STRUCTURED INTERFERENCE CANCELLATION AT A WCDMA TERMINAL

Ahmet Baştuğ, Dirk T.M. Slock

Eurecom Institute
2229 Route des Cretes B.P. 193 06904 Sophia Antipolis Cedex FRANCE
ahmet.bastug@eurecom.fr, dirk.slock@eurecom.fr

ABSTRACT

In third generation CDMA-based cellular mobile systems, loading fraction is going to be less than one due to interference limitations. In other words, to achieve an acceptable BER, the number of served users and hence used codes will be less than the spreading factor. Accordingly, the existence of unused codes implies the existence of a noise subspace, which can be exploited to partially cancel the intracell and intercell interference. When aperiodic scrambling is used, as in the FDD mode of UMTS, the noise subspace varies with symbol period. In this case, multi-user interference cancelling receivers should contain time varying parts. We discuss two-stage structured receivers for mobile terminals that contain (de)scrambling, (de)spreading, projection/hard-limiting/tangent-hyperbolic decision operations and linear filters the time-varying parts of which are limited to descrambling and scrambling. The first stages estimate and delete the intracell and/or intercell interference whereas the second stages operate on the resultant improved SINR signal to detect the user symbols. Performances of various different schemes are compared via simulations.

1. BASEBAND DOWNLINK TRANSMISSION MODEL

The baseband downlink transmission model of a CDMA-based cellular system is given in Fig. 1. We consider only one interfering base station, BS 2, which, in practice, can be generalized to multiple base stations. For each base station $j, j = 1, 2$, the K^j linearly modulated user signals are transmitted over the same linear multipath channel $h^j(t)$, for we assume that downlink user signals are chip-synchronous and there is no beamforming. The symbol and chip periods T and T_c are related through the spreading factor L as $T{=}LT_c$, which here is assumed to be common for all the users and for the two base stations. The total chip sequences b^j_l are the sums of the chip sequences of all the users for each BS. Every user chip sequence is given by the product between the n^{th} symbol of the k^{th} user $a^j_{k,n}$ and an aperiodic spreading sequence $w^j_{k,l}$ which is itself the product of a periodic, unit energy Walsh-Hadamard spreading sequence $\boldsymbol{c}^j_k = [c^j_{k,0} c^j_{k,1} \cdots c^j_{k,L-1}]$, and a base-station specific unit magnitude complex scrambling sequence s^j_l with variance 1 as $w^j_{k,l} = c^j_{k,l \; mod \; L} s^j_l$:

$$b^j_l = \sum_{k=1}^{K^j} b^j_{k,l} = \sum_{k=1}^{K^j} a^j_{k,\lfloor \frac{l}{L} \rfloor} w^j_{k,l} \qquad j = 1, 2 \tag{1}$$

where $\lfloor x \rfloor$ is the largest integer which is less than or equal to x. The chip sequences $b^{1,2}_l$ get transformed into continuous-time signals by filtering them with the pulse shape $p(t)$ and passing them through the multipath propagation channels $\boldsymbol{h}^1(t)$ and $\boldsymbol{h}^2(t)$. These channels become multi-channels if there are more than one sensors at the mobile terminal or oversampling is applied. We consider that there are q sensors and the receiver samples the low-pass filtered received signal m times per chip period. As a result, the total number of samples per chip becomes mq. Stacking these mq samples in vectors, we get the sampled received vector signal

$$\boldsymbol{y}_l = \boldsymbol{y}^1_l + \boldsymbol{y}^2_l + \boldsymbol{v}_l, \quad \boldsymbol{y}^j_l = \sum_{k=1}^{K^j} \sum_{i=0}^{N-1} \boldsymbol{h}^j_i b^j_{k,l-i} \quad j = 1, 2 \tag{2}$$

where

$$\boldsymbol{y}^j_l = \begin{bmatrix} y^j_{1,l} \\ \vdots \\ y^j_{mq,l} \end{bmatrix}, \boldsymbol{h}^j_l = \begin{bmatrix} h^j_{1,l} \\ \vdots \\ h^j_{mq,l} \end{bmatrix}, \boldsymbol{v}_l = \begin{bmatrix} v_{1,l} \\ \vdots \\ v_{mq,l} \end{bmatrix}. \tag{3}$$

Here $\boldsymbol{h}^j_l$ $(mq \times 1)$ represents the vectorized samples (represented at chip rate) of the overall channel $h^j(t)$, including pulse shape, propagation channel and receiver filter. The overall channels $h^j(t)$ are assumed to have the same delay spread of N chips.

When the scrambling sequences are aperiodic and when we model the symbol sequences as i.i.d., stationary, white

Eurécom's research is partially supported by its industrial partners: Ascom, Swisscom, France Télécom, La fondation CEGETEL, Bouygues Télécom, Thales, ST Microelectronics, Motorola, Hitachi Europe and Texas Instruments.

sequences, then since the chip sequences $b_l^{1,2}$ are sums of these sequences, they are also stationary and white. The intracell contribution to $\boldsymbol{y}_l$ then is a stationary vector process the continuous-time counterpart of which is cyclostationary with chip period. The intercell interference is a sum of contributions that are of the same form as the intracell contribution and therefore its contribution to $\boldsymbol{y}_l$ is also a stationary vector process. The remaining noise is assumed to be white stationary noise. Hence the sum of intercell interference and noise, $\boldsymbol{v}_l$, represented at chip rate, is stationary.

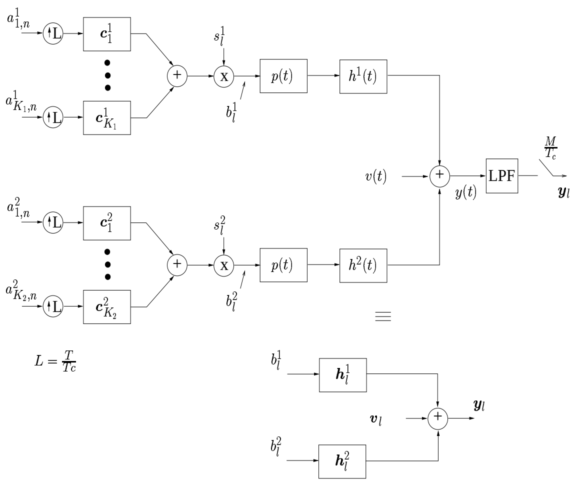

Figure 1: Baseband downlink transmission model

2. INTERFERENCE CANCELLING RECEIVER STRUCTURES

2.1. Simple Linear Receiver Structures

Fig. 2 shows a general linear receiver in which a descrambler and a correlator follow a linear FIR filter f_l whose input is at the sampling rate and its output is at the chip rate. In the case of a RAKE receiver implementation, f is the channel matched filter h^{1^H} maximizing the SNR at the filter output which considers the BS 2 signal as an additive noise. To maximize the SINR at the receiver output while also taking into account the intracell and intercell interference by modeling it as chip rate cyclostationary noise, the filter f can be adapted to be the Max-SINR filter [1].

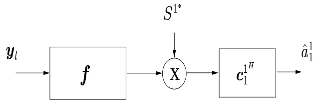

Figure 2: Filter-correlator receiver structure

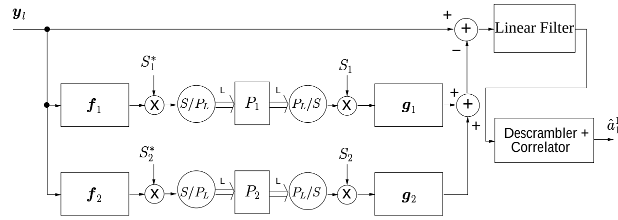

Figure 3: Hybrid structure

2.2. Two-stage Cascade Receiver Structures

Fig. 3 shows a two-stage cascade receiver structure where an Interference Cancellation (IC) block which is composed of two time-variant IC branches precedes a linear receiver. Each branch in the IC block is formed in order by an equalizer, a descrambler, a serial to parallel converter, a projector, a parallel to serial converter, a scrambler and a linear filter. Exploiting the structure of the received signal, this approach is conceived to partially perform both intracell and intercell interference cancellation to improve the performance of the subsequent simple linear receivers. Using only one of the branches and other different implementations are possible.

The purpose of the processing in the first IC branch is to approximate the interference originating from base station 1 by exploiting the orthogonality between the user codes of BS 1 after the equalizer and descrambler blocks and by projecting the symbol by symbol parallelized form of this signal into the subspace of the used spreading codes, except for the user of interest. In terms of complexity, projection operation corresponds to roughly $L\,log_2 L$ additions/subtractions by using fast Walsh Hadamard transforms. By rescrambling and "re-channeling" the serialized form of projected signal, the intracell interference is approximated at the output which somewhat also partially contains the intercell interference plus noise due to the projection operation. Similarly, interference originating from base station 2 is approximated in the second IC branch with an only difference that projection is onto the complete subspace of used spreading codes of BS 2. Provided that every interfering user's signal amplitude is available by estimation in a long term like during one slot period, then the projections on the two IC branches, which might be interpreted as interference approximation with linear soft decisions, can be replaced by despreading, hard limiting (nonlinear), maximum ratio weighting by slotwise estimated signal amplitude and finally respreading operations on each interfering user as in Fig. 4. When applied to the first branch, the approximation of the signal of user of interest (user 1 here) is excluded as indicated with the dashed line in the bottom chain of Fig. 4. In this scheme, by taking the energy differences of the interfering users into account, weighted hard interference approximation is done and hence an increase in the output SINR is contemplated. However note that hard decision might result in an even worse performance unless the estimates of interfering signals are reliable enough. Therefore a compromise solution could be a tangent hyperbolic decision as shown in Fig. 5 only for the k^{th} interfering user of base station j with slotwise estimated signal magnitude α_k^j in place of the corresponding hard decision block element in Fig. 4, see [4].

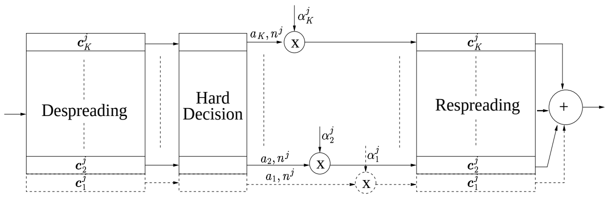

Figure 4: Hard decision scheme in place of projection

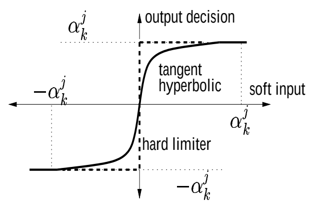

Figure 5: Tangent hyperbolic and hard limiter decisions

There are various filtering alternatives. First of all, the equalizers can either be of Zero-Forcing (ZF) or Minimum Mean Square Error (MMSE) types. The former type of an equalizer enhances the intercell interference plus noise alot but preserves the code orthogonality, whereas the latter type perturbs the structure of the signal received from the base station but does not increase so much the intercell interference plus noise. Note that although it is not possible to completely equalize a single FIR channel by an FIR equalizer, it is possible with multichannels [2]. In this case, multichannels exist due to oversampling. Secondly, the "re-channeling" filters can either be the channels $\boldsymbol{h}^j$ themselves or Wiener (LMMSE) filters adapted for minimizing the noise. In this context, we know that MMSE equalizers outperform ZF-equalizers and when MMSE equalizers are used Wiener filters are practically not necessary [3]. Thirdly, the linear receiver after the subtractor can either be a RAKE receiver, an MMSE-ZF equalizer or a Max-SINR (unbiased MMSE) filter. Rake receiver and MMSE-ZF equalizer are the two extreme approaches such that the former assumes the simultaneous cancellation of all intracell and intercell structured interference and considers the remaining signal as pure AWGN whereas the latter considers that the remaining signal has the same statistics as the received signal. However MMSE-ZF equalizer can be constructed in a better way by roughly estimating/ approximating the spillover intracell/intercell interference power and noise, exploiting every possible side information such as the number of active users in the current and neighboring cells and their respective overall power profiles.

3. SIMULATIONS AND RESULTS

In this section we present some of the simulations performed to evaluate the various structures. The K^j users of base station j are considered synchronous between them, with the same spreading factor $L = 32$ and using the same downlink channel $\boldsymbol{h}^j$ which is a FIR filter, convolution of a sparse Vehicular A UMTS channel and a root-raised cosine pulse shape with roll-off factor of 0.22. The channel length is $N = 19$ chips due to the UMTS chip rate of 3.84 Mchips/sec. An oversampling factor of $m = 2$ and one receive antenna $q = 1$ are used. Two possible user power distributions are simulated: The user of interest has either the same power of average single interferer power or 10dB less which represents the near-far situation.

Fig. 6 to Fig. 9 show the performances of linear and cascade receivers. On the figures, the terms S1, S2 and S3 refer to the usage of branch1, branch2 and both branches; ch1, ch2 and H refer to the rechannels $\boldsymbol{h}^1$, $\boldsymbol{h}^2$ and both; Rake, MS and LMMSE refer to the linear filters after the subtraction and finally HardDec and THyp refer to the usage of hard-limiting and tangent hyperbolic decision schemes instead of projections. Generalizing for all these simulation results, all the receivers are sensitive to higher loading fractions and near-far cases. In all the cases, Rake receiver performs the worst. Contrary to our expectations, S1-ch1-Rake does not improve the performance much; however S2-ch2-Rake comparably improves the performance and S2-ch2-MS is even better. MS receiver performs in between S1-ch1-Rake and S2-ch2-MS and the gap between S2-ch2-MS and MS closes as the interference level increases (higher loading fraction and/or near far situation). As we have expected, using both branches increases the performance and as for the usage of linear filters after the subtractor for this case, we see that MMSE equalizers outperform Rake receivers. We see that there are remarkable performance improvements when the nonlinear hard-limiting or tangent hyperbolic decision structures are used instead of projections and they are more robust to higher loading fractions and near-far situations. Comparing the two, tangent hyperbolic decision gives slightly better performance than hard-limiter at moderate to high SNR values and is levelled only at very high SNR regions.

4. CONCLUSIONS

In the downlink of CDMA-based cellular mobile systems, the classical RAKE receiver exhibits significant limitations in the existence of strong interfering users which is always the case when perfect beamforming is not applied or there is another strong interfering base station. In this paper we proposed two-stage reasonably complex receiver structures which allow important performance improvements by considering noise cancellation in the direction of unused codes.

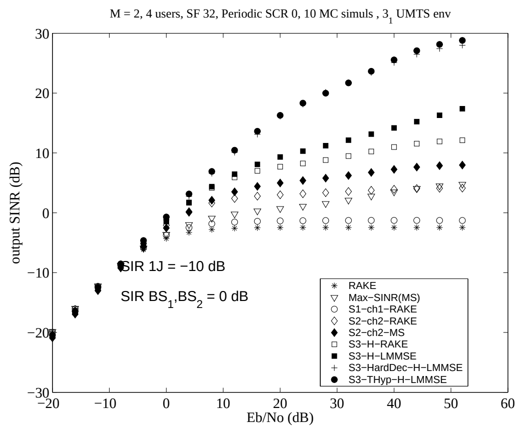

Figure 6: Output SINR versus Eb/No, 12.5% loaded BSs, spreading factor 32 and near-far situation, aperiodic scrambling, MMSE equalizers

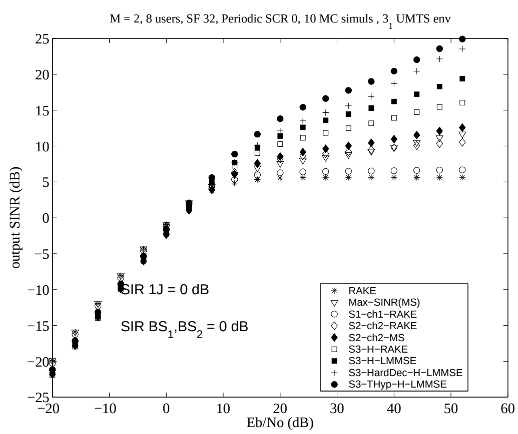

Figure 7: Output SINR versus Eb/No, 25% loaded BSs, spreading factor 32 and equal power distribution, aperiodic scrambling, MMSE equalizers

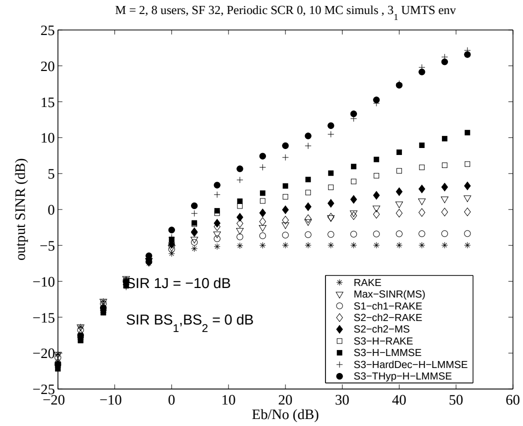

Figure 8: Output SINR versus Eb/No, 25% loaded BSs, spreading factor 32 and near-far situation, aperiodic scrambling, MMSE equalizers

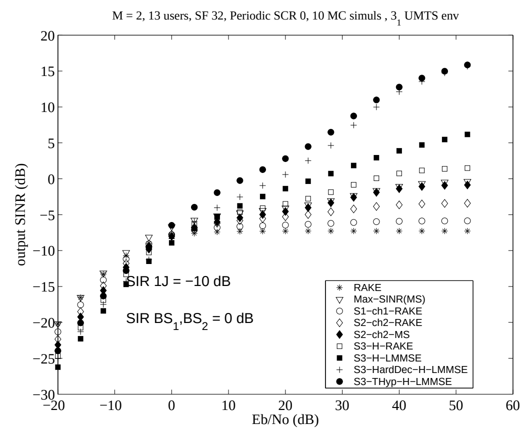

Figure 9: Output SINR versus Eb/No, 40% loaded BSs, spreading factor 32 and near-far situation, aperiodic scrambling, MMSE equalizers

5. REFERENCES

[1] M. Lenardi and D.T.M. Slock, "A rake receiver with intracell interference cancellation for a ds-cdma synchronous downlink with orthogonal codes," *Proc. VTC 2000, Tokyo, Japan*, May 2000.

[2] C. Papadias and D.T.M. Slock, "Fractionally spaced equalization of linear polyphase channels and related blind techniques based on multichannel linear prediction," *IEEE Transactions on Signal Processing, Vol.47, No.3*, March 1999.

[3] M. Lenardi and D.T.M. Slock, "Downlink intercell interference cancellation in wcdma by exploiting excess codes," *Proc. SAM 2000, Boston, Massachusset - USA*, March 2000.

[4] A. Nahler R. Irmer and G. Fettweis, "On the impact of soft decision functions on the performance of multi-stage parallel interference cancelers for cdma systems," *Proc. VTC Spring 2001, Rhodes, Greece*, May 2001.