DS-CDMA 시스템에서 반복적인 구조를 갖는 간섭제거기와 복호기

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Iterative Interference Cancellation and Decoding for DS-CDMA

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ABSTRACT

In this paper, we propose an iterative interference cancellation and decoding scheme for direct-sequence code-division multiple access (DS-CDMA) system. The proposed scheme is reached to single user performance only when second iteration is completed. The major improvement of the bit error probability (BER) performance is achieved by iterative multi-user interference (MUI) canceller using the soft estimated value of maximum a posteriori (MAP) decoder output. Also the minor improvement of the BER performance is achieved by iterative MAP decoder using a posteriori probability (APP).

요 약

본 논문에서는 DS-CDMA 시스템에서 반복적인 간섭제거와 복호화를 수행하는 수신기 구조를 제안하였다. 제안된 구조를 갖는 수신기는 두 번의 반복적인 간섭제거와 복호화를 통해서 단일 사용자 성능에 근접하는 성능을 보여줌을 시뮬레이션을 통해서 보여준다. 주요한 BER 성능향상 요인으로는 최대사후확률 복호기의 출력을 반복적인 구조를 갖는 다중사용자 간섭제기기에 이용함으로서 이루어지며, 또한 보충적인 성능향상요인으로 최대사후확률 복호기의 출력을 다시 사전확률에 이용함으로써 이루어진다.

Ⅰ. 서 론

Wireless and mobile communications have been one of the fastest-growing fields in the electronics and telecommunications industry over the past ten years. Mobile communication systems are broadly divided into the first, second, and third generation of services, whether the system is analog or digital, and voice or multimedia[1]. The first generation systems use frequency modulation for speech and frequency division multiple access (FDMA) for analog cellular systems. The second-generation systems provide digital speech and data. The third-generation systems using 2GHz carrier frequency, called IMT-2000, are expected to provide multimedia service with 2Mbps data rate and same quality as fixed networks[2]. A general trend in multiple access scheme is moved from FDMA to time division multiple access (TDMA) or CDMA. Especially, applications of the DS-CDMA scheme to digital cellular mobile communication system have attracted much intention because of a great capacity than FDMA or TDMA. Also wideband wireless access based on DS-CDMA technology is under extensive

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development for IMT-2000 systems.

In DS-CDMA system, many users transmit messages simultaneously over the same communication channel, each using a specific spread-spectrum pseudo-noise (PN) code. At the receiver, an users message can be extracted from a PN code pre-assigned to each user. However the extraction of each users message is suffered by MUI from other simultaneously transmitted signal. If PN codes of each user are orthogonal, MUI will be reduced. However these codes cannot be exactly orthogonal due to multipath. Thus the capacity of DS-CDMA system is limited by MUI[3].

There has been a large amount of interest in the design of multi-user receiver for DS-CDMA system[4],[5],[8]. They focus on MUI cancellation and decoding to increase system performance and coverage and improve capacity of the system. However they consider MUI cancellation and decoding as an independent operation.

In this paper, we propose a scheme that MUI cancellation and decoding have an iterative operation using the soft estimated value and APP from decoder output. In the proposed scheme, more MUI cancellation derives better decoding performance at decoder, and more accurate decoding leads better MUI cancellation at matched filter.

The iterative decoding method has been introduced by Berrou *et al.* as a coding technique called Turbo codes[6]. In that paper, the soft output of MAP decoder is used as an extrinsic information at next iteration. Iterative decoding method of Turbo codes motivates the application of iterative MUI cancellation in this paper.

The MAP decoding method has been introduced by Bahl *et al.*[7] as an optimal decoding of linear codes for minimizing symbol error rate. The viterbi decoding algorithm minimizes the probability of word error and MAP decoding method minimizes the probability of symbol (or bit) error. In this paper, soft output of coded bits is required for iterative MUI cancellation. Thus we use the MAP decoding method.

Throughout this paper, scalar is expressed by lower case, vector is expressed by bold lower case, and matrix is expressed by bold upper case. The symbol $(\cdot)^T$ denotes matrix transposition operator. The subscript of variable means the time increment and the superscript of variable means the user index, except if stated otherwise.

This paper is organized as follows. In Section II, a system model is presented. In Section III and IV, interference cancellation and MAP decoding method are described. In Section V, iterative interference and decoding method is presented. Simulation results are presented in Section VI. Finally, conclusions are given in Section VII.

II. System Model

The uplink of a DS-CDMA communication is considered. Each user transmits the coded symbols over a slowly time varying and frequency non-selective fading channel with an additive white Gaussian noise (AWGN) with zero mean and variance $\sigma^2 = N_o/2$, where N_o is the single sided noise power spectrum density. Synchronous transmission is assumed. Block diagram of multi-user transmitter is shown in Figure 1.

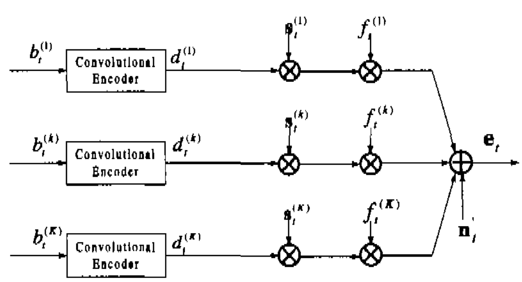

Fig. 1 DS-CDMA transmitter and channel model

Information bit $b_t^{(k)} \in \{0, +1\}$ is coded into $d_{t,c}^{(k)} \in \{+1, -1\}$ by $1/n$ convolutional encoder with specific structure, where $k \in \{1, 2, \cdots, K\}$ is the user index and $t \in \{0, 1, \cdots, L-1\}$ is the time index, and $c \in \{1, 2, \cdots, n-1\}$ is coded bit index at time t. The spreading code used by user k at symbol interval t consists of N chips and is denoted $s_{t,c}^{(k)} \in \{-1/\sqrt{N}, +1/\sqrt{N}\}^N$. The power of

N chips is normalized by 1. Assuming perfect channel estimation at the receiver, the received signal can be expressed as

$$\boldsymbol{e}_{t,c} = \boldsymbol{A}_{t,c}\boldsymbol{d}_{t,c} + \boldsymbol{n}'_{t,c} \tag{1}$$

where $\boldsymbol{d}_{t,c} = (d^{(1)}_{t,c}, d^{(2)}_{t,c}, \cdots, d^{(K)}_{t,c})^T$ is all user's c th coded data vector at time t, $\boldsymbol{A}_{t,c} = (\boldsymbol{s}^{(1)}_{t,c}f^{(1)}_{t,c}, \boldsymbol{s}^{(2)}_{t,c}f^{(2)}_{t,c}, \cdots, \boldsymbol{s}^{(K)}_{t,c}f^{(K)}_{t,c})$ is the bank of spreading codes multiplied by each fading variable $f^{(k)}_{t,c}$, and $\boldsymbol{n}_{t,c}'$ is noise vector.

Ⅲ. Interference Cancellation

At the receiver, the matched filter output at time $t$can be expressed as

$$\boldsymbol{y}_{t,c} = \boldsymbol{H}_{t,c} \cdot \boldsymbol{d}_{t,c} + \boldsymbol{n}_{t,c} \tag{2}$$

where $\boldsymbol{H}_{t,c} = \boldsymbol{A}^T_{t,c} \cdot \boldsymbol{A}_{t,c}$ is the correlation matrix of the spreading sequences and $\boldsymbol{n}_{t,c} = \boldsymbol{A}^T_{t,c} \cdot \boldsymbol{n}_{t,c}'$. These noise samples $\boldsymbol{n}_{t,c}$ are Gaussian distributed and have zero mean and variance $E\{\boldsymbol{n}_{t,c} \cdot \boldsymbol{n}^T_{t,c}\} = \boldsymbol{H}^T_{t,c}\sigma^2_n$.

For the purpose of describing the interference cancellation, the matched filter output $\boldsymbol{y}_{t,c}$ can be rewritten as a perturbed version of $\boldsymbol{d}_{t,c}$

$$\boldsymbol{y}_{t,c} = \boldsymbol{W}_{t,c}\boldsymbol{d}_{t,c} + \boldsymbol{M}_{t,c}\boldsymbol{d}_{t,c} + \boldsymbol{n}_{t,c} \tag{3}$$

where $\boldsymbol{W}_{t,c} = \mathrm{diag}(\boldsymbol{H}_{t,c})$ is the diagonal matrix of $\boldsymbol{H}_{t,c}$ and $\boldsymbol{M}_{t,c} = \boldsymbol{H}_{t,c} - \boldsymbol{W}_{t,c}$. $\boldsymbol{W}_{t,c}$ is the auto-correlation matrix ofthe spreading sequences and $\boldsymbol{M}_{t,c}$ is the crosscorrelation matrix of the spreading sequences which causes MUI.

Figure 2 shows the proposed iterative interference canceller. The decoder output $\hat{d}^{(k)}_{t,c}$ is respreaded by using each spreading sequence and summed. Summation for kth user is $\sum_{i=1(i\neq k)}^{K}(\hat{d}^{(i)}_{t,c}\boldsymbol{s}^{(i)}_{t,c}f^{(i)}_{t,c})$. When iteration is done, (3) can be expressed as

$$\boldsymbol{y}_{t,c} = \boldsymbol{W}_{t,c\ t,c} + \boldsymbol{M}_{t,c}\boldsymbol{d}_{t,c} - \boldsymbol{M}_{t,c}\hat{\boldsymbol{d}}_{t,c} + \boldsymbol{n}_{t,c}. \tag{4}$$

Let $\boldsymbol{\varepsilon}_{t,c} = \boldsymbol{d}_{t,c} - \hat{\boldsymbol{d}}_{t,c}$ be an error sequence. Then (4) is rewritten as

$$\boldsymbol{y}_{t,c} = \boldsymbol{W}_{t,c}\boldsymbol{d}_{t,c} + \boldsymbol{M}_{t,c}\boldsymbol{\varepsilon}_{t,c} + \boldsymbol{n}_{t,c}. \tag{5}$$

If the decoder output $\hat{d}^{(k)}_{t,c}$ is equal to original coded bit $d^{(k)}_{t,c}$, $\boldsymbol{M}_{t,c}\boldsymbol{\varepsilon}_{t,c} = \boldsymbol{0}$ and thus single user performance is achieved. Here $\boldsymbol{0}$ is null vector. Also we find the relationship between the variance reduction of interference canceller and the accuracy of decoder. Let σ^2_{IC} be the variance of the matched filter output. For sufficiently large number of users, the random variable $\boldsymbol{M}_{t,c}\boldsymbol{\varepsilon}_{t,c}$ is Gaussian distributed with zero mean and variance σ^2_{MU}. Assuming that the $\boldsymbol{n}_{t,c}$ and $\boldsymbol{M}_{t,c}\boldsymbol{\varepsilon}_{t,c}$ are statistically independent, the variance of the matched filter output can be expressed as

$$\sigma^2_{IC} = \sigma^2_{MU} + \sigma^2_n \tag{6}$$

where $\sigma^2_{MU} = E\{(\boldsymbol{M}_{t,c}\boldsymbol{\varepsilon}_{t,c})^2\}$ is given by

$$\sigma_{MU} \cong w\sigma^2_{CC}\frac{K-1}{N} \tag{7}$$

Here w is the received power for symbol under perfect power control and σ^2_{cc} is the average power per bit in the error sequence $\boldsymbol{\varepsilon}_{t,c}$ given by $\sigma^2_{CC} = \mathrm{variance}(d^{(k)}_{t,c} - \hat{d}^{(k)}_{t,c})$.

Finally (6) can be rewritten as

$$\sigma^2_{Ic} = w\sigma^2_{CC}\frac{K-1}{N} + \sigma^2_n \tag{8}$$

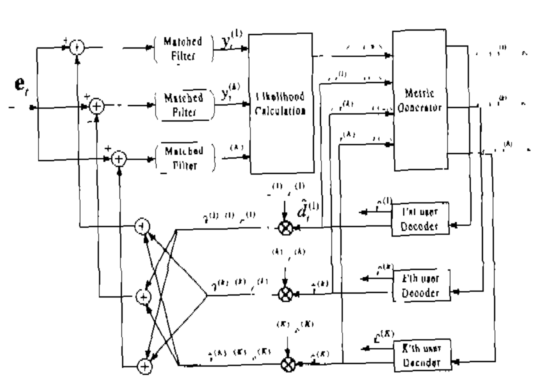

Fig. 2 Iterative MUI cancellation and decoding structure

Note that the attenuation of σ^2_{MU} through interference canceller is dependent on the load of system $(K-1)/N$.

Ⅳ. MAP Decoding Method

The MAP decoding method is to estimate the APP of channel symbol $p(d_{t,c}^{(k)}|y^{(k)})$ or the APP of information bit $p(b_t^{(k)}|y^{(k)})$ for received sequences $y^{(k)}$. The MAP decoding method typically takes as input the metric $p(y_{t,c}^{(k)}|d_{t,c}^{(k)})$ and output $p(d_{t,c}^{(k)}|y^{(k)})$ as the APP. In multi-user environment, the probability of $p(d_{t,c}^{(k)}=d)$ is dependent on the whole vector y because of multi-access interference (MAI). Thus we take as input the metric $p(y_t|d_{t,c}^{(k)})$ and output $p(d_{t,c}^{(k)}|y)$ as the APP.

The metric for single user decoder $p(y_{t,c}|d_{t,c}^{(k)})$ is generated from conditional probability $p(y_{t,c}|d_{t,c})$. In Figure 2, $p(y_{t,c}|d_{t,c})$ is the output of the likelihood calculation block and can be expressed as

$$p(y_{t,c}|d_{t,c})=p(y_{t,c}^{(1)},y_{t,c}^{(2)},\cdots,y_{t,c}^{(K)}|d_{t,c}^{(1)},d_{t,c}^{(2)},\cdots,d_{t,c}^{(K)})$$

$$=\frac{1}{\sqrt{2\pi}\sigma_{IC}}\exp\left\{\frac{-\sum_{i=1}^{K}(d_{t,c}^{(i)}-y_{t,c}^{(i)})^2}{2\sigma_{IC}^2}\right\}. \tag{9}$$

The metric for single user decoder $p(y_{t,c}|d_{t,c}^{(k)})$ is calculated by using joint probability as follows

$$p(y_{t,c},d_{t,c}^{(k)})=\sum_{\substack{d_{t,c}\\(d_{t,c}^{(k)}=d)}}p(y_{t,c}|d_{t,c})p(d_{t,c}). \tag{10}$$

Using (9), the metric generator block calculates

$$p(y_{t,c}|d_{t,c}^{(k)})=\frac{p(y_{t,c},d_{t,c}^{(k)})}{p(d_{t,c}^{(k)})}$$

$$=\sum_{\substack{d_{t,c}\\(d_{t,c}^{(k)}=d)}}p(y_{t,c}|d_{t,c})\prod_{i=1(i\neq k)}^{K}p(d_{t,c}^{(i)}) \tag{11}$$

where $\prod_{i=1(i\neq k)}^{K}p(d_{t,c}^{(k)})=p(d_{t,c}^{(1)},d_{t,c}^{(2)},\cdots,d_{t,c}^{(K)})$

From now the decoder block of the MAP decoding method is briefly described. In the single user decoder, the APP is estimated by traversing the decoding trellis in the forward and backward directions. The state of the coded trellis at time t is denoted by S_t. The objective of the decoder is to estimate the APP given by

$$p(b_t^{(k)}|y)=p(S_{t-1}=m',S_t=m|y) \tag{12}$$

Let us define the probability functions as

$$\alpha_t(m)=p(S_t=m,y_1^t) \tag{13}$$

$$\beta_t(m)=p(y_{t+1}^L|S_t=m) \tag{14}$$

$$\gamma_t(m'm)=p(S_t=m,y_t|S_{t-1}=m') \tag{15}$$

where $y=(y_1^{t-1},y_t,y_{t+1}^L)$ denotes total received bits for K user and each L coded bit. Then (12) can be expressed as

$$p(S_{t-1}=m',S_t=m|y)=\frac{\alpha_{t-1}(m')\beta_t(m)\gamma_t(m',m)}{p(y)}. \tag{16}$$

And $\alpha_t(m)$ and $\beta_t(m)$ are calculated recursively as follows:

$$\alpha_t(m)=\sum_{m'}\alpha_{t-1}(m')\gamma_t(m',m) \tag{17}$$

and

$$\beta_t(m)=\sum_{m'}\beta_{t+1}(m')\gamma_{t+1}(m,m'). \tag{18}$$

Here the boundary conditions are given by

$$\alpha_0(0)=1,\alpha_0(m)=0,(m\neq 0) \tag{19}$$

and

$$\beta_L(0)=1,\beta_L(m)=0,(m\neq 0). \tag{20}$$

Also (15) can be rewritten as

$$\gamma_t(m',m)=\sum_{b=b_t}\{p(y_t|b)p(b|S_t=m,S_{t-1}=m')\cdot p(S_t=m|S_{t-1}=m')\} \tag{21}$$

where $p(S_t=m|S_{t-1}=m')$ is the state transition probability.

The distribution $p(y_t|b)$ is the memoryless channel symbol distribution given the transmitted information bit. When the transition from state m' to state m exits for a known value of b and there is only one transition, $p(b|S_t=m, s_{t-1}=m')=1$, and (21) can be simplified to

$$\gamma_t(m',m)=p(y_t|b)p(S_t=m|S_{t-1}==m'). \qquad (22)$$

This can also be rewritten in terms of the coded bits as

$$\gamma_t(m',m)=p(S_t=m|S_{t-1}=m')\prod_{c=0}^{n-1}p(y_{t,c}|d_{t,c}^{(k)}). \qquad (23)$$

If there is no a priori information, then $p(S_t=m|S_{t-1}=m')=1/2$ if the transition exist.

The APP of information bit is given by

$$p(b_t^{(k)}=b|y)=\sum_{\substack{m',m\\(b_t^{(k)}=b)}}\alpha_{t-1}(m')\gamma_t(m',m)\beta_t(m). \qquad (24)$$

The APP of coded bit is given by

$$p(d_t^{(k)}=d|y)=\sum_{\substack{m',m\\(d_{t,c}^{(k)}=d)}}\alpha_{t-1}(m')\gamma_t(m',m)\beta_t(m). \qquad (25)$$

The APP of information bit is used for decision of information bit and the APP of coded bit is used for decision of coded bit and iteration. The step is summarized as:

1) $\alpha_o(m)$ and $\beta_L(m)$ are initialized according to (19), (20).

2) As soon as y_t is received, the decoder compute $\gamma_t(m',m)$ using (23) and $\alpha_t(m)$ using (17). The value $\gamma_t(m',m)$ and $\alpha_t(m)$ is stored for all t and m.

3) After the all sequences y have been received, compute $\beta_t(m)$ using (18).

4) Estimate the APP using (24), (25). And decide information bit or symbol with greater probability.

V. Iterative MUI Cancellation and Decoding

In Sections III and IV, the interference cancellation and MAP decoding method are described. In this section, we describe how the two operations are jointed as shown in Figure 2.

When the iteration is used, the metric generator block generates the newly updated probability $p(y_{t,c}|d_{t,c}^{(k)}=d)$ using probability $p(y_{t,c}|d_{t,c})$ of the likelihood calculator block output and the APP $p(d_{t,c}^{(k)}=d|y_{t,c})$ of MAP decoder output. These operations give MAP decoder to more accurate decoding and estimation of the APP. Also the accurate estimation of the channel symbol $\hat{d}_{t,c}^{(k)}$ gives more interference cancellation at the interference canceller. After decoding, the estimation of the channel symbol $\hat{d}_{t,c}^{(k)}$ for respreading is calculated as

$$\begin{aligned}\hat{d}_{t,c}^{(k)}&=E\{p(d_{t,c}^{(k)}|y_{t,c})\}\\&=p(d_{t,c}^{(k)}=+1|y_{t,c})-p(d_{t,c}^{(k)}=-1|y_{t,c}).\end{aligned} \qquad (26)$$

This conditional expectation is called a soft estimate which the interference canceller uses to cancel the MUI. When a desired number of iterations is completed, the information bit decision may be computed by hard decision.

VI. Simulation Results

In this Section, we show the performance of the proposed scheme. The forward error control code employed is rate 1/2, 4 state non-recursive convolutional code with generator polynomials $g_1(D)=1+D+D^2$ and $g_2(D)=1+D^2$. The number of users is 5 and spreading code length is 7. Figure 3 shows the performance of the proposed system on AWGN channel. In this figure, "Iter=1" and "Iter=2" denote the iteration number and "No Iter" denotes that iteration is not used. Also "Single user" denotes single user performance. "(left)" denotes a scheme to use only left iteration. "(right)" denotes a scheme to use only right iteration. The left iteration means that the system uses the conventional MAP decoding and the iterative interference canceller. The right iteration proposed in [8] means that the system only uses iterative decoding, not using iterative interference canceller.

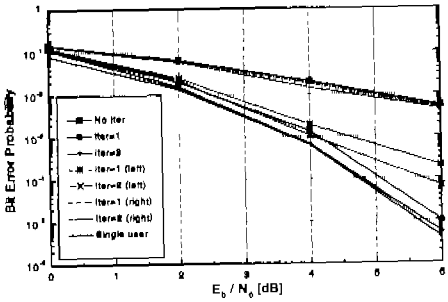

Fig. 3 Bit error probability over AWGN channel

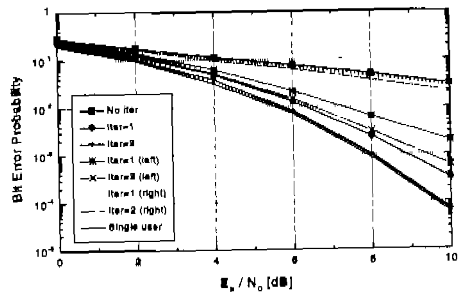

Fig. 4 Bit error probability over Rayleigh fading channel

To investigate the effects of the left and right iterations, the performances of all three schemes are plotted simultaneously. One can see that the right iteration scheme achieves minor improvement of BER performance than the left iteration scheme. The performance of the proposed scheme is better than that of only left iteration.

Figure. 4 and 5 show the BER performance of the proposed system on Rayleigh and Rician channels, respectively. In Figure 4, the mean squared value of Rayleigh distribution is 1/2. Also in Figure 5, the Rician factor is 3. The major improvement of BER performance in the fading channel is achieved when the first iteration is completed as like in the AWGN channel.

It is noticeable that the performance of the proposed scheme using left iteration and right iteration simultaneously reaches to the single user performance only when second iteration is completed.

Ⅶ. Conclusions

In this paper, we propose an iterative interference cancellation and decoding scheme for DS-CDMA system. The proposed scheme is reached to single user performance only when second iteration is completed. Also we investigate the effects of left and right iterations. The major improvement of the BER performance is achieved by iterative MUI canceller using the soft estimated value of MAP decoder output. Also the minor improvement of the BER performance is achieved by iterative MAP decoder using APP.

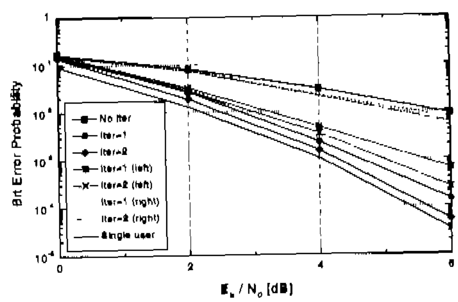

Fig. 5 Bit error probability over Rician fading channel

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