MAI-Suppression with Fractional T-equalizer for CDMA

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Abstract

Wideband CDMA systems with orthogonal spreading codes suffer severely due to the loss of orthogonality by multipath propagation. This yields Multiple Access Interference (MAI) which gravely reduces the performance of classical systems with Rake-receivers.

In our approach we attempt to restore orthogonality again by using a fractional T-equalizer. The results are compared with a system using the tapped delay line equalizer that works in chip duration and the classical solution for wideband CDMA systems, the Rake-receiver.

1. Introduction

UMTS is the 3rd generation mobile cellular communication system of the (near) future. Upon others it defines a FDD-Wide-Band-CDMA scheme. In numerous simulation environments the simulations are done in the chiprate domain. Due to this, impulse shaping and oversampling effects can be neglected. In the literature there are a lot of schemes which try to reestablish orthogonality using the Wiener approach [1] and many of them use some iterative algorithm like RLS and LMS to adapt its coefficients to a time variant channel [2]. Classical derivations of the Zero-Forcing Equalizer show that the normal tapped delay line equalizer (T-equalizer) can not fully equalize the channel. This can be achieved by using a fractional T-equalizer. The Main task of this paper is to show the feasibility of this approach. As stated before most simulation systems are working in Chip domain, therefore the Tx and Rx impulse shaping filter are neglected. Since our approach involves a fractional T-equalizer we can not do this anymore. To allow comparisons all simulations have to be performed in oversampling domain.
The receiver scheme we used as reference within our simulation environment is depicted in fig. 1. In our case the spread signal $s(j)$ is oversampled by the factor $w = 2$ and then put trough the impulse shaping filter $g_{TX}(k)$. In UMTS $g_{TX}(k)$ is a root raised cosine filter with $\alpha = 0.22$. The signal is transmitted over a multipath channel $h(k)$. The classical receiver structure for such a system is the coherent Rake-receiver.

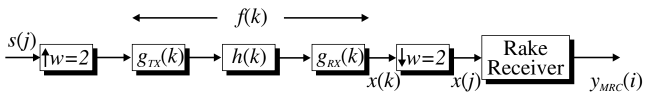

Figure 1: Used simulation scheme

After impulse shaping with another raised root cosine filter at the receiver

$$g_{RX}(k) = g_{TX}^*(T_0 - k) = g_{TX}(k) \tag{1}$$

and downsampling by factor w the received signal is processed by a Rake-receiver as shown in fig. 2. The signal $y_{ID,l}(i)$ for a specific Rake-finger l which can be detected after the integrate and dump operation for the i-th symbol is represented as[1]

$$y_{ID,l}(i) = \frac{1}{SF} \sum_{j=i\cdot SF}^{(i+1)\cdot SF-1} x_l(j) \cdot \frac{1}{\sqrt{2}} \cdot c_{scr}^*(j) \cdot c_{ch}(j) \tag{2}$$

where $c_{ch}(j)$ is the real valued orthogonal channelization code also known as OVSF-code with spreading factor (SF) and c_{scr} represents the complex scrambling code. The received signal after maximum ratio combining for a L_R-finger Rake-receiver is denoted as

$$y_{MRC}(i) = \sum_{l=0}^{L_R} y_{ID,l}(i) \cdot h_l^*(i) \tag{3}$$

where $h_l(i)$ is the channel coefficient for the l-th finger.

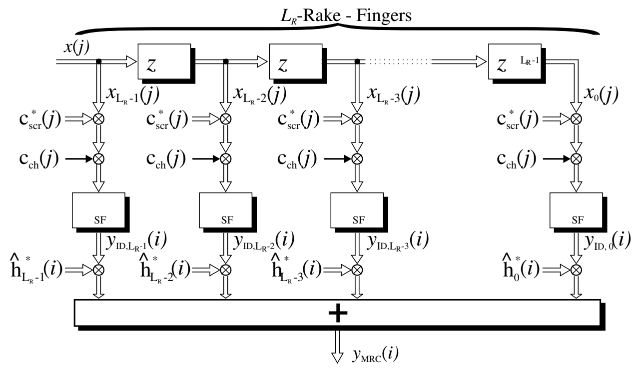

Figure 2: Rake-receiver

This is the optimal approach in the single user case but does not take into account the interference of all other users. Since all channelization codes are orthogonal to each other the bad performance of such a system can be interpreted by the loss of orthogonality due to multipath propagation. Therefore restoring orthogonality with an equalizer is a prudent thing to do especially in systems with a high load. In [2] for example the feasibility of a Wiener approach using adaptive algorithms such as RLS or LMS to adapt to the time variance of a channel is shown. In our approach the main idea is quite similar with the exception, that we try to use an oversampled input signal and that we omit the receiving root raised cosine filter.

2. Fractional T-Equalizer in CDMA-Systems

The overall sampled impulse response of the Tx- and Rx- impulse shaping filter and the impulse response of the channel is given by:

$$f(k) := f_s(kT_s) = g_{TX}(t) * h(t) * g_{RX}(t)|_{t=kT_s} \tag{4}$$

[1] (*) denotes conjugate complex

$$\text{with} \qquad T_s = \frac{T_c}{2}. \qquad \text{yielding} \qquad w = 2$$

In our case the oversampling factor w between Chip duration T_c and sampling period T_s is 2 and the transmitter and receiver root raised cosine filter are denoted as $g_{TX}(t)$ and $g_{RX}(t)$. The Zero-Forcing solution of this problem can be derived by reducing eq. 5 to eq. 6 and can be reformed to eq. 7 [3].

$$\underbrace{\begin{pmatrix} f_0 & 0 & \cdots & \\ f_1 & f_0 & & \mathbf{0} \\ f_2 & f_1 & & \\ & & \ddots & \\ \vdots & \vdots & & f_0 \\ f_{m_2} & f_{(m_2-1)} & \cdots & \\ 0 & f_{m_2} & & \\ \vdots & & \ddots & \vdots \\ \mathbf{0} & & \cdots & f_{m_2} \end{pmatrix}}_{\mathbf{F}} \cdot \begin{pmatrix} e_0 \\ e_1 \\ \vdots \\ e_{n_2-1} \\ e_{n_2} \end{pmatrix} \stackrel{!}{=} \begin{pmatrix} * \\ 0 \\ * \\ \vdots \\ * \\ 0 \\ * \\ 1 \\ * \\ 0 \\ * \\ \vdots \\ * \\ 0 \\ * \end{pmatrix} \tag{5}$$

This yields for example for a f of order 6 with $*$ denoting "do not cares" and the even f_κ denoting are in chip duration and the odd ones are directly between them.

$$\underbrace{\begin{pmatrix} f_1 & f_0 & 0 & 0 & 0 \\ f_3 & f_2 & f_1 & f_0 & 0 \\ f_5 & f_4 & f_3 & f_2 & f_1 \\ 0 & f_6 & f_5 & f_4 & f_3 \\ 0 & 0 & 0 & f_6 & f_5 \end{pmatrix}}_{\mathbf{F}_2} \cdot \underbrace{\begin{pmatrix} e_0 \\ e_1 \\ e_2 \\ e_3 \\ e_4 \end{pmatrix}}_{\mathbf{e}} = \underbrace{\begin{pmatrix} 0 \\ 0 \\ 1 \\ 0 \\ 0 \end{pmatrix}}_{\mathbf{i}} \tag{6}$$

$$\mathbf{e} = \mathbf{F}_2^{-1}\mathbf{i} \tag{7}$$

Alas the Zero-Forcing approach is very sensitiv to noise due to the noise magnification of this filter.
Instead of eq. 7 the MMSE (Wiener) solution can be found [4] for the fractional T-equalizer by introducing an additional error vector ϵ

$$\mathbf{F}_2 \cdot \mathbf{e} = i + \epsilon \tag{8}$$

and a cost function

$$\Psi = \epsilon^T \epsilon + \gamma \cdot \mathbf{e}^T \cdot \mathbf{e}, \tag{9}$$

which has to be minimized in respect to $\mathbf{e}$. This results in the MMSE solution for the fractional T-equalizer

$$\mathbf{e} = \mathbf{F}_2^H \left(\mathbf{F}_2 \mathbf{F}_2^H + \gamma \mathbf{I}\right)^{-1} \mathbf{i}. \tag{10}$$

The factor γ can be seen as compromise between noise magnification and remaining MAI. In our case we assumed that each signal for every user is broadcast with the same power. In this case the number of users has to be taken into account. For the classical MMSE-filter in the single user narrow-band environment γ is normally set to:

$$\gamma^2 = \frac{N_0}{E_S} \tag{11}$$

In this case the system normally deals with the effects of ISI of one user. In a spread system in a single user case[2] $\frac{N_0}{E_S}$ will change to $\frac{N_0}{E_c}$ with E_c denoting the energy per chip and SF denoting the spreading factor resulting in

$$\gamma^2 = \frac{N_0}{E_c} = \frac{N_0}{E_s} \cdot SF\ . \tag{12}$$

For a higher number of users N_u the MAI raises in a linear manner therefore γ^2 has to be reduced accordingly.

$$\gamma^2 = \frac{N_0}{E_s} \cdot \frac{SF}{N_u} \tag{13}$$

For the MMSE-solution the number of coefficients of $\mathbf{e}$ can be enlarged to gain better performance. Unfortunately the larger the number of coefficients the smaller the performance enhancement will be.
Besides of the normal approach using a root raised cosine filter at the receiver, this approach is neglected because the equalizer itself will perform this task. The resulting S/N loss is expected to be negligible. In our case a root raised cosine filter for $w = 2$ with 21 taps is taken. For the Zero-Forcing-solution with a filter of the same order and in case of a perfect channel $c(t) = \delta(t)$ this yields a S/N loss of about 0.26 dB. The loss itself will degrade while using the MMSE version and or more filter coefficients. The resulting system is depicted in fig. 3b). Figure 3a) represents the system using a normal MMSE-Equalizer working in the T_C-domain. The smaller number of coefficients due to the missing receiver filter is an additional advantage.

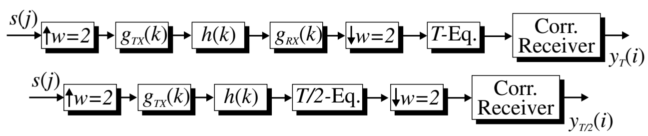

Figure 3: Simulation scheme for the a) T and b) fractional T-equalizer

3. Simulation Results

In this section simulation results are presented. The channel estimation is considered to be perfect. For all examined cases the Rake-receiver as depicted in figure 1 and the T-Equalizer (MMSE) approach depicted in fig. 3a) are simulated as a reference for the fractional T-equalizer according to fig. 3b). All simulations are done for block fading channels. Therefore the time variance of the channel is not taken into account.

[2]Of course we would prefer the MRC-receiver in this case

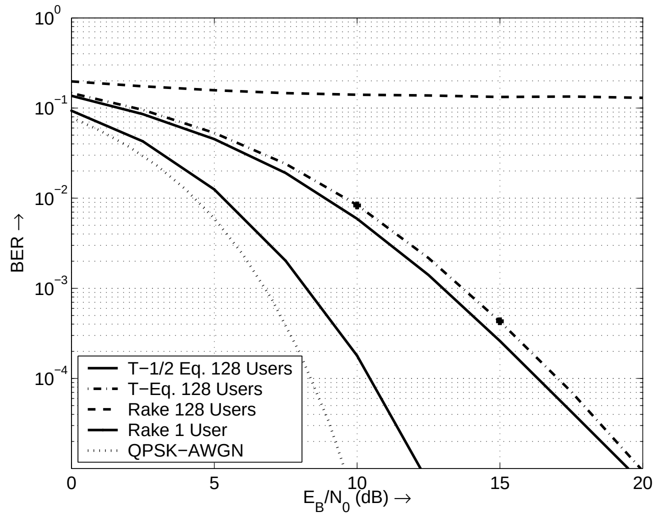

Figure 4: BER curve for 20 path Rayleigh block fading channel and 128 users

In fig. 4 a spreading factor of 128 is taken and signals for 128 users with the same transmission power are broadcast. Beside the following simulations the classical AWGN-curve and a curve for a system with only one user is displayed as reference. The Rayleigh channel has 20 coefficients in the oversampled case and the length of the fractional T-Equalizer filter is set to 161 taps. The number of filter coefficients of the reference T-Equalizer filter is set to 121. Keep in mind that the resulting delay for the T-Equalizer is even higher compared to the fractional one. A simulation with a 161-tap T-Equalizer yields nearly no gain at all which can be seen as crosses at 10 and 15 dB in figure 4.
Due to the high interference power, the Rake-receiver is not capable to reach acceptable results for this full loaded system. The T-equalizer performs quite well and the fractional T-equalizer outperforms it by approximately 0.75dB. It is needlessly to say that it is still much worse than the Rake-receiver for the one user case. The Rake-receiver itself is in this kind of environment not usable anymore with an error floor of approximately 12 % For a half loaded system (64 users) the relationship of all curves to each other are mostly the same compared to fig. 4 beside that all BERs are better.
In fig. 5 the BER curves for a 8-path Rayleigh channel are displayed. The fractional T-equalizer is still about 0.75 dB better compared to the T-equalizer. Considering fig. 4 the steepness of the curves for the 8-Path case is a little bit lower because of the lesser diversity due to the channel.
For channel models with delays according to the chip duration like most 3GPP models, defined in [5], the fractional T-equalizer can not achieve any gain since due to the zeros for the fractional taps for this channel it degrades to a normal T-equalizer.

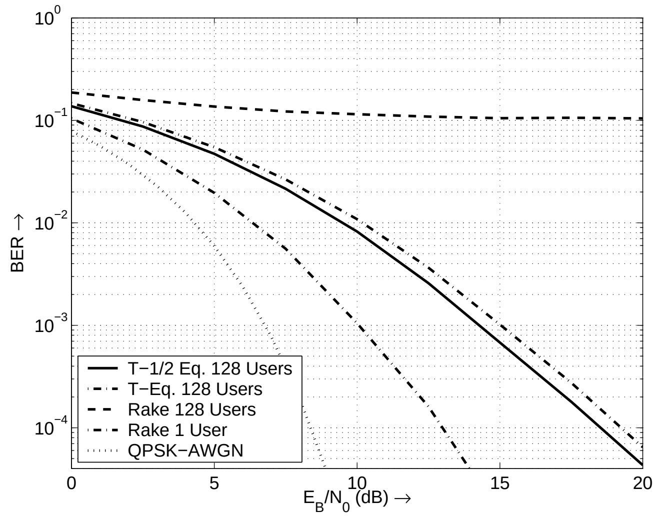

Figure 5: BER curve for 8 path Rayleigh block fading channel and 128 users

4. Conclusions

In this paper the use of a fractional tapped delay line equalizer is proposed. Under perfect conditions it results in a gain of nearly 1 dB compared to the use of a tapped delay line equalizer in a full loaded system. The normal rake receiver is no match for these high numbers of users. Nevertheless this paper gives only a glimpse that this scheme may work in a modern communication system. This results in further activities that have to examine first. In our simulations we used the direct solution by inverting matrices. Unfortunately this is not a good idea for a practical system due to ist high computational cost $O(n^3)$. Therefore an iterative method, like using the RLS or LMS algorithms, have to be applied analogue to that in [2] for the normal Wiener case in chip domain. Another problem is of course the channel estimation which has to be estimated in the oversampling domain.

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