DS-CDMA MMSE turbo equalization using 2-step maximum likelihood channel estimation

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Abstract—**Frequency-domain equalization (FDE) based on the minimum mean square error (MMSE) criterion and frequency-domain inter-chip interference (ICI) cancellation can be incorporated into turbo decoding (this is called the MMSE turbo equalization in this paper) to improve the bit error rate (BER) performance for direct sequence code division multiple access (DS-CDMA). DS-CDMA MMSE turbo equalization requires accurate channel estimation. The performance of MMSE turbo equalization is sensitive to channel estimation accuracy. Recently, we proposed a 2-step maximum likelihood channel estimation (MLCE) for DS-CDMA with FDE. In this paper, we apply 2-step MLCE to DS-CDMA MMSE turbo equalization and evaluate by computer simulation the achievable BER performance in a frequency-selective block Rayleigh fading channel.**

Keywords-component; DS-CDMA, frequency-domain equalization, frequency-domain ICI cancellation, MMSE turbo equalization, channel estimation

I. INTRODUCTION

A broadband wireless access technique of 100 Mbps to 1 Gbps is required for the 4th generation (4G) mobile communication systems [1]. In the present 3rd generation (3G) systems, direct sequence code division multiple access (DS-CDMA) is adopted as the wireless access technique [2]. However, since the broadband wireless channel is severely frequency-selective, the bit error rate (BER) performance of DS-CDMA with rake combining significantly degrades. Replacing rake combining by frequency-domain equalization (FDE) based on the minimum mean square error (MMSE) criterion can significantly improve the BER performance of DS-CDMA [3, 4]. However, the presence of residual inter-chip interference (ICI) after MMSE-FDE limits the improvement of the BER performance. Frequency-domain ICI cancellation can significantly suppress the residual ICI and bring the BER performance close to the theoretical lower-bound [5]. Frequency-domain ICI cancellation can be incorporated into turbo decoding (we call this MMSE turbo equalization) [6].

FDE requires the accurate estimation of the channel transfer function. Pilot-assisted channel estimation (CE) can be used. Time-domain pilot-assisted CE was proposed for single-carrier transmission in [7]. After the channel impulse response is estimated according to the least-sum-of-squared-error (LSSE) criterion, the channel transfer function is obtained by applying fast Fourier transform (FFT). Frequency-domain pilot assisted CE was proposed in [8], [9]. The received pilot signal is transformed into the frequency-domain pilot signal and then the pilot modulation is removed using zero forcing (ZF) or least square (LS) technique. As the pilot signal, the Chu sequence [10] having the constant amplitude in both time- and frequency-domain is used. However, the number of Chu sequences is limited. For example, it is only 128 for the case of 256-bit period [10].

PN sequences can be used as the pilot. Using a partial sequence taken from a long PN sequence, a very large number of pilots can be generated. However, since the frequency spectrum of the partial PN sequence is not constant, the use of ZF channel estimation (ZFCE) produces the noise enhancement [11]. The noise enhancement can be mitigated by using the minimum mean square error based channel estimation (MMSECE) [11]. Using MMSECE, the channel estimation accuracy is almost insensitive to the used pilot sequence. We also proposed a 2-step maximum likelihood channel estimation (MLCE) to further improve estimation accuracy [12]. In this paper, we apply 2-step MLCE to DS-CDMA MMSE turbo equalization and evaluate by computer simulation the BER performance in a frequency-selective block Rayleigh fading channel. It is shown that the 2-step MLCE improves the BER performance compared to the conventional decision feedback MMSECE at the cost of increasing complexity of the receiver.

The rest of the paper is organized as follows. Sect. II presents transmission system model. In sect. III, 2-step MLCE is presented. In sect. IV, the achieved BER performance in a frequency-selective block Rayleigh fading channel is evaluated by computer simulation. Sect. V offers some conclusions.

II. TRANSMISSION SYSTEM MODEL

A. Overall transmission system model

The transmission system model for DS-CDMA MMSE turbo equalization is illustrated in Fig 1. Throughout the paper, the chip-spaced discrete-time signal representation is used. At the transmitter, after turbo encoding and channel interleavering, a coded binary data sequence of $(U/SF)\cdot N_c\cdot\log_2 M$ bits is transformed into the nth data-modulated symbol block $\{d_n(m'); m'=0\sim(U/SF)\cdot N_c-1\}$, $n=0\sim N-1$, where SF is the spreading factor, M is the modulation level and N_c is the FFT block size. The resulting data modulated symbol block $\{d_n(m')\}$ is transformed, by serial-to-parallel (S/P) converter, into U parallel sub-blocks $\{d_{n,u}(m); m=0\sim N_c/SF-1\}$, $u=0\sim U-1$.

The uth sub-block $\{d_{n,u}(m)\}$ is spread by multiplying it with an orthogonal spreading sequence $\{c_u(t); t=0\sim SF-1\}$. The resultant U chip-block of N_c chips each are added and further multiplied by a common scramble sequence $\{c_{scr}(t); t=\ldots,-1,0,1,\ldots\}$ to make the resultant multicode DS-CDMA chip-block like white-noise. The last N_g chips of each N_c chip-block is copied as a cyclic prefix and inserted into the guard interval (GI) placed at the beginning of each chip-block, as illustrated in Fig. 2. For channel estimation, one pilot chip-block is transmitted every N–1 data chip-block (one pilot chip-block plus N–1 data chip-block constitutes a frame) as shown in Fig. 3.

The GI-inserted chip-block is transmitted over a frequency-selective fading channel and is received at a receiver. After the removal of the GI, the received chip-block is decomposed by N_c-point fast Fourier transform (FFT) into N_c frequency components. After MMSE-FDE, ICI cancellation is performed in the frequency-domain. N_c-point inverse FFT (IFFT) is applied to obtain the time-domain received chip-block for de-spreading. After performing channel deinterleaving and maximum *a posteriori* (MAP) decoding, a sequence of soft symbol replicas is generated using the decoder output (log likelihood ratio (LLR)). This replica is fedback for MMSE weight updating and ICI cancellation. A series of MMSE-FDE, ICI cancellation, de-spreading and MAP decoding is repeated a sufficient number of times before obtaining the final received data.

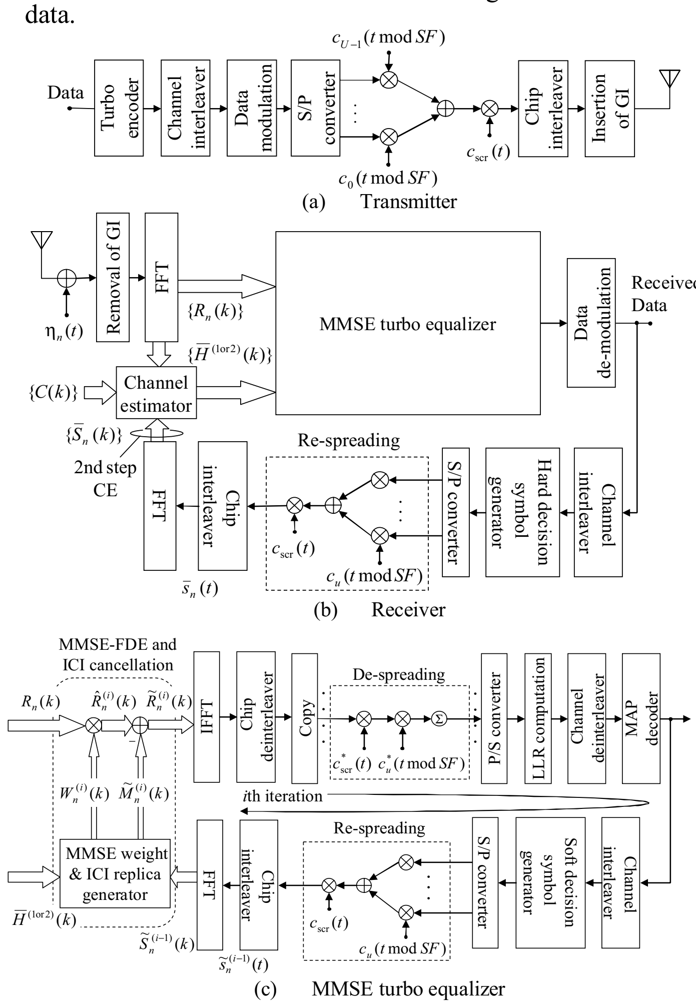

Figure 1. Transmitter/receiver structure for multicode DS-CDMA MMSE turbo equalization.

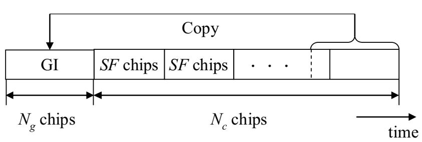

Figure 2. Chip-block structure.

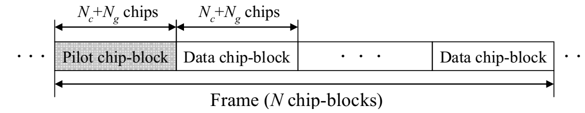

Figure 3. Frame structure.

B. Signal representation

The nth chip-block $\{\tilde{s}_n(t);\ t=0\sim N_c-1\}$, $n=0\sim N-1$, can be expressed, using the equivalent lowpass representation, as

$$\tilde{s}_n(t)=\sqrt{2P}s_n(t) \tag{1}$$

with

$$s_n(t)=\left[\sum_{u=0}^{U-1} d_{n,u}\left(\left\lfloor \frac{t}{SF}\right\rfloor\right)c_u(t \bmod SF)\right]c_{\mathrm{scr}}(t), \tag{2}$$

where P is the transmit power and $\lfloor x \rfloor$ represents the largest integer smaller than or equal to x. After inserting the GI of N_g chips, the nth chip-block is transmitted. The propagation channel is assumed to be a frequency-selective block fading channel having chip-spaced L discrete paths, each subjected to independent fading. We assume that the channel gains stay constant over N blocks. The channel impulse response $h(\tau)$ can be expressed as

$$h(\tau)=\sum_{l=0}^{L-1} h_l \delta(\tau-\tau_l), \tag{3}$$

where h_l and τ_l are the complex-valued path gain and time delay of the lth path ($l=0\sim L-1$), respectively , with $\sum_{l=0}^{L-1}E[|h_l|^2]=1$ ($E[.]$ denotes the ensemble average operation). In this paper, we assume that the maximum time delay difference $\tau_{L-1}-\tau_0$ of the channel is shorter than the GI length.

The nth received chip-block $\{r_n(t);\ t=-N_g\sim N_c-1\}$ can be expressed as

$$r_n(t)=\sqrt{2P}\sum_{l=0}^{L-1} h_l s_n(t-\tau_l)+\eta_n(t), \tag{4}$$

where $\eta_n(t)$ is a zero-mean complex Gaussian process with variance $2N_0/T_c$ with T_c and N_0 being respectively the chip duration and the single-sided power spectrum density of the zero-mean additive white Gaussian noise (AWGN).

C. MMSE-FDE & ICI cancellation

A joint MMSE-FDE and ICI cancellation is repeated in an iterative fashion. Below, the ith iteration ($i=0\sim I-1$) is described.

After the removal of the GI, the received chip-block is decomposed by N_c-point FFT into N_c frequency components. The kth frequency component can be written as

$$\begin{aligned} R_n(k)&=\sum_{t=0}^{N_c-1} r_n(t)\exp\left(-j2\pi k\frac{t}{N_c}\right) \\ &=H(k)S_n(k)+\Pi_n(k) \end{aligned}, \tag{5}$$

where $H(k)$ is the channel gain, $S_n(k)$ is the signal component, and $\Pi_n(k)$ is the noise due to the AWGN. They are given by

$$\begin{cases} S_n(k)=\sum_{t=0}^{N_c-1} s_n(t)\exp\left(-j2\pi k\dfrac{t}{N_c}\right) \\ H(k)=\sqrt{2P}\sum_{l=0}^{L-1} h_l \exp\left(-j2\pi k\dfrac{\tau_l}{N_c}\right) \end{cases}. \quad (6)$$

One-tap MMSE-FDE is carried out as

$$\hat{R}_n^{(i)}(k)=W^{(i)}(k)R_n(k), \quad (7)$$

where $W^{(i)}(k)$ is the MMSE-FDE weight at ith iteration and is given by [5]

$$W^{(i)}(k)=\frac{H^*(k)}{\rho^{(i-1)}\left|H(k)\right|^2+2\sigma^2} \quad (8)$$

with $2\sigma^2$ ($=2N_0N_c/T_c$) being the variance of $\Pi_n(k)$ and $\rho^{(i-1)}$ is the interference coefficient.

After MMSE-FDE, ICI cancellation is performed as

$$\tilde{R}_n^{(i)}(k)=\hat{R}_n^{(i)}(k)-\tilde{M}_n^{(i)}(k), \quad (9)$$

where $\tilde{M}_n^{(i)}(k)$ is the residual ICI replica and given by

$$\tilde{M}_n^{(i)}(k)=\begin{cases} 0 & \text{for } i=0 \\ \{H(k)W^{(i)}(k)-A^{(i)}\}\tilde{S}_n^{(i-1)}(k) & \text{for } i\geq 1 \end{cases}. \quad (10)$$

In the above, $\tilde{S}_n^{(i-1)}(k)$ is the kth frequency component of the chip-block replica $\{\tilde{s}_n^{(i-1)}(t)\}$ generated after the (i–1)th iteration, and $A^{(i)}$ is given by

$$A^{(i)}=\frac{1}{N_c}\sum_{k=0}^{N_c-1}H(k)W^{(i)}(k). \quad (11)$$

$H(k)$ and σ^2 are unknown at the receiver and need to be estimated. $H(k)$ is estimated by using 2-step MLCE [12]. σ^2 can be estimated following to [11].

D. De-spreading

N_c-point IFFT is applied to transform the frequency-domain signal $\{\tilde{R}_n^{(i)}(k)$; k=0~N_c–1} into the time-domain chip-block $\{\tilde{r}_n^{(i)}(t)$; t=0~N_c–1} as

$$\tilde{r}_n^{(i)}(t)=\frac{1}{N_c}\sum_{k=0}^{N_c-1}\tilde{R}_n^{(i)}(k)\exp\left(j2\pi t\frac{k}{N_c}\right). \quad (12)$$

De-spreading is carried out on $\{\tilde{r}_n^{(i)}(t)$; t=0~N_c–1} to obtain the uth code's decision variable associated with $d_{n,u}(m)$, giving

$$\hat{d}_{n,u}^{(i)}(m)=\frac{1}{SF}\sum_{t=mSF}^{(m+1)SF-1}\tilde{r}_n^{(i)}(t)c_u^*(t \bmod SF)c_{\mathrm{scr}}^*(t). \quad (13)$$

E. MAP decoding, ICI replica generation & MMSE-FDE weight updating

After de-spreading, the *a priori* log-likelihood ratio (LLR) $\{\lambda_n^{(i)}(b_{m',x})$; m'=0~(U/SF)·N_c–1, x=0~$\log_2 M$–1}, where $b_{m',x}$ is the xth bit in the m'th symbol, is obtained by performing the maximum *a posteriori* (MAP) decoding [6, 13]. The soft symbol replica $\{\tilde{d}_n^{(i)}(m')$; m'=0~(U/SF)·N_c–1} is generated using *a priori* LLR as [14]

$$\tilde{d}_n^{(i)}(m')=\begin{cases} \dfrac{1}{\sqrt{2}}\tanh\left(\dfrac{\lambda_n^{(i)}(b_{m',0})}{2}\right)+j\dfrac{1}{\sqrt{2}}\tanh\left(\dfrac{\lambda_n^{(i)}(b_{m',1})}{2}\right) \\ \qquad\qquad\qquad\qquad \text{for QPSK} \\ \dfrac{1}{\sqrt{10}}\tanh\left(\dfrac{\lambda_n^{(i)}(b_{m',0})}{2}\right)\left\{2+\tanh\left(\dfrac{\lambda_n^{(i)}(b_{m',1})}{2}\right)\right\} \\ +j\dfrac{1}{\sqrt{10}}\tanh\left(\dfrac{\lambda_n^{(i)}(b_{m',2})}{2}\right)\left\{2+\tanh\left(\dfrac{\lambda_n^{(i)}(b_{m',3})}{2}\right)\right\} \\ \qquad\qquad\qquad\qquad \text{for 16QAM} \end{cases}, \quad (14)$$

After S/P conversion to transform $\{\tilde{d}_n^{(i)}(m')$; m'=0~(U/SF)·N_c–1} into $\{\tilde{d}_{n,u}^{(i)}(m)$; m=0~N_c/SF}, u=0~U–1, the transmit chip-block replica $\{\tilde{s}_n^{(i)}(t)\}$ is obtained by re-spreading $\{\tilde{d}_{n,u}^{(i)}(m)\}$. Then, the kth frequency component of the transmitted chip-block replica $\{\tilde{S}_n^{(i)}(k)\}$ is obtained by applying N_c-point FFT to $\{\tilde{s}_n^{(i)}(t)\}$. The residual ICI replica is obtained using $\tilde{S}_n^{(i)}(k)$ as in Eq. (10).

The FDE weight is updated using Eq. (8). The interference coefficient $\rho^{(i)}$ is computed using

$$\rho^{(i)}=SF\sum_{m=0}^{N_c/SF-1}\sum_{u=0}^{U-1}\left\{E\left[\left|d_{n,u}(m)\right|^2\right]-\left|\tilde{d}_{n,u}^{(i)}(m)\right|^2\right\}, \quad (15)$$

where $\rho^{(-1)}=UN_c$, $E[|d_{n,u}(m)|^2]$ is the expectation of $|d_{n,u}(m)|^2$ when $\{r_n(t)$; t=–N_g~N_c–1} is received and computed as follows. First, the expectation of the symbol block $E[|d_n(m')|^2]$ is obtained as [14]

$$E\left[|d_n(m')|^2\right]=\begin{cases} 1 & \text{for QPSK} \\ \dfrac{4}{10}\tanh\left(\dfrac{\lambda_n^{(i)}(b_{m',1})}{2}\right)+\dfrac{4}{10}\tanh\left(\dfrac{\lambda_n^{(i)}(b_{m',3})}{2}\right) \\ +1 & \text{for 16QAM} \end{cases}. \quad (16)$$

Then, $\{E[|d_n(m')|^2]$; m'=0~(U/SF)·N_c–1} is transformed, by S/P converter, into $\{E[|d_{n,u}(m)|^2]$; m=0~N_c/SF}, u=0~U–1.

After the (I–1)th iteration, the LLR of the information bit obtained by MAP decoding is used to recover the transmitted data.

III. 2-STEP MAXIMUM LIKELIHOOD CHANNEL ESTIMATION[12]

2-step MLCE is illustrated in Fig. 4. We apply the pilot-assisted MMSECE [11] and delay time-domain windowing technique [15, 16] to obtain the instantaneous channel estimate $\bar{H}^{(1)}(k)$. The MMSE-FDE weight and ICI replica are computed using Eqs. (8) and (10) with replacing $H(k)$ by $\bar{H}^{(1)}(k)$. A series of MMSE-FDE, ICI cancellation, de-spreading and MAP decoding is repeated I–1 times, followed by tentative data symbol decision.

Next, using N chip-blocks (one received pilot chip-block and N–1 data chip-block replicas), the maximum likelihood channel estimation is performed and the channel estimate $\bar{H}^{(2)}(k)$ is obtained. The MMSE-FDE weight and ICI replica are computed using Eqs. (8) and (10) with replacing $H(k)$ by $\bar{H}^{(2)}(k)$ once again. A series of MMSE-FDE, ICI cancellation, de-spreading and MAP decoding is iterated I–1 times, and the final received data is obtained.

A. 1st step

The kth frequency component of the received pilot chip-block (n=0) can be represented as

$$R_0(k)=H(k)C(k)+\Pi_0(k)\,, \tag{17}$$

where $C(k)$ is the kth frequency component of the transmitted pilot chip-block $\{\sqrt{U}c(t)\,;\ t=0\sim N_c-1\}$ with $|c(t)|=1$ (the pilot power is set to UP to keep it the same as the U-order code-multiplexed data chip-block power). $C(k)$ is given by

$$C(k)=\sqrt{U}\sum_{t=0}^{N_c-1}c(t)\exp\left(-j2\pi k\frac{t}{N_c}\right). \tag{18}$$

Using MMSECE, the instantaneous channel gain estimate $\tilde{H}^{(1)}(k)$ is obtained as

$$\tilde{H}^{(1)}(k)=X(k)R_0(k)\,, \tag{19}$$

where

$$X(k)=\frac{C^*(k)}{\left|C(k)\right|^2+(P/\sigma^2)^{-1}} \tag{20}$$

is the reference to remove the pilot modulation [11]. The signal power P and the noise power σ^2 can be estimated following to [11].

The instantaneous channel gain estimate $\{\tilde{H}^{(1)}(k)\,;\ k=0\sim N_c-1\}$ obtained from the received pilot chip-block is noisy. The noise can be suppressed by applying delay time-domain windowing technique [15, 16]. $\{\tilde{H}^{(1)}(k)\,;\ k=0\sim N_c-1\}$ is transformed by N_c-point IFFT into the instantaneous channel impulse response $\{\tilde{h}^{(1)}(\tau)\,;\ \tau=0\sim N_c-1\}$ as

$$\tilde{h}^{(1)}(\tau)=\frac{1}{N_c}\sum_{k=0}^{N_c-1}\tilde{H}^{(1)}(k)\exp\left(j2\pi\tau\frac{k}{N_c}\right). \tag{21}$$

The actual channel impulse response is present only within the GI length, while the noise is spread over an entire delay-time range. Replacing $\tilde{h}^{(1)}(\tau)$ with zero's for $N_g\le\tau\le N_c-1$ and applying N_c-point FFT, the improved channel gain estimate $\{\bar{H}^{(1)}(k)\,;\ k=0\sim N_c-1\}$ is obtained as

$$\bar{H}^{(1)}(k)=\sum_{\tau=0}^{N_g-1}\tilde{h}^{(1)}(\tau)\exp\left(-j2\pi k\frac{\tau}{N_c}\right). \tag{22}$$

The MMSE-FDE weight and ICI replica are computed using Eqs. (8) and (10) with replacing $H(k)$ by $\bar{H}^{(1)}(k)$. A series of MMSE-FDE, ICI cancellation, de-spreading and MAP decoding is iterated I–1 times, followed by tentative data symbol decision.

The tentatively detected sub-block $\{\bar{d}_{n,u}(m)\,;\ m=0\sim N_c/SF-1\}$, $u=0\sim U-1$, is spread to obtain the transmitted chip-block replica $\{\bar{s}_n(t)\,;\ t=0\sim N_c-1\}$ as

$$\bar{s}_n(t)=\left[\sum_{u=0}^{U-1}\bar{d}_{n,u}\left(\left\lfloor\frac{t}{SF}\right\rfloor\right)c_u(t \bmod SF)\right]c_{\mathrm{scr}}(t)\,. \tag{23}$$

Applying N_c-point FFT to $\{\bar{s}_n(t)\}$, the replica of the kth frequency component of the transmitted chip-block is obtained as

$$\bar{S}_n(k)=\sum_{t=0}^{N_c-1}\bar{s}_n(t)\exp\left(-j2\pi k\frac{t}{N_c}\right). \tag{24}$$

B. 2nd step

Using the kth frequency components of the pilot chip-block and N–1 transmitted chip-block replicas $\{\bar{S}_n(k)\,;\ n=1\sim N-1\}$, the maximum likelihood estimation [12] is performed as

$$\tilde{H}^{(2)}(k)=\frac{R_0(k)C^*(k)+\sum_{n=1}^{N-1}R_n(k)\bar{S}_n^*(k)}{\left|C(k)\right|^2+\sum_{n=1}^{N-1}\left|\bar{S}_n(k)\right|^2}\,. \tag{25}$$

By applying delay time-domain windowing technique to $\{\tilde{H}^{(2)}(k)\,;\ k=0\sim N_c-1\}$ as in the 1st step, the improved channel gain estimate $\{\bar{H}^{(2)}(k)\,;\ k=0\sim N_c-1\}$ is obtained. The MMSE-FDE weight and ICI replica are computed using Eqs. (8) and (10) with replacing $H(k)$ by $\bar{H}^{(2)}(k)$ once again. A series of MMSE-FDE, ICI cancellation, de-spreading and MAP decoding is iterated I–1 times, and the final received data is obtained.

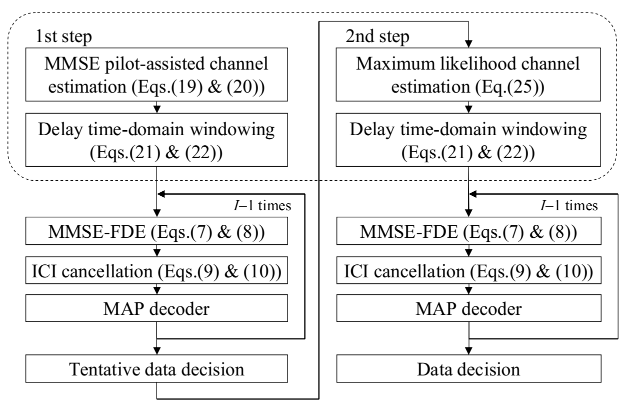

Figure 4. 2-step MLCE.

IV. COMPUTER SIMULATION

The simulation condition is shown in Table 1. We assume an FFT block size of N_c=256 chips and a GI of N_g=32 chips. One pilot chip-block is transmitted every 15 data chip-block (i.e., N=16). We assume the spreading factor SF=16, an L=16-path frequency-selective block Rayleigh fading channel having uniform power delay profile. A turbo encoder with two (13, 15) RSC encoders and Log-MAP decoder is used. The number of iterations in the turbo equalization is assumed to be I=6 (i.e.,

i=0~5) since I=6 is found by computer simulation to be enough to get a good BER performance [6].

The simulated BER performance of multicode DS-CDMA using MMSE turbo equalization is plotted in Fig. 5 as a function of the average received bit energy-to-AWGN noise power spectrum density ratio $E_b/N_0=(1/\log_2 M)(P\cdot SF\cdot T_c/N_0)(1+N_g/N_c)\{N/(N-1)\}R^{-1}$. Also plotted are the BER performances using the pilot-assisted decision feedback MMSECE [11] and that with ideal CE for comparison. When the coding rate is R=1/2 (Fig5. (a)), with the pilot-assisted decision feedback MMSECE, the E_b/N_0 loss of about 0.8 dB from the ideal CE can be seen for BER=10^{-4}. This E_b/N_0 loss includes a pilot insertion loss of 0.28 dB. The use of 2-step MLCE improves the BER performance and reduces the E_b/N_0 loss to about only 0.5 dB for QPSK and 16QAM.

When the coding rate is R=3/4 (Fig5. (b)), the same performance improvement as when R=1/2 can be obtained.

TABLE I. SIMULATION CONDITION

Transmitter	Data modulation	QPSK, 16QAM
	FFT block size	$N_c = 256$
	Guard interval length	$N_g = 32$
	Spreading sequence	Product of Walsh sequence and PN sequence
	Spreading factor	$SF = U = 16$
	Pilot chip sequence	PN sequence
Channel	Fading	Frequency-selective block Rayleigh
	Power delay profile	L=16-path uniform power delay profile
Turbo coding and equalization	Coding rate	R = 1/2, 3/4
	Encoder	(13,15) RSC encoder
	Decoder	Log-MAP decoding
	Number of iterations	I = 6
Receiver	Frequency-domain equalization	MMSE
	Channel estimation	2-step MLCE

V. CONCLUSIONS

In this paper, we applied 2-step MLCE to DS-CDMA MMSE turbo equalization and evaluated by computer simulation the BER performance improvement in a frequency-selective block Rayleigh fading channel. The 2-step MLCE improves the BER performance compared to the conventional decision feedback MMSECE. The required E_b/N_0 loss for BER=10^{-4} from the ideal CE is about 0.5 dB (about 0.28 dB of which is due to the pilot insertion) at the cost of increasing complexity of the receiver.

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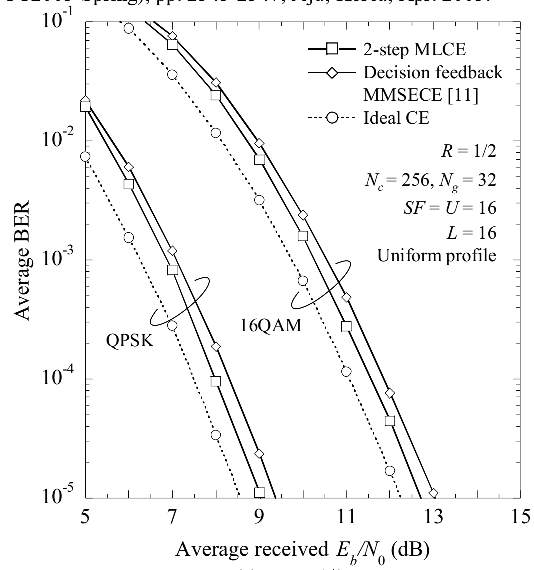
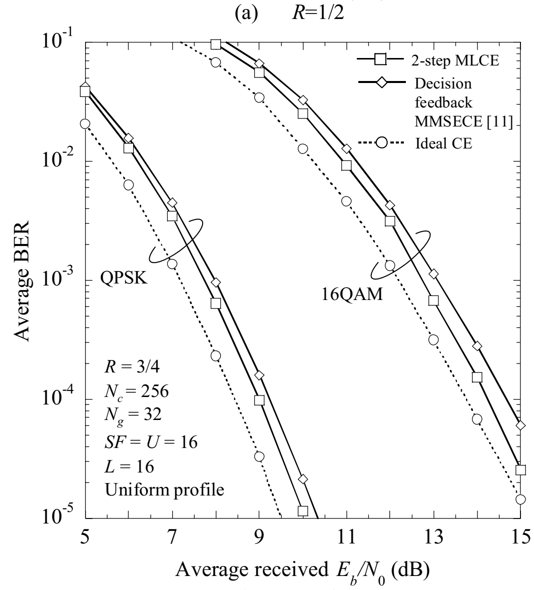


(a) R=1/2



(b) R=3/4

Figure 5. Average BER performance.