A Low Complexity Turbo Adaptive Interference Cancellation Using Antenna Arrays for W-CDMA

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Abstract - **This paper investigates an adaptive iterative space-time interference cancellation receiver for Wideband Code Division Multiple Access (W-CDMA). Each iteration of the interference cancellation includes adaptive beamforming, adaptive path-combining and soft-input soft-output decoding. The complexity of the demodulation process using very simple Normalized Least Mean Square (N-LMS) adaptive algorithms is limited with no need of space-time channel estimation. Adaptive space-time processing algorithms are shown to benefit from the iterative process. Simulation results on a UMTS-like scenario show that this receiver achieves performance that is very close to the single user bound, even for highly interfered networks.**

I. INTRODUCTION

Third generation mobile radio systems (IMT-2000, UMTS) aim at allowing the transmission of data rates up to 2 Mbit/s by using a physical layer based on the W-CDMA technology. In a first step, simple receiver techniques such as Rake receivers will probably be employed to cope with the relatively low cell load coming from 3rd generation first users. In a second step, to face an increase of traffic, *i.e.*, to deal with highly interfered networks, advanced receiver techniques will have to be implemented, namely in Base Stations, where additional computational complexity will be tolerable. Thus, extensive research has been done on interference mitigation techniques for W-CDMA. Among solutions, those based on interference cancellation (IC) [1] and beamforming [2] are likely to be implemented in Base Stations for the UMTS Frequency Division Duplex mode. During the last few years, several studies and research projects have focused on space-time interference cancellation techniques [3]-[5]. Iterative techniques using the channel decoding process, often called Turbo techniques, are proposed in [6] to improve the performance of space-time interference cancellation, assuming perfect space-time channel estimation. In [7], actual space-time channel estimation is included in each iteration to make it also benefit from the iterative process. Preliminary results on iterative space-time interference cancellation comparing the performance of explicit space-time channel estimation and adaptive space-time signal processing are presented in [8]. Thus, including more and more signal-processing functions in the iterative process always improves the transmission performance. However, it increases the receiver complexity, which might also become a limitation for an implementation at the Base Station.

This paper proposes an adaptive iterative space-time interference cancellation technique, which aims at ensuring the good trade-off between performance and complexity by avoiding space-time channel estimation and its related difficulties (*e.g.*, the necessity to know the number of impinging signals for the root-MUSIC algorithm with spatial smoothing). Further advantages of using adaptive signal processing techniques are their capacity to track short term space-time channel variations and the insensitivity of adaptive beamforming to phase array calibration mismatches. In the sequel, the general principle of space-time interference cancellation is presented in section II, whereas the specificity of the proposed adaptive implementation is detailed in section III. Simulation results showing the convergence of the algorithm and the performance in terms of bit-error-rate (BER) for different numbers of antennas are proposed in section IV for a highly loaded UMTS-like scenario. Concluding remarks are finally given in section V.

II. ITERATIVE SPACE-TIME INTERFERENCE CANCELLATION

The general transmitter structure of a Direct Sequence CDMA (DS-CDMA) reverse link system is represented in Fig. 1 for K users.

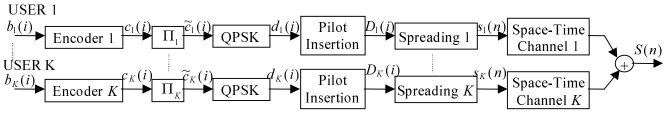

Fig. 1: Structure of DS-CDMA reverse link transmitters.

Information bits $b_k(i)$ of user k are convolutionally encoded. The trellis of each convolutional code is properly terminated to divide the data stream into finite blocks of N_c coded bits $c_k(i)$, $i = 0, \ldots, N_c - 1$. These coded blocks are then interleaved by a user-specific interleaver Π_k to guaranty independence between the different users' codes if a very

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low spreading factor is assigned. The interleaved coded bits $\tilde{c}_k(i)$ are then mapped onto $N_c / 2$ QPSK data symbols $d_k(i)$ and N_p QPSK pilot symbols $p_k(i)$ are inserted as header. The obtained sequence, with $N_D = N_c / 2 + N_p$ symbols, $D_k(0), \ldots, D_k(N_D - 1)$, is then spread by a user-specific signature. For each code block and for a spreading factor SF, $N_{chips} = SF.N_D$ chips $s_k(n)$ are sent on user k space-time channel. Each user's channel has P paths, each path p having a delay equal to $\tau_{k,p}$ chips and a complex Gaussian coefficient $h_{k,p} = \rho_{k,p}\exp(j\,\nu_{k,p})$.

At the receiver side, each path arrives from a direction of arrival (DOA) $\theta_{k,p}$ on a Uniform Linear Array (ULA) with L antennas, as depicted in Fig. 2. The ULA geometry induces a constant additional propagation length from one antenna to the next, equal to $d.\cos(\theta_{k,p})$ where d is the distance between adjacent antennas. Thus, for DOA $\theta_{k,p}$, the space-time channel coefficient resulting from the specific phase rotation $\varphi_{k,\ell,p}$ on antenna ℓ is:

$$h_{k,\ell,p} = \rho_{k,p} e^{j\xi_{k,\ell,p}} \tag{1}$$

where

$$\begin{aligned}\xi_{k,\ell,p} &= \nu_{k,p} + \varphi_{k,\ell,p} = \nu_{k,p} + (\ell - 1).\varphi_{k,p} \\ &= \nu_{k,p} + 2\pi \frac{d}{\lambda}(\ell - 1)\cos(\theta_{k,p})\end{aligned} \tag{2}$$

where λ is the carrier wavelength and $\varphi_{k,p}$ is the constant phase rotation between two consecutive antennas.

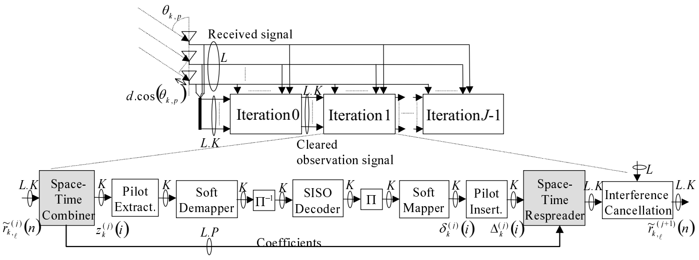

Fig. 2: Structure of the iterative space-time interference cancellation receiver.

The received signal on each antenna is:

$$r_\ell(n) = \sum_{k=1}^{K}\sum_{p=1}^{P} h_{k,\ell,p} s_k(n - \tau_{k,p}) + w_\ell(n) \tag{3}$$

where $w_\ell(n)$ is the additive white Gaussian noise (AWGN) on antenna ℓ at time n.

The basis of proposed DS-CDMA reverse link receiver structure is composed of J iterations including antenna processing and soft decoding as originally presented in [6]. In each iteration j and for each user k, a vector of L antenna observation signals $\tilde{r}_{k,\ell}^{(j)}(n)$ ($\ell = 1,\ldots,L$) issued from the previous iteration $j - 1$, is processed by a user-specific space-time combiner. In the first iteration, this vector is equal to the received signal vector for all users, *i.e.*,

$$\tilde{r}_{k,\ell}^{(0)}(n) = r_\ell(n)\,,\ k = 1,\ldots,K \tag{4}$$

On the basis of pilot symbols, the space-time combiner ensures a despreading of each path on each antenna, a beamforming for each path and a path combining in order to generate the symbol estimate $z_k^{(j)}(i)$ of transmitted symbol $D_k(i)$. After pilot extraction and soft de-mapping, this estimate is sent to the de-interleaver followed by a Soft-Input Soft-Output (SISO) decoder, which can provide soft estimates on both the information bits and the coded bits with an increase of reliability. The coded bits are interleaved again and mapped into soft symbols $\delta_k^{(j)}(i)$. Pilot symbols are inserted to form the estimated sequence of transmitted symbols $\Delta_k^{(j)}(i)$:

$$\begin{aligned}&\Delta_k^{(j)}(i) = p_k(i) && \text{for } i = 0,\ldots,N_p - 1 \\ &\Delta_k^{(j)}(i) = \delta_k^{(j)}(i - N_p) && \text{for } i = N_p,\ldots,N_D - 1\end{aligned} \tag{5}$$

For each user k, this sequence is processed by a so-called space-time respreader that generates the summed contribution of the P paths on each antenna with proper channel attenuation and phase rotation. Interference cancellation finally consists in subtracting on each antenna the K - 1 contributions of interfering users from the received signal in order to yield a cleared observation signal $\tilde{r}_{k,\ell}^{(j+1)}(n)$ for each user with less interference. As some interference may remain, this process is iterated so that the observation signal for each user approaches an interference-free signal.

Performance of the space-time interference cancellation structure was proposed in [6] in the case of perfect space-time channel estimates. As in a real transmission scenario, this performance strongly depends on the estimation reliability, an explicit space-time channel estimation was included in the iterative process as proposed in [7]. Alternatively, we propose in this paper a turbo (or iterative) adaptive space-time interference cancellation that does not need any explicit space-time channel estimation.

III. Adaptive Combining and Respreading

In the iterative space-time interference cancellation receiver represented on Fig. 2, both the space-time combiner and the space-time respreader require some knowledge about space-time channel characteristics. Thus, avoiding space-time channel estimation thanks to adaptive signal processing impacts both the combining and the respreading techniques.

A. Adaptive Space-Time Combining

We include in each iteration an adaptive minimum mean square error (MMSE) space-time combiner. The beamformer may thus be able to cancel interfering signals by placing nulls in their direction while pointing its main lobe towards the direction of the desired signal. By contrast with beamforming based on explicit DOA estimation algorithms, *e.g.*, the root-MUSIC algorithm with spatial smoothing, adaptive MMSE beamforming can form several lobes towards the different DOAs of an observation signal without any a priori knowledge of the number of DOAs. To ensure a good trade-off between performance and complexity, we choose the Normalized-Least Mean Square (N-LMS) adaptive algorithm [9]. Concatenated and joint structures have been investigated to perform space-time combining [10]. We focus on a disjoint adaptive structure, *i.e.*, an adaptive N-LMS beamforming filter followed by an adaptive N-LMS path combining filter, since dealing with two short filters instead of one larger (for the joint approach) brings better convergence speed properties. The resulting structure of the proposed space-time combiner is represented on Fig. 3.

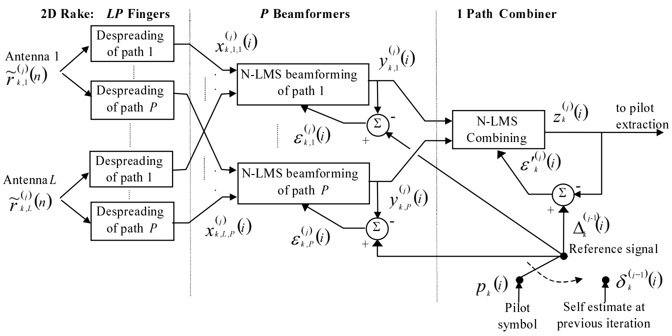

Fig. 3: Structure of the adaptive MMSE space-time combiner.

In a first step, for each user k, the despreading of each path p on each antenna ℓ by a filter matched to the path delay $\tau_{k,p}$ generates samples $x_{k,\ell,p}^{(j)}(i)$ for $i = 0,\ldots, N_D - 1$.

In a second step, a beamforming is processed for each path to mitigate the remaining interference from other directions than those of the desired signal. Since despread samples are narrow-banded, the beamforming uses a single coefficient $\beta_{k,\ell,p}$ per antenna. Denoting $\overline{x}_{k,p}^{(j)}(i)^T = \left(x_{k,1,p}^{(j)}(i),\ldots,x_{k,L,p}^{(j)}(i)\right)$,

$$y_{k,p}^{(j)}(i) = \overline{\beta}_{k,p}^{(j)}(i)^T \cdot \overline{x}_{k,p}^{(j)}(i) = \sum_{\ell=1}^{L} \beta_{k,\ell,p}^{(j)}(i) \cdot x_{k,\ell,p}^{(j)}(i) \tag{6}$$

where $\beta_{k,p}^{(j)}(i)^T = \left(\beta_{k,1,p}^{(j)}(i),\ldots,\beta_{k,L,p}^{(j)}(i)\right)$ results from the N-LMS update rule as follows:

$$\overline{\beta}_{k,p}^{(j)}(i+1) = \overline{\beta}_{k,p}^{(j)}(i) + \mu_{k,p}^{(j)}(i) \cdot \varepsilon_{k,p}^{(j)}(i) \cdot \overline{x}_{k,p}^{(j)*}(i) \tag{7}$$

$\varepsilon_{k,p}^{(j)}(i)$ is the error signal controlling the convergence of the algorithm and $\mu_{k,p}^{(j)}(i)$ is the step size of the N-LMS algorithm [9]. Initialization of vector $\overline{\beta}_{k,p}^{(0)}(0)$ ensures an omnidirectional beamforming, *i.e.*, $\overline{\beta}_{k,p}^{(0)}(0) = (1,0,0,\ldots,0)$. A classical way [11] to generate the error signal $\varepsilon_{k,p}^{(j)}(i)$ consists in considering a reference signal that is passed through a time channel model to take into account the complex channel coefficient since the degradations caused by the multipath channel have not been compensated yet. However, as this method would require explicit knowledge of the complex channel attenuation, we prefer the following error signal definition to avoid channel estimation:

$$\varepsilon_{k,p}^{(j)}(i) = \Delta_k^{(j-1)}(i) - y_{k,p}^{(j)}(i) \tag{8}$$

Note that in iteration 0, $\Delta_k^{(-1)}(i)$ is only composed of pilot symbols $p_k(i)$ since soft estimates $\delta_k^{(-1)}(i)$ are not available. Therefore, adaptive processes of the first iteration can only be based on N_p observation signals. This first filtering process can be considered as a "rotated beamforming" since applying (7) in (6) forces the beamformer to compensate the channel phase rotation $\nu_{k,p}(i)$.

In a third step, observations for all paths of user k are combined to provide us with a single observation $z_k^{(j)}(i)$ for each transmitted symbol $D_k(i)$:

$$z_k^{(j)}(i) = \overline{\alpha}_k^{(j)}(i)^T \cdot \overline{y}_k^{(j)}(i) = \sum_{p=1}^{P} \alpha_{k,p}^{(j)}(i) \cdot y_{k,p}^{(j)}(i) \tag{9}$$

where $\overline{y}_k^{(j)}(i)^T = \left(y_{k,1}^{(j)}(i),\ldots,y_{k,P}^{(j)}(i)\right)$ and the combining coefficient vector $\overline{\alpha}_k^{(j)}(i)^T = \left(\alpha_{k,1}^{(j)}(i),\ldots,\alpha_{k,P}^{(j)}(i)\right)$ is generated using another N-LMS update rule as follows:

$$\overline{\alpha}_k^{(j)}(i+1) = \overline{\alpha}_k^{(j)}(i) + \mu_k'^{(j)}(i) \cdot \varepsilon_k'^{(j)}(i) \cdot \overline{y}_k^{(j)*}(i) \tag{10}$$

where $\varepsilon'^{(j)}_k(i)$ is the error signal leading the convergence of the algorithm and $\mu'^{(j)}_k(i)$ is the step size of the N-LMS algorithm. Initialization of vector $\overline{\alpha}_k^{(0)}(0)$ ensures an equal gain path-combining, *i.e.*, $\overline{\alpha}_k^{(0)}(0) = (1,1,\ldots,1)$. The error signal $\varepsilon'^{(j)}_k(i)$ is generated as:

$$\varepsilon_k'^{(j)}(i) = \Delta_k^{(j-1)}(i) - z_k^{(j)}(i) \tag{11}$$

Using the same reference signal in equations (8) and (11) avoids estimating the channel and enables to jointly optimize the space and time combining filters. Indeed, the beamforming provides us with samples $y_{k,p}^{(j)}(i)$ that have received an initial phase correction and the path combining finalizes this correction. Particularly when the structure has converged, beamforming fully compensates the channel phase rotation and path combining only weights each path with a distinct real-valued coefficient.

Besides, channel decoding induces a block-by-block processing of received signals. Thus, at the end of the processing of a given block at iteration j, vectors $\overline{\alpha}_k^{(j)}(N_D - 1)$ and $\overline{\beta}_{k,p}^{(j)}(N_D - 1)$ are assumed to be closer to the MMSE optimum vectors. Hence, both final sets of coefficients are re-used to process the same block again, so that the output samples $z_k^{(j)}(i)$ can be generated with additional reliability.

This adaptive space-time combiner benefits from the iterative structure in two ways. On one hand, the reference signal $\Delta_k^{(j-1)}(i)$ can be either a pilot symbol or a self-estimate, which becomes more reliable with the number of iterations. On the other hand, the initialization of adaptive filters for one block at iteration j ($j \neq 0$) is based on the filter coefficients obtained after the processing of the same block at iteration j - 1, since these coefficients are expected to be all the closer to the optimum MMSE coefficients as the number of iterations increases. Hence, including adaptive combining in each iteration is expected to improve the performance.

B. Space-Time Respreading

As depicted on Fig. 2, for each iteration, the interference contribution of each user must be regenerated at each antenna connector in order to subtract the space-time interference from the received signal. Thus, the signal estimate $\Delta_k^{(j)}(i)$ of each user must be respread by the associated spreading sequence and filtered by a model of the space-time channel, as represented on Fig. 4.

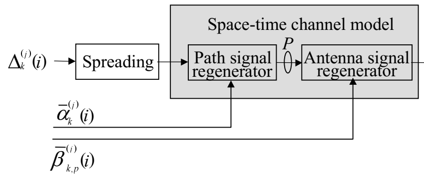

Fig. 4: Structure of the space-time respreader.

In order to avoid any explicit estimation of the space-time channel, the interference regeneration process simply uses coefficients issued from the adaptive space-time combining filters. However, since the reference signal is common to both adaptive algorithms, vectors $\overline{\beta}_{k,p}^{(j)}(i)$ and $\overline{\alpha}_k^{(j)}(i)$ do not explicitly contain the separated influence of the multipath channel and of the DOAs. Therefore, we model the space-time channel in two parts: a path signal regenerator and an antenna signal regenerator. The vector of P coefficients $\overline{\gamma}_k^{(j)}(i)$ used to perform path signal regeneration is derived from the vector of P coefficients $\overline{\alpha}_k^{(j)}(i)$ issued from the path-combining filter, which is assumed to converge to maximum ratio combining (MRC):

$$\overline{\gamma}_k^{(j)}(i) = \overline{\alpha}_k^{(j)*}(i) \qquad (12)$$

Similarly, the vector of coefficients $\overline{\zeta}_{k,p}^{(j)}(i)$ for antenna signal generation relies on vector $\overline{\beta}_{k,p}^{(j)}(i)$ issued from the adaptive beamformer. Assuming that the phase of $\overline{\beta}_{k,p}^{(j)}(i)$ elements gives a modified DOA as seen by the beamformer (including the influence of the channel phase rotation), the elements $\zeta_{k,\ell,p}^{(j)}(i)$ of vector $\overline{\zeta}_{k,p}^{(j)}(i)$ are defined as:

$$\zeta_{k,\ell,p}^{(j)}(i) = \frac{\beta_{k,\ell,p}^{(j)}{}^{*}(i)}{\left|\beta_{k,\ell,p}^{(j)}(i)\right|} \qquad (13)$$

This regeneration process assumes that the adaptive MMSE space-time combining structure converges to a space-time matched filter as the number of iterations increases. If interference cancellation performs successfully and additive noise is omnidirectional, this condition is validated for an observation signal coming from a single DOA, since MRC and MMSE beamforners have the same antenna diagram. For an observation signal issued from multiple DOAs, this approximation may lead to slight degradations.

IV. SIMULATION RESULTS

The performance of the turbo adaptive space-time interference cancellation is analyzed in terms of Mean Square Error (MSE) and BER on a UMTS-like highly interfered scenario. We consider the reverse link transmission of the signals of 16 users, each transmitting with the same power 328 kbps on 2 distinct paths using a spreading factor equal to 8 on a 3.84 MHz bandwidth channel. For each user, blocks of 328 information bits and 8 tail bits are coded by a (561,753) rate ½ convolutional code. N_p = 144 QPSK pilot symbols are added to the obtained $N_c/2$ = 336 QPSK data symbols to form blocks of 480 symbols transmitted in 1 ms. Thus, 30% of the transmitted energy is dedicated to pilots. The space-time channel is assumed stationary over each block. Every new block, for each delay, a channel complex coefficient and a single DOA (within a 120 degrees sector) are randomly chosen, which can be considered as a poor case for adaptive systems. 3 to 8 antennas compose the ULA. Perfect power control is assumed. In order to initiate the convergence of the space-time combining filters, the two first iterations only use pilot symbols with large step sizes. As the iteration number increases, smaller step sizes are employed with soft decoded bits from the previous iteration as pilots. For these results, we use Eb/N0 values that are normalized with respect to the number of antennas so that a fair comparison between different ULA configurations is possible for the same Eb/N0.

For Eb/N0 = 5 dB, Fig. 5 represents the influence of both the number of antennas and the number of iterations on the MSE between the symbol estimates $\delta_k^{(j)}(i)$ and the transmitted symbol $d_k(i)$ computed over all the users and averaged over 100 blocks of N_D - N_p symbols corresponding to 100 different multi-user space-time channel configurations. As a lower bound, we consider the MSE, which is achieved for a single user transmission with the same Eb/N0 on AWGN channel. For any number of antennas, we first observe that one initial detection stage (iteration 0) is not sufficient to output blocks of symbols with a low MSE and that adaptive space-time interference cancellation can benefit from the iterative process. For 5 antennas, the MSE progressively falls from –2.3 dB down to –4.3 dB after 4 iterations (iterations 0 to 3), which is only a 0.7 dB loss compared to the lower bound. This loss remains almost constant after the fourth iteration showing that 4 iterations with 5 antennas are quite enough to deal with such a highly interfered system. Reducing to 3 the number of antennas significantly degrades the performance of the receiver, whereas 4 antennas is an appropriate configuration to benefit from the whole gain provided by the 5 iterations. Finally, this figure emphasizes the complexity trade-off between having less iterations with more antennas and having more iterations with less antennas since 3 iterations with 8 antennas perform just as 5 iterations with 4 antennas.

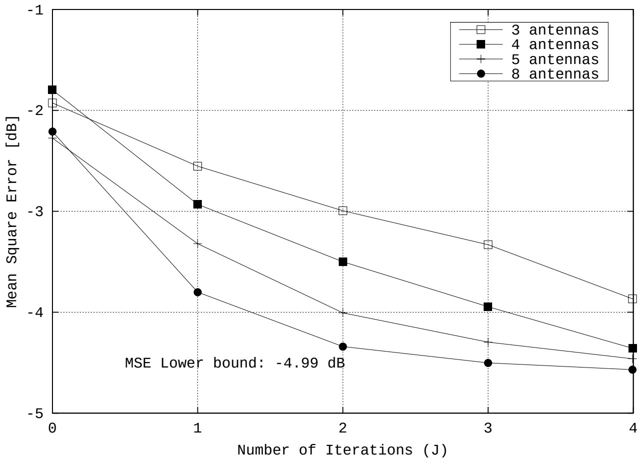

Fig. 5: Influence of the number of antennas on the convergence of the MSE (normalized Eb/N0 = 5 dB).

On Fig. 6, the BER averaged over the 16 users is plotted for the same transmission scenario with 4 antennas. Iteration 0 performs worse than a 2D Rake receiver with perfect space-time channel estimation, which can be considered as the best achievable performance of first implemented UMTS Base Stations. In this case, both N-LMS algorithms have not properly converged to the optimum MMSE solution because of a lack of observations. However, as long as the number of iterations increases, performance of the adaptive space-time interference cancellation is improved. Thus, with only 4 iterations and 4 antennas, the proposed receiver experiences a 0.3 dB loss to achieve a BER equal to 10^{-4}, as compared with the single user reference.

V. Conclusion

Compared to Rake receivers, iterative space-time interference cancellation techniques are good candidates to improve the capacity of the W-CDMA systems. As a good trade-off between performance and complexity, we proposed an adaptive solution that avoids space-time channel estimation. On a high data rate UMTS-like scenario, simulation results showed that adaptive space-time signal processing based on simple N-LMS algorithms can really benefit from an iterative process and achieve near single user performance, even in very high loaded system and for a low number of antennas.

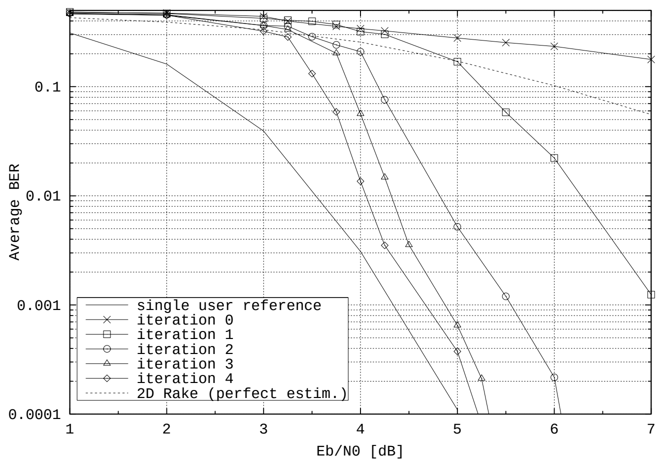

Fig. 6: Influence of the number of iterations on the BER performance (normalized Eb/N0, 4 antennas).

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