논문 2007-44TC-1-16

TD-SCDMA 이동국용 FIR-Rake 수신기 설계

(Design of FIR-Rake Receiver for TD-SCDMA Mobile Stations)

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요 약

본 논문에서는 채널 파라미터 추정을 위해 TD-SCDMA 기지국에 의해 방송되는 파일럿 신호를 이용하기 위한 FIR-Rake 수신기를 개발한다. FIR-Rake 수신기는 이동국에 대한 다중경로 간섭을 완화할 수 있다. 주파수 영역과 이산 영역에서의 해석, 그리고 컴퓨터 시뮬레이션은 제안 FIR-Rake 수신기가 전형적인 다중경로 페이딩 전파 환경에서 더 우수한 성능을 나타내는 것을 확인하였다.

Abstract

This paper developed a FIR-Rake receiver which makes use of a pilot signal broadcast by the TD-SCDMA (division-synchronous code division multiple access) base station to estimate the channel parameters. The FIR-Rake receiver can reduce multipath interference for the mobile stations. The analysis in frequency and discrete domain and computer simulations confirm that the proposed FIR-Rake receiver achieves a better performance under typical multipath fading propagation conditions.

Keywords : TD-SCDMA, FIR-Rake receiver, channel estimation, communication circuits.

Ⅰ. Introduction

TD-SCDMA standard seems to be a promising approach for implementing cellular communication services[1,2]. Though TD-SCDMA has absorbed many new techniques recent proposed, it did not consider Pre-Rake and the Rake techniques for its base stations and mobile stations[3-8].

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※ This work was supported by the International Cooperation Research Program of the Ministry of Science & Technology, KOTEF, the IT Foreign Specialist Inviting Program (ITFSIP) supervised by IIFA under ITRC supervised by IITA, and 2nd stage BK21, Korea.
접수일자: 2005년8월16일, 수정완료일: 2007년1월12일

To improve the communication quality of the downlinks of TD-SCDMA, this paper extends the continuous-time algorithm of Ref [5] for simple case (one user and broadcast pilot channel) to our complicated and general case (discrete time, multiusers and adaptive channel estimation from the burst frame structure with pilot slot [1,2], shown as

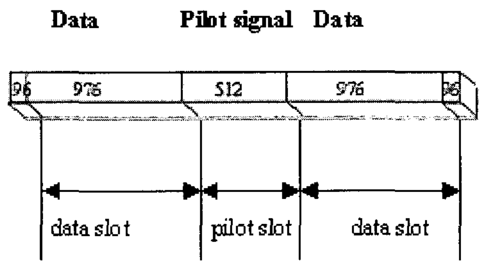

그림 1. TD-SCDMA 버스트 시컨스 구조
Fig. 1. TD-SCDMA burst sequence structure.

Fig. 1). The estimation model in FIR-Rake can provide channel parameter to adjust the FIR model in FIR-Rake receiver for the TD-SCDMA mobile terminals. Time domain and frequency domain analysis and numerical simulation results are used to demonstrate the performance of the proposed FIR-RAKE receiver in terms of bit error probability under multipath fading propagation conditions.

II. TD-SCDMA system and channel model

A frequency-selective slowly fading transmission channel has been assumed as the Ref [5-8]. The low-pass impulse response of the channel for the kth user is given by

$$h(k,t)=\sum_{l=0}^{L-1}\alpha(k,l)\delta[t-\tau(k,l)],\ \ k=0,\ldots,L \tag{1}$$

where L is the number of resolvable paths, where $\alpha(k,l)$ and $\tau(k,l)$ is the delay of the lth path of the kth user, respectively,

The estimation of the parameters $\{\alpha(k,l),\tau(k,l)\}$ is achieved by means of a pilot signal at burst dada frame, which is broadcasted by TD-SCDMA base station to all the mobile users in the cell.

For the sake of simplicity, the TD-SCDMA system under consideration is assumed as formed by a single isolated circular cell of radius R with a centrally located base-station. In particular, the focus here is on downlink communications. It is assumed that the base station communicates with the K spatially dispersed mobile users by a TD-SCDMA scheme. All the K transmitted signals from the TD-SCDMA base-station have experienced identical fading when received by a particular mobile user. These users are assumed to be uniformly distributed throughout the cell area.

We have assumed parameter L(i.e., the number of resolvable paths in our channel model) equal to four and considered negligible the inter-symbol interference introduced by the downlink channel, i.e.,

$$0\le\tau(k,l)<0.2T_b \tag{2}$$

for $1\le l\le L$, with T_b, the user data bit duration.

III. FIR-RAKE algorithm

QPSK modulation scheme is used in the transmission of TD-SCDMA communication system.

The baseband representation of the in-phase and quadrature components of the transmitted signal from the base station to the kth mobile user in its cell is

$$s(k,t)=\begin{cases}b(k,t)c_I(k,t)+jd(k,t)c_Q(k,t), \text{dada slot}\\ c_E(t), \text{pilot slot}\end{cases} \tag{3}$$

where

$$b(k,t)=\sum_{m=0}^{\infty}b(k,m)q(t-mT_b) \tag{4}$$

$$d(k,t)=\sum_{m=0}^{\infty}d(k,m)q(t-mT_b) \tag{5}$$

$$q(t)=\begin{cases}1 & 0\le t<T_b\\ 0 & otherwise\end{cases}$$

$\{b(k,m)\}$ is in-phase data sequence of kth mobile user, with $b(k,m)$ equal to ±1 with equal probability; $\{d(k,m)\}$ is quadrature data sequence, with $d(k,m)$ equal to ±1 with equal probability, and T_b is the user data bit duration.

The following pseudonoise (PN) code sequences in (1) have following forms

$$c_I(k,t)=\sum_{n=0}^{N-1}c_I(k,n)p(t-nT_c) \tag{6}$$

$$c_Q(k,t)=\sum_{n=0}^{N-1}c_Q(k,n)p(t-nT_c) \tag{7}$$

where $\{c_{k,I}(n)\}$ is PN sequence associated with the in-phase component, with $c_{k,I}(n)$ equal to ±1 (binary chip); $\{c_{k,Q}(n)\}$ is PN sequence associated with the quadrature component, with $c_{k,Q}(n)$ equal to ±1 (binary chip); T_c is the chip duration, $T_b=QT_c$, Q is spreading gain; n is the number of chips in the PN sequence (or chips/bit); N is the length of PN sequence, and

$$p(t)=\begin{cases}1 & 0\le t<T_c \\ 0 & otherwise\end{cases} \tag{8}$$

Additionally, a PN signal (pilot signal) from a TD-SCDMA base station is added to the downlink transmitted signal in order to perform the channel parameter estimation.

The pilot signal in (3) is defined as

$$c_E(t)=\sqrt{P}\sum_{n=0}^{N-1}c_E(n)p(t-nT_c) \tag{9}$$

where P denotes the power imbalance of the pilot signal with respect to the information bearing signals and $c_E(n)(=\pm 1)$ are the binary chips forming the PN sequence.

We assumed that the base-band signal of the system is sampled by two times of the chip frequency:

$$f_s=2f_c=1/T_s=2/T_c \tag{10}$$

then $t=nT_s$. We have the discrete channel pulse response

$$h(k,nT_s)=\sum_{l=0}^{L-1}\alpha(k,l)\delta[nT_s-\tau(k,l)],\ \ k=0,...,K \qquad)$$

and the discrete transmitted signal

$$s(k,nT_s)=\begin{cases}b(k,nT_s)c_I(k,nT_s)+jd(k,nT_s)c_Q(k,nT_s),\text{ data slot} \\ c_E(nT_s),\text{ pilot slot}\end{cases} \tag{12}$$

where

$$b(k,nT_s)=\sum_{m=0}^{\infty}b(k,m)q(nT_s-mT_b) \tag{13}$$

$$d(k,nT_s)=\sum_{m=0}^{\infty}d(k,m)q(nT_s-mT_b) \tag{14}$$

$$c_I(k,nT_s)=\sum_{m=0}^{N-1}c_I(k,m)p(nT_s-mT_c) \tag{15}$$

$$c_Q(k,nT_s)=\sum_{m=0}^{N-1}c_Q(k,m)p(nT_s-mT_c) \tag{16}$$

The discrete pilot signal is

$$c_E(nT_s)=\sqrt{P}\sum_{m=0}^{N-1}c_E(n)p(nT_s-mT_c) \tag{17}$$

From (12), if K users are active in a TD-SCDMA cell, it is straightforward to see that the received signal (baseband) for the kth mobile is

$$r(nT_s)=\sum_{k=1}^{L}r(k,nT_s)+v(nT_s) \tag{18}$$

where $v(n)$ is an AWGN term with zero mean and two-sided power spectral density $N_0/2$, and

$$r(k,nT_s)=\begin{cases}\sum_{l=1}^{L}\alpha(k,l)s[k,nT_s-\tau(k,l)],\text{ data slot} \\ \sum_{l=1}^{L}\alpha(k,l)c_E[nT_s-\tau(k,l)],\text{ pilot slot}\end{cases} \tag{19}$$

$r(k,nT_s)$ is given by the sum of the output of the channel (11) having a randomly time-varying impulse response. The multipath effects are represented as a sequence of replicas of the transmitted signal (12), each one characterized by a particular delay and attenuation.

Notice: Though the transmitted sequence $s(k,t)$ is a binary one, however, it becomes multilevel one-$r(k,nT_s)$, when it passes the multipath channel. Here, at the receiver, we use soft-demodulation. In this paper, we assume that the sequence at the input of FIR-Rake is a multilevel and discrete sequence. At the output of FIR-Rake receiver there are two binary sequences.

Classical Rake to collect the time diversity is to adjust the marching filters to the lager fingers, then combine the marching results, such as

$$c_I(k,nT_s)=\sum_{l=0}^{L-1}r(nT_s)c_I[k,nT_s-\tau(k,l)] \tag{20}$$

and

$$c_Q(k,nT_s)=\sum_{l=0}^{L-1}r(nT_s)c_Q[k,nT_s-\tau(k,l)] \tag{21}$$

For each finger, the Rake despreads $r(nT_s)$ with the Walsh sequence $\{c_I(k,nT_s)\}$ and $c_Q(k,nT_s)$.

If not use data buffers, from (20) and (21), there will be L fingers, and we need 2L marched filters. It brings a great difficulty for the circuits' implementation of Rake mobile receiver.

Whether there is a simple approach to solve the complexity of Rake implementation? To solve the problem, we design a FIR filter to process the signal $r(nT_s)$, and the filter can collect the energy of the main fingers.

We notice that (20) and (21) can be written another form:

$$c_I(k,nT_s)=\sum_{l=0}^{L-1} r[nT_s-\tau(k,l)]c_I(k,nT_s) \tag{22}$$

and

$$c_Q(k,nT_s)=\sum_{l=0}^{L-1} r[nT_s-\tau(k,l)]c_Q(k,nT_s) \tag{23}$$

In this case, the number of marched filters reduced into 2. We only need move the received sequence to deferent main delays. This function can be realized by a FIR filter. The FIR filter is derived from $\alpha(k,l)$, the attenuation is introduced by the l-path, and $\tau(k,l)$, time delay is introduced by the l-path for $1\le l\le L$, we call it as FIR-Rake, shown in Fig. 2, and with the following pulse response,

$$h(k,nT_s)=\sum_{l=0}^{L-1}\alpha(k,L-l)\delta[nT_s-\tau(k,l)],\ \ k=0,\dots,K \tag{24}$$

When the received sequence $r(nT_s)$ is processed by the filter, from the output the filter we get

$$y(k,nT_s)=\sum_{i=1}^{L-1}\alpha(k,L-l)r(nT_s-\tau(k,l)) \tag{25}$$

If we use two matched filters with PN sequences $\{c_I(k,nT_s)\}$ and $c_Q(k,nT_s)$ to multiply the output the filter, we get

$$c_{FI}(k,nT_s)=\sum_{l=0}^{L-1}\alpha(k,L-l)r[nT_s-\tau(k,l)]c_I(k,nT_s) \tag{26}$$

and

$$c_{FQ}(k,nT_s)=\sum_{l=0}^{L-1}\alpha(k,L-l)r[nT_s-\tau(k,l)]c_Q(k,nT_s) \tag{27}$$

Compare (26) and (27) with (22) and (23), we find that they have similar forms except the filter coefficients $\{\alpha(k,L-l)\}$. In following simulation we will see that the correct filter coefficients $\{\alpha(k,L-l)\}$ can enhance the effect of main fingers, though we need channel estimation model to acquire the parameters. Compare with the classical Rake using scan correlation to search main figures, the channel estimation for the filter coefficients $\{\alpha(k,L-l)\}$ is not complicated. The (26) and (27) are a FIR like filtering of the received baseband signal $r(nT_s)$ in (18).

Now, we can despread the data sequences $\{b(k,m)\}$, the in-phase data sequence and $\{d(k,m)\}$, the quadrature data sequence of kth mobile user ,

$$b(k,m)=\begin{cases}1, if\ c_{FI}(k,nT_s)>0\\ 0, else\end{cases} \tag{28}$$

and

$$d(k,m)=\begin{cases}1, if\ c_{FQ}(k,nT_s)>0\\ 0, else\end{cases} \tag{29}$$

for $(m-1)T_b\le nT_s<(m-1)T_b$.

Different from the classical Rake, we need not use 2L matching filters with different delays to collect the time diversity signals. We only use two matched filter to detect the output signal $y(k,n)$ of the FIR-Rake, which is shown in Fig. 2.

In Fig. 2, the model of C-Est is a channel estimator, which adjusts the coefficients of FIR filter according to the received pilot sequence, and the algorithm of the C-Est model will be given in next section. In this section, we only consider that the FIR filter has obtained the parameters of multipath channel from pilot sequence.

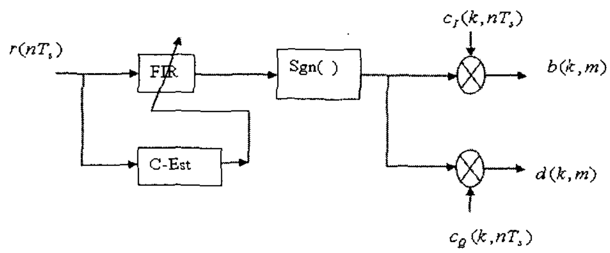

그림 2. FIR-Rake 수신기의 회로
Fig. 2. The circuits of FIR-Rake receiver.

Ⅳ. The Acquirement of the FIR filter Coefficient

Before we establish our estimation algorithm for the FIR filter coefficients $\{\alpha(k,L-l)\}$, we need make some simplification for channel model and FIR filter. We try to express $\tau(k,l)$ in a simple form. For $1 \le l \le L$, since

$$0 \le \tau(k,l) < 0.2T_b = 0.2QT_c = 0.1QT_s \tag{30}$$

and $\tau(k,l) \in \{0,1,\ldots,0.1QN\}T_s$, N is the length of PN sequence, let $M = 0.1QN$, we can extend channel model and FIR filter into a general form,

$$h(k,nT_s) = \sum_{m=0}^{M-1} \alpha(k,m)\delta(nT_s - mT_s),\ \ k = 0,\ldots,K \tag{31}$$

and

$$h(k,nT_s) = \sum_{m=0}^{M-1} \alpha(k,M-m)\delta(nT_s - mT_s),\ \ k = 0,\ldots,K \tag{32}$$

Consider first the pilot slot, in which the system is only corrupted by noise. The pilot digital signal

$\mathbf{c}_E = [c_E(0)\, c_E(1)\, \ldots, c_E(M-1)]$,
whose components are given by Eq. (9), is transmitted over a fading multipath channel $\mathbf{h}(k)$ of the kth user, after which the signal has memory of N symbols. Thermal noise is generated at the receiver and it is modeled by additive white Gaussian noise $\mathbf{v}(k)$, which is sampled at the symbol rate.

The demodulation problem here is to detect the transmitted bits $\mathbf{c}_E$ from the received signal

$\mathbf{y}(k) = [y(k,0)\, y(k,1)\, \ldots y(M-1)]$.

Besides the received signal the detector needs also the channel estimates, which are provided by a specific channel estimator device.

The received signal $\mathbf{y}(k)$ can be expressed as follows

$$\mathbf{y}(k) = \mathbf{C}\mathbf{h}(k) + \mathbf{v}(k) \tag{33}$$

where the channel impulse response $\mathbf{h}(k)$ of the wanted signal is expressed as

$$\mathbf{h}(k) = [\alpha(k,0)\ \alpha(k,1)\ \alpha(k,2)\ \ldots\ \alpha(k,M\text{-}1)]^T \tag{34}$$

Within each transmission burst the transmitter sends a unique training sequence $\mathbf{c}_E$ which is the length of M and denoted by

$$\mathbf{c}_E = [c_E(0)\, c_E(1)\, \ldots, c_E(M-1)] \tag{35}$$

having bipolar elements $m_i \in \{\text{-}1,1\}$ Finally to achieve Eq. (9) the circulate training sequence matrix $\mathbf{C}$ is formed as

$$\mathbf{C} = \begin{bmatrix} c_E(0) & 0 & \cdots & 0 & 0 \\ c_E(1) & c_E(0) & \cdots & 0 & 0 \\ \vdots & \vdots & & & \\ c_E(M-1) & c_E(M-2) & \cdots & c_E(1) & c_E(0) \end{bmatrix} \tag{36}$$

The LS channel estimates are found by minimising the following squared error quantity

$$\hat{\mathbf{h}}(k) = \arg\min_{\mathbf{h}(k)} \|\mathbf{y}(k) - \mathbf{C}\mathbf{h}(k)\|^2$$

Assuming white Gaussian noise the solution is given by [9]

$$\hat{\mathbf{h}}_{LS}(s) = (\mathbf{C}^H\mathbf{C})^{-1}\mathbf{C}^H\mathbf{y}(k) \tag{37}$$

where $()^H$ and $()^{-1}$ denote the Hermitian and inverse matrices, respectively. The given solution (37) is also the best linear unbiased estimate (BLUE) for the channel coefficients.

The given solution is further simplified to

$$\hat{\mathbf{h}}(k) = \mathbf{C}^H\mathbf{y}(k) \tag{38}$$

Provided that the periodic autocorrelation function (ACF) of the training sequence is ideal with the small delays from 1 to M, because the correlation matrix $\mathbf{C}^H$ becomes diagonal. This holds for TD-SCDMA training sequences, whenever reference length 512 is chosen. The estimations given by Eq. (38) are simply scaled correlations between the received signal and training sequence.

Ⅴ. FIR-RAKE time-domain and frequency domain analysis

In Section 4, we show that in discrete time domain, the FIR-rake can only use one matched filter to extract the data flow from an in-phase or quadrature channel, based on Eq. (26) and Eq. (27). When combine the multipath channel and FIR filter, they can be regarded as a serial system, with discrete time pulse response,

$$g(k,nT_s)=\sum_{l=0}^{L-1}\alpha(k,L-l)\sum_{m=0}^{L-1}\alpha(k,m)\delta[nT_s-\tau(k,m)-\tau(k,l)]$$
$$k=0,\ldots,K \tag{39}$$

Take z-transform to $g(k,nT_s)$, we have

$$G(k,z)=\sum_{l=0}^{L-1}\alpha(k,L-l)\sum_{m=0}^{L-1}\alpha(k,m)z^{-\tau(k,m)-\tau(k,l)}$$
$$k=0,\ldots,K \tag{40}$$

and let $z=\exp(jnT_s\omega)$,we get the frequency response of the combined system

$$G(k,\mathbf{j}\omega)=\sum_{l=0}^{L-1}\alpha(k,L-l)\sum_{m=0}^{L-1}\alpha(k,m)\exp(-\mathrm{j}\omega T_s(\tau(k,m)-\tau(k,l)))$$
$$k=0,\ldots,K \tag{41}$$

Based on the Eq. (39) - Eq. (41), we can do the frequency analysis for the system combined multipath channel and FIR filter, further to explain the time characteristics of the FIR-Rake.

Consider a multipath channel for pilot slot with the parameter,

h=[0.6 0 0.4 0 0 0.3 0 0] (42)

According to Eq.(24), we should have the FIR filter with coefficients

hf=[0 0 0.3 0 0 0.4 0 0.6] (43)

We see the channel pulse response and pulse co-response of channel and FIR filter, shown in Fig. 3. In discrete time domain,

The FIR filter changes the channel pulse response

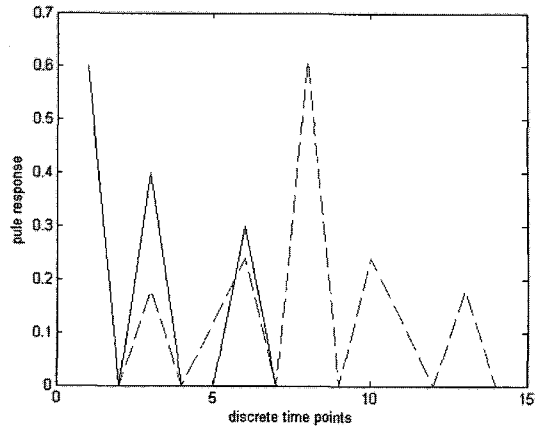

그림 3. 채널과 FIR 필터의 채널 펄스응답(직선)과 여응답(점선)

Fig. 3. The channel pulse response (real line) and pulse co-response (dotted line) of channel and FIR filter.

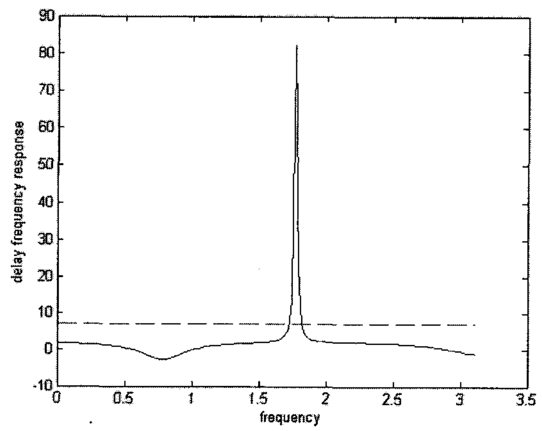

그림 4. 채널과 FIR 필터의 채널 시간지연 주파수 응답(직선)과 시간지연 주파수 여응답(점선)

Fig. 4. The channel time delay frequency response (real line) and time delay frequency co-response (dotted line) of channel and FIR filter.

from unsymmetrical into symmetric one, though the symmetric point is not at original. It is the symmetry of co-response of channel and FIR filter, to let nonlinear time delay characteristics of the multipath channel be changed into flat one, shown in Fig.4.

Under the linear phase or flat time delay frequency co-response of channel and FIR filter, we can use two matched filters to despread the signals in in-phase channel and the quadrature channel.

Now, we send a sequence

s=[1 1 1 1 -1 -1 -1 -1] (44)

into the multipath channel h given by Eq. (42), to see the output of FIR filter hf given by Eq. (43).

From the output of the channel, we get the distorted sequence

$$y=[0.6\ 0.6\ 1.0\ 1.0\ -0.2\ 0.1\ -0.7\ -0.7] \qquad (45)$$

which is shown in Fig. 7. If use hard justification, sgn(y(k)), k=1,···8, we will get

$$s'=[1\ 1\ 1\ 1\ -1\ 1\ -1\ -1] \qquad (46)$$

It is obvious that that there is an error on the 6th bit's position. The error is caused by the different multipath time delays of the channel. Whether combining FIR filter can correct the error? The output of FIR filter is the sequence

$$yf=[0.79\ 0.55\ 0.31\ 0.61\ -0.61\ -0.31\ -0.55\ -0.79] \qquad (47)$$

if use hard justification for Eq. (47), i.e. sgn(yf(k)), k=1,···8, we will get s'=[1 1 1 1 -1 -1 -1 -1]. The result is agreed with the transmitted sequence s which is given by Eq. (44).

Notice: In the practical application, the length of FIR filter is selected to be same as that of pilot sequence. Assume the length is M, then the length of the output FIR filter will be 2M-1. To keep the same length of the transmitted sequence, only the sequence located from the bit position from M to 2M-1.

To verify the LS channel estimation of Eq. (37) given in Section 4 to be valid, we use the output of the channel, y is given by Eq. (42), and training sequence ce=s, and s is given by Eq. (41), according to Eq. (36), we have

$$\mathbf{C}=\begin{bmatrix} 1 & 0 & \cdots & 0 & 0 \\ 1 & 1 & \cdots & 0 & 0 \\ \vdots & \vdots & & & \\ -1 & -1 & \cdots & 1 & 1 \end{bmatrix}$$

Substitute the matrix C and the sequence y into Eq. (37), we get the channel parameters, h' =[0.6 0 0.4 0 0 0.3 0 0], which is agreed with the h value in Eq. (37)

VI. The system simulation

Now, we apply the results in above sections to the TD-SCDMA case. The parameters $\tau(k,l)$, for $1 \le l \le L$, have been assumed fixed. Conversely, parameters $\alpha(k,l)$ for $1 \le l \le L$, has been considered as statistically independent and identically distributed random variables. The random variables $\alpha(k,l)$, $1 \le l \le L$, have been characterized statistically by independent Releigh distributions.

Numerical results determined by means of computer simulations are presented here. Each in-phase, quadrature data bit and each pilot signal bit is spread by a specific sequence from a set of 127 Gold codes operating at a rate of 200Mchip/s (all the four paths considered in our channel systems under the situation of 20 users and 4 paths model may be resolved). The sliding correlations between the received signal and the local generated version of the pilot signal are performed over a time interval 635 ns long. For each trial, the codes assigned to the users and to the pilot signal are randomly chosen from the set of available codes.

We have noted that BER as a function of parameter SINR, is influenced by the number of tracked replicas. In particular, we have found that the best choice is to track the first four most powerful replicas. This result is due to the fact that the

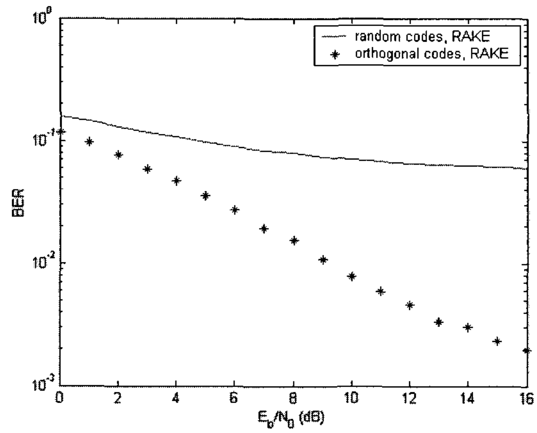

그림 5. Rake에 대한 비트 오율 대 Eb/N0

Fig. 5. The bit error rate versus the Eb/N0 for the rake.

multipath interference is relatively too strong in the two weaker replicas. Hence, in deriving the numerical results shown in the following figures we have always assumed L equal to 4. The curves in Fig. 5 depict the performance of RAKE techniques under the situation of 20 users and 4 paths. From this figure, we can notice that when orthogonal codes are used, the performance of TD-SCDMA mobile receivers can be improved.

Ⅶ. Conclusions

This paper has investigated the performance of FIR-Rake for TD-SCDMA user end. The FIR-Rake includes a channel estimator, which extract parameters of the multipath from the downlink pilot signal in the burst frame of TD-SCDMA system. The numerical results of frequency and discrete time domain have been used to demonstrate the performance of the proposed FIR-RAKE receiver in terms of bit error probability under multipath fading propagation conditions. The circuits of FIR-Rake are also suitable to UTRA-TDD WCDMA systems.

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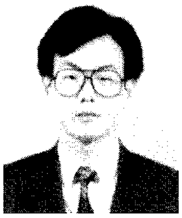
저 자 소 개



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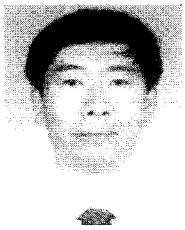

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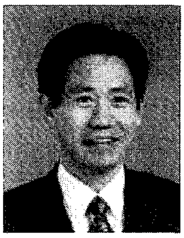

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