Linear baseband modelling of Direct-Sequence Code-Division Multiple Access systems *

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Abstract: In this paper, the entire process of transmission and reception in a DS-CDMA system, including the spreading, is described as a tapped delay line with multiple inputs and a single output. The model is then used to design a fractionally spaced decision feedback equalizer with a single input and multiple outputs. In the proposed approach, long codes can be used and channel estimation can be performed efficiently. The suggested detector is near-far resistant. Monte Carlo simulations demonstrate the utility of the proposed approach in a heavily loaded system with and without power control where multipath is present.

Introduction

Much of the flexibility in the design of a DS-CDMA system relies on the use of *long spreading codes*. The use of long codes makes it unnecessary to use code planning and makes it easy to supply variable rate services.

The use of non-orthogonal channels makes the DS-CDMA system very sensitive to the *near-far problem*. The common solution, power control in combination with the conventional detector, has serious drawbacks. To circumvent the near-far problem and thus eliminate the need for power control, the use of so-called *multiuser detectors* has been proposed.

Many proposed multiuser detectors operate on the outputs from a bank of matched filters. They are frequently presented as block detectors, and multipath propagation and long codes are rarely considered explicitly. One example is the decorrelating detector [1]. A disadvantage with these detectors is that the propagation delay must be well known. In a near-far scenario, the propagation delay can be hard to estimate, as illustrated in [2].

Other multiuser detectors [3, 4, 5] operate directly on the received wideband signal, which is sampled at the chip rate. The received signal is passed through an FIR filter followed by a decision device. The filters are adaptive and they are updated recursively directly from the received signals, making estimation of the propagation delay unnecessary. However, long codes are impossible to use, and low bitrates could result in a difficult tracking problem.

An ideal detector for DS-CDMA should be near-far resistant like the decorrelating detector. It should also be insensitive to errors in propagation delay estimates. Finally, the detector should be able to operate in a system where long codes are used.

In this paper, we reformulate the standard DS-CDMA system transmission and reception as an equivalent, discrete time, tapped delay line model. This model takes the data symbols as input and has the wideband, chip sampled, received signal as output. The spreading operation is represented as linear filtering. This model makes it possible to design equalizers which, directly from the chip sampled signal, filter out the signals of the respective users.

As an example of a detector which has been derived from this model, we subsequently outline a generalization of the fractionally spaced decision feedback equalizer (DFE). This

*This work was supported by the NUTEK research program in Telecommunications. The research forms a part of the project "Robust and adaptive methods for equalization, tracking and diversity".

DFE has one input and the same number of outputs as the number of users. We then investigate the performance of the proposed DFE without and without any power control. It is demonstrated that the DFE works without problems in both scenarios.

System model

In this section a DS-CDMA system will be reformulated as a completely discrete time, tapped delay line model with multiple inputs and a single output.

The DS-CDMA system

We are considering an asynchronous DS-CDMA system with K users. The modulation scheme is BPSK. The symbol period is denoted by T_s, whereas T_c represents the duration of a chip. The processing gain, i.e. the ratio T_s/T_c, is denoted by N_c. Without loss of generality, we set $T_c = 1$. User k transmits the symbol $d_k(tN_c)$ during the time period $[tN_c, (t+1)N_c[$. Each symbol is spread by a wideband signature sequence.

The transmitted passband signal propagates through a frequency selective, wideband channel. The receiver front end consists of a conventional IQ-stage, where the I and Q signals are down-converted to the baseband. The received baseband signal is then passed through a filter matched to the pulse shape of the chip. At the output of this *chip-matched filter*, the signal is sampled at the chip rate, resulting in the complex-valued signal $r(\cdot)$.

A linear baseband model: the single-user case

Traditionally, both the symbol sequences and the signature sequences in a DS-CDMA system are represented as *continuous time* signals. To arrive at a *discrete time* model of the system, all pulse shaping is assumed to take place *after* the spreading. Next, the pulse shaping, the frequency up-conversion, the physical channel, the frequency down-conversion and the chip-matched filter are lumped together and replaced by an equivalent discrete time channel. This equivalent representation will in the following be called the *physical channel*. This means that the transmission and reception can be represented as in Figure 1.

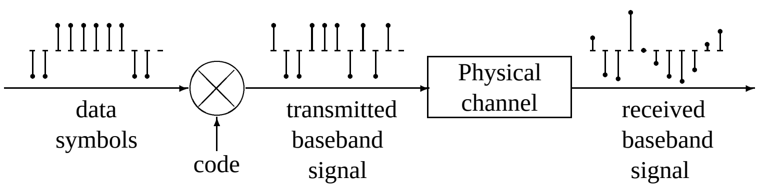

Figure 1: A completely discrete time model, sampled at the chip rate.

To arrive at a tapped delay line model, the spreading must be represented by a linear filter. We would then obtain an *equivalent channel* which can be represented by the convolution of this *spreading filter* and the physical channel. Let us therefore view the symbol sequence as an input and the spread signal as an output of a linear filter. Consider the situation during one symbol period, as depicted in Figure 2.

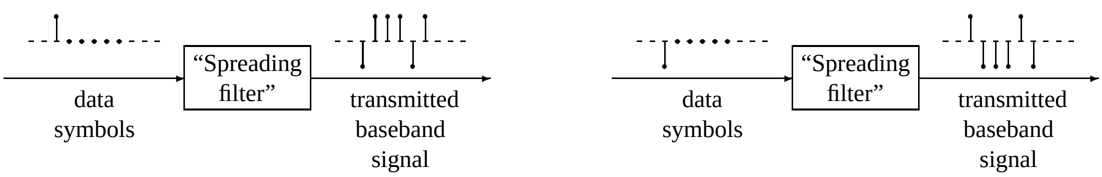

Figure 2: The spreading viewed as a filtering operation. Input symbol $+1$ (left) and -1 (right). Note that the symbol sequence is assumed to be non-zero only every N_c chips!

This means that during one symbol period, *the spread signal can be thought of as having been generated by passing the symbol through a discrete time filter of length N_c with impulse response identical to the code sequence.*[1]

A multiple-input single-output channel model

Each tapped delay line is the convolution between a spreading filter and a physical channel. Since the length of the spreading filter is equal to the symbol period, intersymbol interference will be present if any physical channel has more than one tap.

The entire channel can now be represented by a linear model with K inputs and one output. To formulate this mathematically, introduce the quantities

$$\begin{array}{rl} h^k_{iN_c+j}(tN_c+j): & \text{tap } iN_c+j \text{ in the channel impulse response from user } k \text{ at time } tN_c+j \\ m \quad : & \text{maximum extent (over all channels) of the intersymbol interference.} \end{array}$$

The received signal will be the noise-corrupted sum of the signals from the K users:

$$r(tN_c+j) = \sum_{k=1}^{K}\sum_{i=0}^{m} h^k_{iN_c+j}(tN_c+j)d_k((t-i)N_c) + n(tN_c+j) \tag{1}$$

where $n(tN_c+j)$ is assumed to be wide sense stationary zero mean noise. Introduce

$$\begin{aligned} d(tN_c) &\triangleq \begin{pmatrix} d_1(tN_c) & \ldots & d_K(tN_c) \end{pmatrix}^T \\ \mathbf{h}^{tN_c+j}_{iN_c+j} &\triangleq \begin{pmatrix} h^1_{iN_c+j}(tN_c+j) & \ldots & h^K_{iN_c+j}(tN_c+j) \end{pmatrix} . \end{aligned} \tag{2}$$

Then the received signal (1) can be rewritten as

$$r(tN_c+j) = \sum_{i=0}^{m} \mathbf{h}^{tN_c+j}_{iN_c+j} d((t-i)N_c) + n(tN_c+j) . \tag{3}$$

Equation (3) is the desired K-input single-output channel model, which relates the sequence of symbol vectors to the chip sampled output sequence.

To make use of this model to design a DS-CDMA detector, we have to estimate the channel coefficient vector (2) for $i = 0, 1, \ldots, m$ and $j = 0, 1, \ldots, N_c - 1$. We then have to design an equalizer based on the impulse response coefficients (2). When performing the channel estimation, we suggest the use of the transmitted chip sequence as regressors to identify only the physical channel. The complete channel is then obtained by convolving the (estimated) physical channel with the (known) spreading filter.

A good equalizer candidate is the fractionally spaced decision feedback equalizer (DFE), which will be outlined next.

The single-input K-output fractionally spaced DFE

Our proposed decision feedback equalizer operates directly on the chip sampled signal $r(\cdot)$. The outputs from the equalizer are the data symbols of the K users. This means that the proposed DFE will be a single-input multiple-output equalizer.

The structure of the DFE is depicted in Figure 3. The scalar, chip sampled signal $r(\cdot)$ is used as input to the feedforward filter and the K outputs of the feedforward filter are sampled at the symbol rate.

[1]For short codes, the spreading filter will be time-invariant, but for long codes it will be time-varying.

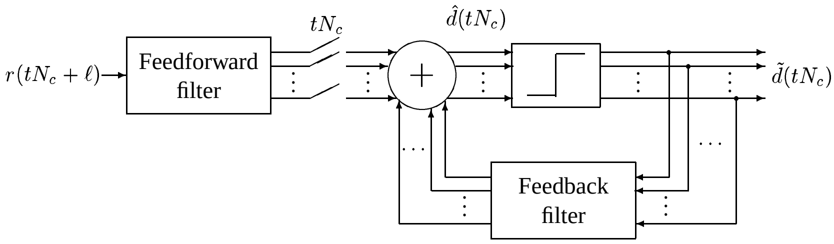

Figure 3: The single-input multiple-output fractionally spaced DFE.

The DFE estimates the symbol vector that was transmitted at time tN_c, given data up to time $tN_c + \ell$. The design variable ℓ is known as the *smoothing lag* or *decision delay.*

The coefficient matrices of the two filters are adjusted so that the mean square error of the symbol vector estimate $\hat{d}(tN_c)$ is minimized. To calculate the optimal filters, we solve a system of linear equations. For the derivation of the optimal filters, see [6].

If long codes are used, the known part of the channel will change completely every symbol period. The time-varying nature of the channel is taken into account by including coefficients of *future* channel models in the calculation of optimal filter coefficients. The proposed DFE is a symbol-by-symbol detector, making it possible to implement without modification.

The computational complexity of the calculations necessary to optimize the equalizer coefficients is rather high. To calculate the coefficients of the feedforward filter approximately $(\ell+1)^3/6 + K(\ell+1)^2$ multiplications have to be performed. The number of multiplications necessary to calculate the feedback filter is proportional to $K(\ell+1)^2$. Using the DFE to estimate a transmitted symbol vector then requires $(\ell+1)K + mK^2$ multiplications. Notice that the complexity is not exponential in any system parameter.

Simulations

We will now investigate the utility of our channel model, as well as the usefulness of our proposed DFE. We are using *long* codes in a system with processing gain $N_c = 8$ and $K = 5$ users. This represents a rather heavily loaded system.

All physical channels have four Rayleigh faded taps. During transmission of a burst, the taps are time invariant. The propagation delays of the users are uniformly distributed in the interval $[0, N_c - 1]$ (in units of the chip period). For the simulation, we assume that we know the impulse response of the channel.

To emphasize the near-far problem in a system with no power control, four of the users have the same average power, whereas the fifth user has an average power that is 10, 20 or 30 dB lower. The weaker user (user 1) has an average E_b/N_0 between 5 and 20 dB. The performance measure is the BER of the weaker user.

The result is shown in Figure 4 on the following page, where it is also shown that the DFE outperforms the conventional receiver with a four-finger RAKE in a system with perfect power control. We see that the near-far effect has an impact on the DFE, but the effect is small. Increasing the power of the four stronger users by 20 dB results in a deterioration of approximately 1 dB for the entire investigated range of SNR:s.

Conclusions

In this paper, we have described the complete transmission and reception process in a DS-CDMA system, including the spreading, by a discrete time tapped delay line model. The model takes as the input the symbols from all users in the system and outputs the received signal, sampled at the chip rate. By using this discrete time model, the problem

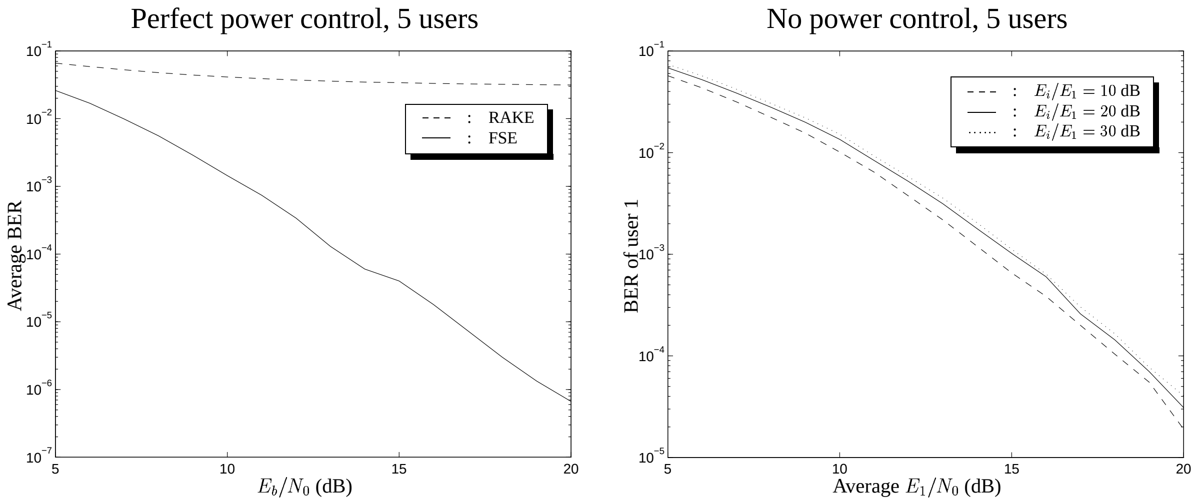

Figure 4: BER as a function of E_b/N_0. Perfect power control (left) and no power control (right)

of detecting the signals of the users has been recast as a deconvolution problem. Also, channel estimation can be handled in a way that reduces problems with fast fading.

Based on the proposed model, we have subsequently derived a fractionally spaced decision feedback equalizer. This DFE detects all signals simultaneously, thereby making it possible to operate in an environment without power control. In contrast to many other multiuser detectors, it is derived as a symbol-by-symbol detector. Another advantage is that due to the fast sampling, the DFE only requires rough synchronization.

The DFE can be used in systems where long codes are employed. Since the DFE effectively decorrelates the users, a large part of the capacity loss commonly associated with long codes can be eliminated.

The amount of computations that has to be performed to update the equalizer is substantial, but it is not exponential in any system parameter.

The simulations in Section 4 show that in a heavily loaded system, the DFE works well, even in a rather severe near-far situation.

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