{-4}. As α increases, the Rayleigh channel becomes less frequency-selective and the BER performances of both coherent FDE and FDDDE with IIR filtering degrade. When α=0dB (τ_{rms}/T_c=4.6), corresponding to the uniform delay profile, both coherent FDE and FDDDE with IIR filtering achieve their best BER performances.

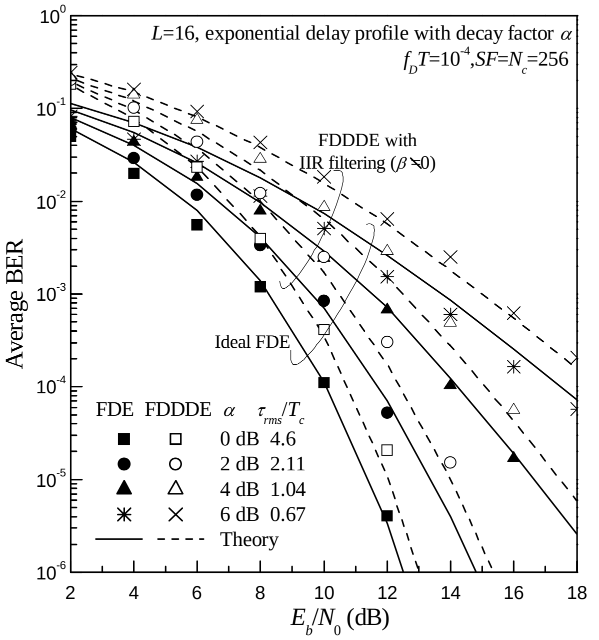


Fig. 5. Impact of τ_{rms}/T_c.

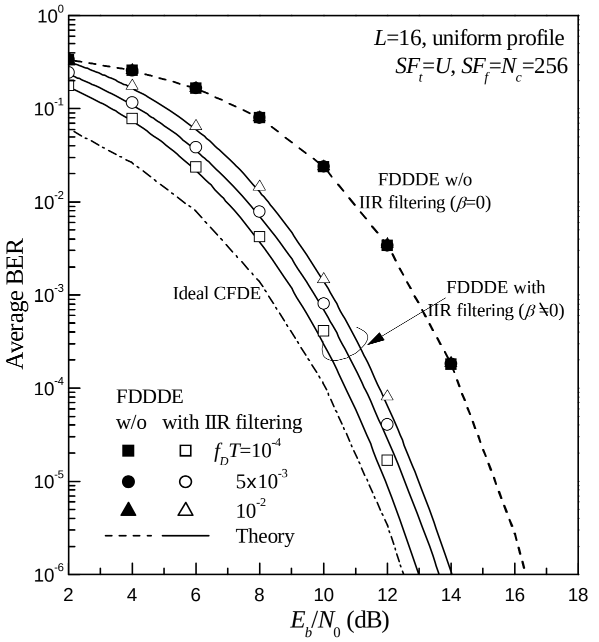


Fig. 6 Impact of f_DT.

Fig. 6 shows the impact of f_DT on the BER performance of DS-CDMA using FDDDE in the case of the uniform delay profile (α=0dB). There is a good agreement between the computer simulated BER performance and the theoretical one. It can be seen that FDDDE without IIR filtering (β=0) is almost insensitive to f_DT, but performs worse than FDDDE with IIR filtering. As f_DT increases, FDDDE with IIR filtering degrades. This is because a high tracking ability against fast fading needs a small value of β, which makes the noise reduction insufficient. However, the E_b/N_0 degradation at BER=10^{-5} from coherent FDE is only 1.5dB, even when f_DT=0.01.

5. Conclusions

In this paper, theoretical analysis of FDDDE for the reception of DQPSK-modulated DS-CDMA signals in a time-frequency selective fading channel was presented. The theoretical BER analysis was confirmed by the computer simulation. It was shown that the BER performance of FDDDE with IIR filtering can approach that of coherent FDE with ideal channel estimation in the case of slow Doppler fading and the E_b/N_0 less from coherent FDE is as small as 0.4dB at BER=10^{-5}. FDDDE with IIR filtering has a high tracking ability against fading and the E_b/N_0 degradation at BER=10^{-5} from coherent FDE is about 1.5dB, even for f_DT=0.01.

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