A Linear Receiver for WCDMA Downlink Exploiting the Coloration of the Interference

Hafez Hadinejad-Mahram
Institute of Communications Engineering
Aachen University
Melatener Str. 23, 52074 Aachen, Germany
E-mail: hafez@ient.rwth-aachen.de

Gholamreza Alirezaei
Institute of High Frequency Technology
Aachen University
Melatener Str. 25, 52074 Aachen, Germany
E-mail: alirezaei@ihf.rwth-aachen.de

***Abstract*— In 3rd generation mobile radio systems based on direct-sequence code-division multiple access the employment of more sophisticated demodulation techniques is indispensable for the provision of high data rates with a reasonable error performance. In this paper, we propose a modification of a recently introduced scheme which exploits the coloration of the interference at receiver input. In this approach, the maximum likelihood criterion is used to derive a time-variant equalizer which outperforms the RAKE receiver significantly. With the modification proposed the performance gain can be increased by several dB.**

I. Introduction

Currently, the performance requirements for wideband code-division multiple access (WCDMA), including the requirements for high-speed downlink packet access (HSDPA), an extension of WCDMA for high data rate packet access, are extensively being discussed. For the latter, it has been proposed, that the performance requirements should be such that compliance cannot be achieved with the standard RAKE receiver [1]. This implies that an alternative receiver architecture with an improved performance compared to that of the standard RAKE receiver is needed. For a reasonable dimensioning of the required improvement, the performance of alternative receivers has to be compared and related to their complexity. Several advanced WCDMA downlink detection algorithms have been proposed in the open literature [2–7].

In WCDMA downlink, MAI is caused only by the multipath channel, and all users' signals reach the mobile station (MS) through the same channel. Thus, the orthogonality between the users' signals can be restored by equalizing the common multipath channel. This notion has lead to the development of linear chip equalizers based on the zero-forcing or the LMMSE (linear minimum mean square error) criterion [2, 6], and numerous adaptive versions thereof [5, 7].

A modification of the RAKE receiver with MAI suppression capability has been proposed in [4] and termed *Generalized* RAKE (G-RAKE). In contrast to the conventional RAKE receiver, G-RAKE combining takes into account that a) the finger outputs are correlated and b) the interference level is in general different for each finger. Using the optimal combining weights is not, however, the only key to the noticeable performance gains over the conventional RAKE as reported in [4]. The most significant part of the improvement arises form using fingers placed at delays not corresponding to channel rays.

Another approach, referred to as colored noise detector (CND), has been proposed in [3]. Here, the MAI is modeled as white Gaussian noise at the transmitter, which results in colored Gaussian noise at the receiver. Based on the maximum likelihood (ML) criterion they derive a receiver algorithm which operates on a specified interval of received samples per detected symbol. While fixing the observation interval of the receiver avoids the need for the cumbersome optimization of the equalizer filter length as e.g with LMMSE based receivers, it also limits the achievable performance gain as was shown in [8].

In this paper, we propose a modification of the approach in [3] by relaxing the restrictions in the design rule. Specifically, both the length of the observation interval and its position relative to the symbol to be detected are adjustable design parameters. Optimum detection of signals corrupted by additive colored noise often requires that the observation interval of the detector be infinitely large due to the correlation of the noise samples. In a practical implementation, however, the detector must make a decision based on a small number of received chips. Therefore, one has to find out which samples to take so as to optimize the performance. In other words, once we have fixed the number of chips to be considered for the decision of a symbol, we have to determine their position with respect to that symbol. In [3], all samples depending on the transmitted symbol of interest are considered, i.e. the received chips corresponding to the symbol itself as well as those of the following symbol interval which are affected by that symbol due to the dispersion of the channel. This choice fixes both the length of the observation interval and its position. The reasoning behind this choice is that these samples include all available information on the symbol of interest. However, this argumentation ignores the fact that not only the information about the symbol is important for the detector but also the information about previous and future noise samples. In our approach, we allow for a trade off between these two

0-7803-7661-7/03/$17.00

information sources. By flexibly choosing the position of the observation interval we can e.g. sacrifice some information about the symbol in order to look back and take into account the information arising from noise samples of the previous symbol intervals.

The paper is organized as follows. Sect. II describes the system model used in this paper. In Sect. III we derive the receiver structure and discuss the differences compared to the algorithm in [3]. Sect. IV is devoted to numerical examples which reveal the superiority of our approach over the one in [3]. Finally, Sect. V presents our conclusions.

II. System Model

The system model used in this paper is based on the 3GPP specifications according to [9, Appendix A.3.3 and A.3.4]. The desired user has two dedicated logical channels, one for data and one for control information, each of which is mapped onto a dedicated transport channel. The data on each dedicated channel is delivered to the physical layer in so called transport blocks (TB) (or transport block sets) which contain a certain number of bits to be transmitted in a predetermined time, the transmission time interval (TTI). Each TB is processed in the physical layer in several steps: first a specified number of CRC (cyclic redundancy check) bits and a coding-type dependent number of termination bits are appended to it. Then, each of the new blocks is passed through a rate R channel (convolutional or turbo) encoder. Further steps include rate matching, first (or inter-frame) interleaving, radio frame segmentation, and transport channel multiplexing (for details see [10]). The latter is a parallel-to-serial multiplexer which creates a single data stream from the two transport channels. Afterwards, the bit sequence passes the second (or intra-frame) interleaving before being split into segments that fit into a slot. In each slot a certain number of bit positions is reserved for control information generated in the physical layer such as power control commands and pilot bits. These bits along with the payload are used to build a slot as described in [11]. Finally the bits are QPSK (quaternary phase shift keying) modulated and spread by a complex spreading sequence which is the product of a user-specific orthogonal variable spreading factor (OVSF) code [12] and the base station (BS) specific long scrambling code. Before transmission over a time-variant multi-path fading channel the spread signals of other users as well as several physical channels carrying only information relevant to the physical layer procedures are added to the signal of the desired user. The latter include the synchronization channel (SCH), the primary common control physical channel (P-CCPCH), the primary common pilot channel (P-CPICH), and the paging indication channel (PICH). Each one of these channels uses a specified fraction of the total transmission power of the BS [9]. Moreover, the other users' signals are simulated by the orthogonal channel noise source (OCNS) according to [9]. Note that all physical channels transmitted from a BS except SCH are mutually orthogonal.

The above description of the system model has been intended to provide a picture of the simulation setup used in this paper. For the derivation of the receiver structure, however, we need a more mathematical description which will be presented in the following.

Consider the equivalent lowpass representation of a K user synchronous and orthogonal direct-sequence CDMA system. The k-th user's source generates a sequence of equiprobable and independent symbols $\{d_k(i) \in \{\pm\frac{1}{\sqrt{2}} \pm \jmath\frac{1}{\sqrt{2}}\}\}$. The i-th symbol $d_k(i)$ is spread by the user-specific and symbol-period-dependent complex spreading sequence $c_{k,i}(m) \in \{\pm\frac{1}{\sqrt{2N}} \pm \jmath\frac{1}{\sqrt{2N}})\}, m = 0, \cdots, N-1$, where N is the spreading factor. Afterwards, the spread symbols of other users are added to it and subsequently, this composite chip sequence is convolved with the unit-energy Nyquist pulse shape $p(t)$ resulting in the total transmit signal

$$s(t) = \sum_{i=-\infty}^{\infty} \sum_{k=1}^{K} \sqrt{E_k} d_k(i) \sum_{m=0}^{N-1} c_{k,i}(m) p(t - mT_c - iT) \tag{1}$$

where T_c and $T = NT_c$ are the chip and the symbol duration, respectively, and E_k is the energy per symbol of k-th user.

The channel between the BS and the desired MS is assumed to be a frequency-selective Rayleigh fading channel modeled as a chip-spaced FIR filter with time-variant coefficients [1] whose impulse response can be expressed as

$$h(t, \tau) = \sum_{l=0}^{L} h_l(t) \delta(\tau - lT_c) \tag{2}$$

where LT_c is the channel length and $h_l(t)$ is the complex-valued coefficient of the l-th tap. In the derivation of the receiver algorithm we assume that the channel is slowly time varying such that changes over the observation window of the receiver are negligible. Thus, we can rewrite (2) as a time-invariant impulse response

$$h(t) = \sum_{l=0}^{L} h(l) \delta(t - lT_c)$$

Furthermore, we assume the average channel energy

$$\sum_{l=0}^{L} \mathrm{E}\{|h(l)|^2\}$$

to be normalized to unity where $\mathrm{E}\{\cdot\}$ denotes expectation.

The received signal composed of the radio channel output superimposed by additive white Gaussian noise (AWGN) $n(t)$ of two-sided power spectral density N_0 which models the inter-cell interference and the receiver noise can be expressed as

$$r(t) = \sum_{l=0}^{L} h(l) s(t - lT_c) + n(t).$$

In the receiver, $r(t)$ is filtered by a chip matched filter and sampled at chip rate resulting in samples $y(nT_c)$ given by

$$
\begin{aligned}
y(nT_c) &= \int_{-\infty}^{\infty} r(\tau)p(\tau - nT_c)\, d\tau \\
&= \sum_{l=0}^{L} h(l) \int_{-\infty}^{\infty} s(\tau - lT_c)p(\tau - nT_c)\, d\tau \\
&\qquad\qquad + z(nT_c)
\end{aligned}
$$

where $z(nT_c) = \int_{-\infty}^{\infty} n(\tau)q(\tau - nT_c)\, d\tau$. Substituting (1) for $s(t)$ and evaluating the integral yields

$$
\begin{aligned}
y(nT_c) = \sum_i \sum_{k=1}^{K} \sqrt{E_k} d_k(i) \sum_{l=0}^{L} c_{k,i}(n-l)h(l) \\
+ z(nT_c) \qquad (3)
\end{aligned}
$$

where we have exploited the fact that $p(t)$ is a Nyquist pulse.

III. Receiver Structure

Since the receiver operates on blocks of chips, it is more convenient to use a vector-matrix notation in deriving the receiver structure. Let vector $\mathbf{y}^{(\nu,M)}$ be the equalizer observation window where M and ν denote the length of $\mathbf{y}^{(\nu,M)}$ and the index of the first sample of $\mathbf{y}^{(\nu,M)}$, respectively, i.e.

$$
\mathbf{y}^{(\nu,M)} = [y(\nu), \cdots, y(\nu + M - 1)]^T
$$

where we have omitted T_c in the argument of y. $\mathbf{y}^{(\nu,M)}$ can be expressed using a matrix-vector product as

$$
\mathbf{y}^{(\nu,M)} = \sum_i \sum_{k=1}^{K} \sqrt{E_k}\, d_k(i)\, \mathbf{H}\mathbf{c}_{k,i}^{(\nu,M+L)} + \mathbf{z}^{(\nu,M)} \quad (4)
$$

where $\mathbf{z}^{(\nu,M)}$ is the vector of the corresponding noise samples,

$$
\mathbf{H} = \begin{bmatrix}
h(L) & \cdots & h(0) & 0 & \cdots & 0 \\
0 & \ddots & & \ddots & \ddots & \vdots \\
\vdots & \ddots & \ddots & & \ddots & 0 \\
0 & \cdots & 0 & h(L) & \cdots & h(0)
\end{bmatrix}
$$

with $\mathbf{H} \in \mathbb{C}^{M \times (M+L)}$, and

$$
\mathbf{c}_{k,i}^{(\nu,M+L)} = [c_{k,i}(\nu - L), \cdots, c_{k,i}(\nu + M - 1)]^T.
$$

Note that $c_{k,i}(n)$ is zero for $n \leq iN$ and $n \geq (i+1)N$.

In the following we assume, without loss of generality, that the 0-th symbol of user 1 is to be detected.

We can rewrite $\mathbf{y}^{(\nu,M)}$ as

$$
\mathbf{y}^{(\nu,M)} = d_1(0)\mathbf{H}\mathbf{c}_{1,0}^{(\nu,M+L)} + \mathbf{H}\,\mathbf{q}^{(\nu,M+L)} + \mathbf{z}^{(\nu,M)}
$$

where the first term in the right-hand side of the equation is the noiseless received signal of user 1 in 0-th period and

$$
\mathbf{q}^{(\nu,M+L)} = \sum_{\substack{i \\ (i,k)\neq(0,1)}} \sum_{k=1}^{K} \sqrt{E_k}\, d_k(i)\, \mathbf{c}_{k,i}^{(\nu,M+L)}.
$$

contains all intersymbol interference (ISI) and MAI terms. Following the suggestion of Jarosch and Dahlhaus [3], we model $\mathbf{q}^{(\nu,M+L)}$ as AWGN. Thus $\mathbf{H}\,\mathbf{q}^{(\nu,M+L)}$ represents colored noise. Now, the ML criterion leads to the following decision variable

$$
\zeta = \left(\mathbf{c}_{1,0}^{(\nu,M+L)}\right)^H \mathbf{H}^H \mathbf{K}_{\mathbf{y}}^{-1}\, \mathbf{y}^{(\nu,M)} \qquad (5)
$$

where $\mathbf{K}_{\mathbf{y}}$ is the covariance matrix of $\mathbf{y}^{(\nu,M)}$ given by

$$
\mathbf{K}_{\mathbf{y}} = E_T \mathbf{H}\,\mathbf{H}^H + N_0 \mathbf{I}
$$

where E_T is the total transmitted energy. Note that in the derivation of $\mathbf{K}_{\mathbf{y}}$ several simplifying assumptions are used which however have been reported in [3] to be non-critical with respect to the error performance of the detector.

The product of $\mathbf{y}^{(\nu,M)}$ by $\mathbf{H}^H\mathbf{K}_{\mathbf{y}}^{-1}$ on the right-hand side of (5) describes the filtering of the input signal $\mathbf{y}^{(\nu,M)}$ by a time-variant (with respect to the chips within one symbol) FIR filter $\mathbf{H}^H\mathbf{K}_{\mathbf{y}}^{-1}$. Herein, $\mathbf{K}_{\mathbf{y}}^{-1}$ includes a finite-length whitening filter and its matched filter, and $\mathbf{H}^H$ denotes the channel matched filter. Furthermore, the multiplication of $\left(\mathbf{c}_{1,0}^{(\nu,M+L)}\right)^H$ from left is the despreading operation.

The receiver derived above has two parameters, ν and M, which should be chosen such as to optimize the receiver performance. The length of the observation window M is usually limited by the maximum tolerable complexity of the receiver. Thus our focus should be the optimization with respect to ν. Unfortunately, this task is not a trivial one. A similar problem occurs in LMMSE chip equalizers and in G-RAKE [4]. In both cases, the solution is usually found by numerically searching for the value that optimizes a target function. We will also follow this approach where the target function is sensibly chosen to be the bit error rate (BER). Note that choosing ν is a tradeoff between matching the channel and whitening the noise.

The CND operates on a an observation interval which includes all samples depending on the symbol $d_1(0)$. Stated in our notation this means that $\nu = 0$ and $M = N + L$. In [3] these choices are understood as a natural result of the ML scheme employed. However, as described above this is a misconception and the optimal choice of ν for a given M has to be found heuristically.

IV. Numerical Results

For a fair comparison we fixed the length of the observation window for both receivers to that of the CND, i.e. $M = N + L$.

We simulated the downlink reference measurement 384 kbps according to [9] A.3.4. The spreading factor of the dedicated physical channels is 8. Furthermore, the relative power of the desired user to the total transmitted power was chosen -9 dB. Moreover, the powers of P-CCPCH, P-CPICH, SCH, and PICH are specified in [9], thus we can calculate the OCNS power by subtracting the powers of the above channels from the total transmit power. A total of 10^7 symbols were simulated.

Fig. 1 shows the BER curves for the CND, the modified CND (MCND), and the conventional RAKE receiver in the ITU out-to-in B channel which is a very dispersive channel. While the gain of CND over the RAKE receiver is huge there is still a considerable gain to be achieved by optimizing ν. For example at a BER of 10^{-2} the MCND displays a gain of 6 dB compared to the CND.

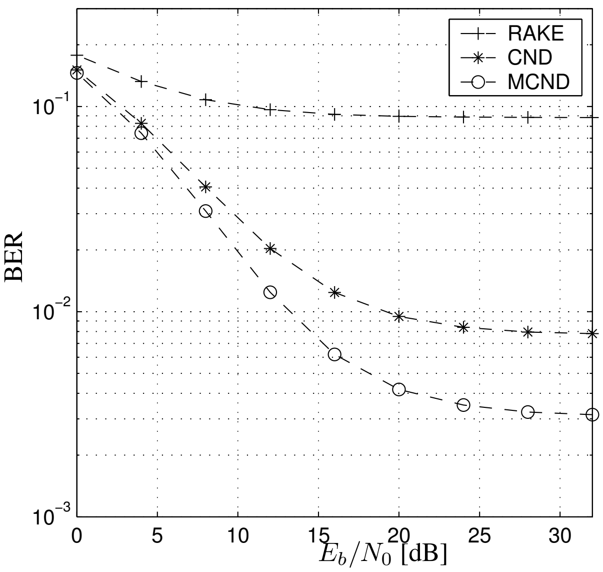

Fig. 1: BER vs. E_b/N_0. ITU out-to-in B channel.

Fig. 2 shows the BER curves for the ITU vehicular A channel which is less dispersive than the ITU out-to-in B channel. However, the performance improvement of the MCND is still noticeable.

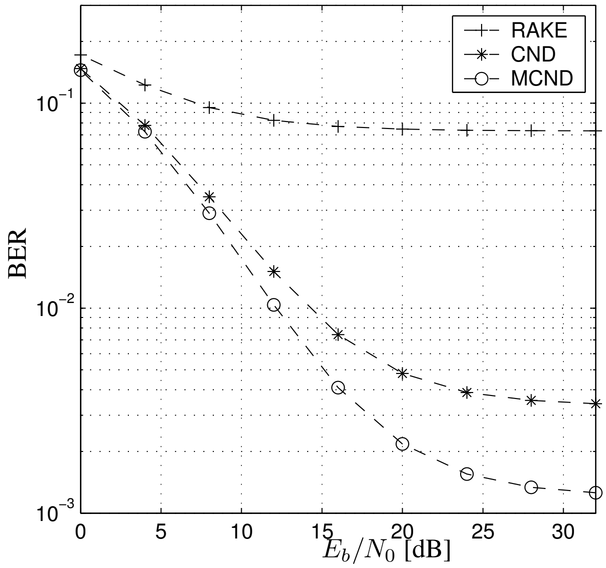

Fig. 2: BER vs. E_b/N_0. ITU vehicular A channel.

V. Conclusions

A modification of the colored noise detector [3] was introduced which leads to a significant performance gain over the original version. The misconception included in the derivation of the CND was identified and a more flexible structure of the receiver was suggested. Performance gains of several dB were demonstrated using simulation results.

VI. Acknowledgment

This work was supported in part by Synopsys in Herzogenrath, Germany.

References

[1] J.G. Proakis, *Digital Communications*, McGraw-Hill, New York, 3rd edition, 1995.

[2] A. Klein, "Data detection algorithms specially designed for the downlink of CDMA mobile radio systems," in *Proc. IEEE Vehicular Technology Conf. VTC'97*, Phoenix, AZ, USA, May 1997, pp. 203–207.

[3] A. Jarosch and D. Dahlhaus, "Linear space-time receivers for downlink of UMTS with wcdma," *European Trans. Telecomm.*, vol. 12, no. 5, pp. 379–391, September/October 2001.

[4] G.E. Bottomley, T. Ottosson, and Y.-P. E. Wang, "A generalized RAKE receiver for interference suppression," *IEEE J. Selected Areas Comm.*, vol. 18, no. 8, pp. 1536–1545, August 2000.

[5] K. Hooli, M. Latva-aho, and M. Juntti, "Performance evaluation of adaptive chip-level channel equalizers in WCDMA downlink," in *Proc. IEEE Int. Conf. Comm ICC'01*, Helsinki, Finland, 2001, pp. 1974–1979.

[6] K. Hooli, M. Latva-aho, and M. Juntti, "Multiple access interference suppression with linear chip equalizers in WCDMA downlink receivers," in *Proc. IEEE Int. Global Comm. GLOBECOM'99*, Rio de Janeiro, Brazil, December 1999, pp. 467–471.

[7] M.J. Heikkilä, P. Komulainen, and J. Lilleberg, "Interference suppression in CDMA downlink through adaptive channel equalization," in *Proc. IEEE Vehicular Technology Conf. VTC'99*, Amsterdam, The Netherlands, October 1999, pp. 978–982.

[8] H. Hadinejad-Mahram, H. Elders-Boll, and G. Alirezaei, "Performance evaluation of advanced receivers for WCDMA downlink detection.," in *Proc. of The Fifth International Symposium on Wireless Personal Multimedia Communications, WPMC'2002*, Honolulu, Hawaii, USA, October 2002.

[9] "UE radio transmission and reception (FDD)," 3GPP TS 25.101, V4.5.0 (2002-06).

[10] "Multiplexing and channel coding (FDD)," 3GPP TS 25.212, V3.3.0 (2002-06).

[11] "Physical channels and mapping of transport channels onto physical channels (FDD)," 3GPP TS 25.211, V3.3.0 (2002-06).

[12] "Spreading and modulation (FDD)," 3GPP TS 25.213, V3.3.0 (2002-06).