

Iterative Space-Time Interference Cancellation for W-CDMA Systems

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Abstract: *Iterative multiuser interference cancellation schemes for Direct Sequence Code Division Multiple Access systems exhibit good performance results for a reasonable complexity. We study an efficient iterative scheme, combining interference cancellation, soft input soft output decoding and beamforming. To deal with unknown channels, we follow two approaches: on one hand the addition of a pilot-aided space-time channel estimation in each iteration and on the other hand the use of adaptive beamforming and path combining. The iterative structure in both approaches has two advantages: the observation signal used for estimation or adaptation contains less interference from one iteration to another and soft estimates of coded bits are available for data-aided estimation or adaptation.*

1. Introduction

The air interface of the third generation Universal Mobile Telecommunication System (UMTS), is based on Wideband Code Division Multiple Access (W-CDMA). Even if both Frequency Division Duplex (FDD) and Time Division Duplex are considered to deal with the available paired and unpaired frequency bands, the FDD mode currently receives more interest from European manufacturers. At first, simple receivers will probably be implemented to face the presumed relatively low traffic coming from 3G first users. Subsequently, advanced receivers will have to be employed to deal with the expected increase of traffic and associated interference. Therefore, many interference mitigation techniques have been studied for W-CDMA receivers, namely for Base Stations where additional computational complexity is tolerable. Due to the large size of spreading sequences in FDD mode, techniques based on Interference Cancellation (IC) [1] seem to be appropriate to mitigate intra-cell interference with a reasonable complexity increase. The interference on a given user is rebuilt from a bank of Rake receivers and subtracted from the received signal to produce a new clearer observation for this user. This process may then be iterated. Several studies, *e.g.* [2][3], have shown that a performance increase could be obtained by using channel coding and performing Soft Input Soft Output (SISO) decoding inside each detection iteration. Alternatively or in conjunction, beamforming antenna arrays may reduce this interference [4]. Recently, several studies have been carried out on the association of IC techniques and antenna arrays for space-time IC [5]. An efficient scheme, combining interference cancellation, SISO decoding and beamforming, has been proposed in [6] assuming perfect channel knowledge. In each iteration, for each user and each path, a conventional beamforming, linearly combining signals from all antennas, allows additional interference reduction, thus improving the iterative process.

Following the same trend as in [6], which leads to the inclusion of the maximum number of signal processing functions in the iterative process to further improve the transmission performance, this paper proposes and compares two different iterative space-time interference cancellation techniques for unknown channels. The first one integrates an explicit space-time channel estimation [7], which is renewed at each iteration to make it also benefit from the iterative process. The second technique deals with an adaptive iterative space-time interference cancellation technique [8] that aims at achieving a good trade-off between performance and complexity. These studies have been realised in the scope of the IST ASILUM project¹ [9], in the part dedicated to the optimisation of interference mitigation techniques in order to improve the capacity of UMTS mobile cellular systems.

The paper is organised as follows: In section 2, the principle of an iterative space-time interference cancellation is presented. The space-time channel estimation is then inserted in the iterative process in section 3. The description of the adaptive approach follows in section 4. Performance results in section 5 allow comparison between both approaches and show that the proposed detection scheme almost achieves single-user performance even with the highly loaded simulated system.

2. Iterative Space-Time Interference Cancellation

Let us consider a Direct Sequence CDMA system with K users as depicted on Fig. 1. Information bits $b_k(i)$ of user k are convolutionally encoded. The trellis of each convolutional code is properly terminated in order to divide the

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data stream into finite blocks of N_c coded bits $c_k(i)$, $i = 0, \dots, N_c - 1$. These coded blocks are then interleaved by a user-specific interleaver Π_k to guaranty independence between the different users' codes if a very low spreading factor is assigned. The obtained interleaved coded bits $\tilde{c}_k(i)$ are then mapped onto $N_c / 2$ QPSK symbols $d_k(i)$. N_p QPSK pilot symbols $p_k(i)$ are inserted before this data sequence. The obtained sequence, with $N_D = N_c / 2 + N_p$ symbols, $D_k(0), \dots, D_k(N_D - 1)$, is finally spread by a user-specific signature. For each code block and for a spreading factor SF , $N_{chips} = SF \cdot N_D$ 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, a complex Gaussian coefficient $h_{k,p} = \rho_{k,p} \exp(j\nu_{k,p})$ and a direction of arrival (DOA) $\theta_{k,p}$.

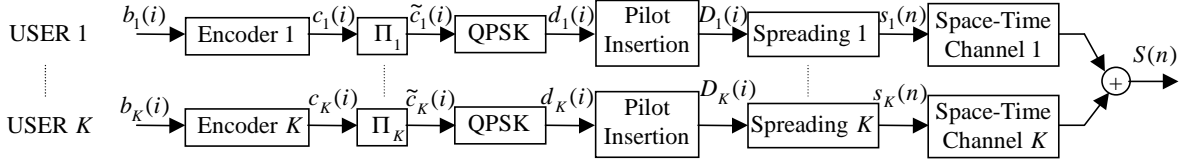


Fig. 1: Structure of a Direct Sequence CDMA transmitter.

Fig. 2 depicts an iterative receiver with J iterations using a Uniform Linear Array (ULA) with L antennas and a SISO decoding, as described in [6]. The ULA geometry induces a constant additional propagation length from one antenna to the next, equal to $d \cdot \cos(\theta_{k,p})$, where d is the distance between 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

$$h_{k,\ell,p} = \rho_{k,p} e^{j\xi_{k,\ell,p}}, \text{ where } \xi_{k,\ell,p} = \nu_{k,p} + \varphi_{k,\ell,p} = \nu_{k,p} + (\ell - 1) \cdot \varphi_{k,p} = \nu_{k,p} + 2\pi \frac{d}{\lambda} (\ell - 1) \cos(\theta_{k,p}) \quad (1)$$

where λ is the wavelength and $\varphi_{k,p}$ is the constant phase rotation between two consecutive antennas. The received signal on each antenna is

$$r_\ell(n) = \sum_{k=1}^K \sum_{p=1}^P h_{k,\ell,p} s_k(n - \tau_{k,p}) + w_\ell(n) \quad (2)$$

where $w_\ell(n)$ is the Additive White Gaussian Noise (AWGN) on antenna ℓ at time n .

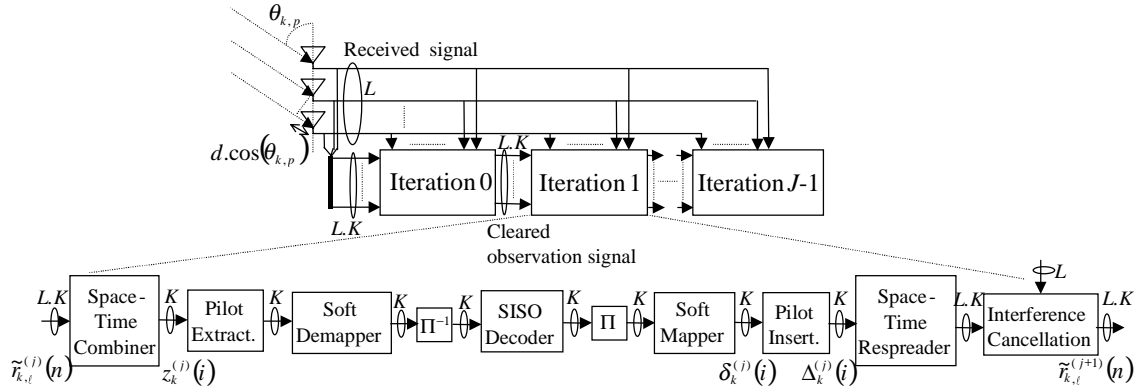


Fig. 2: Structure of the iterative space-time receiver.

In each receiver iteration j , a user-specific multi-antenna observation signal $\tilde{r}_{k,\ell}^{(j)}(n)$, where $\ell = 1, \dots, L$, provided by the previous iteration $j - 1$, is processed by a so-called 2D-Rake receiver or space-time combiner. In the first iteration, this observation signal is equal for all users to the received signal:

$$\tilde{r}_{k,\ell}^{(0)}(n) = r_\ell(n), k = 1, \dots, K \quad (3)$$

The space-time combiner is detailed on Fig. 3. For each user k and each path p , the observation signal on antenna ℓ is filtered by a despreader matched to the path delay $\tau_{k,p}$. We obtain $x_{k,\ell,p}^{(j)}(i)$ for $i = 0, \dots, N_D - 1$, which mainly contains the contribution of a single path generally issued from a single DOA. To cancel residual signals in other directions, a beamforming outputs

$$y_{k,p}^{(j)}(i) = \sum_{\ell=1}^L \beta_{k,\ell,p} x_{k,\ell,p}^{(j)}(i) \quad (4)$$

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

$$z_k^{(j)}(i) = \sum_{p=1}^P \alpha_{k,p} y_{k,p}^{(j)}(i) \quad (5)$$

For each user, after pilot extraction, the observations on coded bits are deinterleaved and decoded using a SISO decoder, e.g., a forward-backward decoder. The soft estimates on coded bits are then mapped on soft QPSK data symbols $\delta_k^{(j)}(i)$ and pilot symbols are inserted to form the estimated sequence of transmitted symbols:

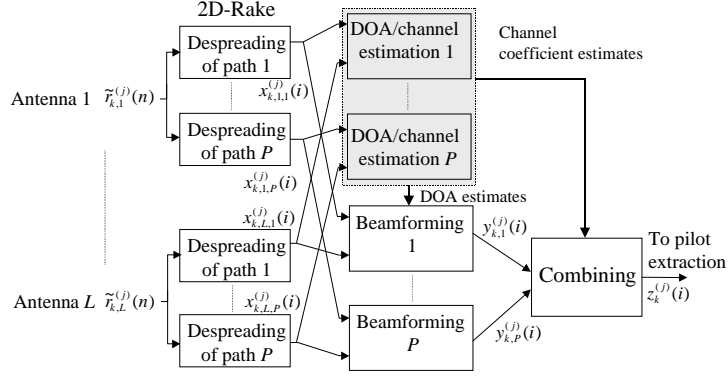


Fig. 3: Space-time combiner with channel estimation for user k .

$$\begin{aligned} \Delta_k^{(j)}(i) &= p_k(i) \quad \text{for } i = 0, \dots, N_p - 1 \\ \Delta_k^{(j)}(i) &= \delta_k^{(j)}(i - N_p) \quad \text{for } i = N_p, \dots, N_D - 1 \end{aligned} \quad (6)$$

Symbols $\Delta_k^{(j)}(i)$ are respread and each user's contribution in the global interference is rebuilt on each antenna, by modelling the space-time channel response. Using these contributions, interference cancellation is performed on the received signal to supply the following iteration with a new cleared observation signal $\tilde{r}_{k,\ell}^{(j+1)}(n)$. Using coded bit estimates after SISO decoding strongly improves the interference cancellation quality, allowing high capacity and near-single-user performance.

3. Iterative Space-Time Interference Cancellation with Channel Estimation

As channels are unknown and pilots are available, we add a pilot-aided space-time channel estimation in each space-time combiner, *i.e.*, in each iteration. Since the quality of the observation signal $\tilde{r}_{k,\ell}^{(j)}(n)$ improves from one iteration to the following, the estimation quality will also improve, thus making the detection more accurate. Hence, we expect a performance improvement compared to a non-iterative estimation, which would be performed in the first iteration only. Furthermore, since the reliability of coded bit estimates is increased thanks to SISO decoding, the soft values $\delta_k^{(j)}(i)$ of data symbols may be used as new pilots in the following iteration for data-aided estimation. This larger number of pilots for iterations 1 to $J - 1$ will further improve the estimation quality and thus the performance.

The space-time channel estimation separately computes the complex coefficient and the DOA of each path p of user k , *i.e.*, it works with signals $x_{k,\ell,p}^{(j)}(i)$ for $\ell = 1, \dots, L$ and $i = 0, \dots, N_D - 1$. Two space-time channel estimation algorithms are considered. The first one is a classical root-MUSIC with spatial smoothing [4]. Originally blind, this algorithm is modified to benefit from the knowledge of pilot symbols [7]. This algorithm has the strong advantage to distinguish between coherent paths with same delay but different DOAs. As described hereafter, a low-complexity estimator is also considered, which assumes that paths from different DOAs arrive with a different delay. It is derived from an approximation of the Maximum Likelihood (ML) criterion. From (1) and (2), assuming that interference from other users and other paths has been perfectly cancelled, we can write the despread signal as

$$x_{k,\ell,p}^{(j)}(i) = h_{k,\ell,p} D_k(i) + w'_{k,\ell,p}(i) \quad (7)$$

where for each user k and each path p , the noise samples on all antennas and all instants are Gaussian and independent. To perform ML estimation, we must minimise the quadratic Euclidean distance between the received sequence and the expected one:

$$d_{k,p}^2 = \sum_{i=0}^{N_p-1} \sum_{\ell=1}^L \left| x_{k,\ell,p}^{(j)}(i) - h_{k,\ell,p} \Delta_k^{(j-1)}(i) \right|^2 \quad (8)$$

In the first iteration, we must restrict the distance computation to the N_p pilot symbols, since no reliable data estimate is available. To simplify the minimisation, we perform it separately on each antenna to find the estimate $\hat{\xi}_{k,\ell,p}^{(j)}$:

$$\hat{\xi}_{k,\ell,p}^{(j)} = \underset{\xi}{\operatorname{argmin}} d_{k,\ell,p}^2 = \underset{\xi}{\operatorname{argmin}} \left(\sum_{i=0}^{N_p-1} \left| x_{k,\ell,p}^{(j)}(i) - \rho e^{j\xi} \Delta_k^{(j-1)}(i) \right|^2 \right) \quad (9)$$

Since the antenna array is a ULA, a simple linear regression yields the estimates $\hat{\theta}_{k,p}^{(j)}$ and $\hat{v}_{k,p}^{(j)}$ from $\hat{\xi}_{k,\ell,p}^{(j)}$, $\ell = 1, \dots, L$. Finally, $\hat{\rho}_{k,p}^{(j)}$ is obtained by minimisation of $d_{k,p}^2$.

These channel estimates are employed in a maximum ratio space-time combiner (MRC) including conventional *i.e.*, MRC beamforming ($\beta_{k,\ell,p} = \exp(-j\hat{\phi}_{k,\ell,p})$) and MRC path combining ($\alpha_{k,p} = \hat{\rho}_{k,p} \exp(-j\hat{v}_{k,p})$).

4. Adaptive Iterative Space-Time Interference Cancellation

By contrast with the previous approach, we propose here an adaptive solution for the iterative space-time interference cancellation. This technique has a low complexity since it avoids explicit channel estimation by using an adaptive space-time combining filter associated with an interference regeneration filter in each iteration.

4.1. Principles of the adaptive space-time combiner

Unlike the space-time MRC filter proposed in section 3, we consider a space-time combiner based on the mean square error (MSE) criterion. In this way, the space combiner, *i.e.*, the beamformer, takes into account the interfering signals by placing nulls in their direction while pointing its main beam towards the direction of the desired signal. By contrast with the MRC-based beamforming, minimum mean square error (MMSE) -based beamforming is also expected to deal with desired coherent paths issued from distinct DOAs since it can point distinct beams towards their directions. Furthermore, using adaptive MMSE algorithms to perform space-time combining may help tracking space-time channel variations. Finally, adaptive beamforming may also compensate potential phase array calibration mismatches.

Among the concatenated and joint structures that have been investigated to perform space-time combining [10], we focus on a disjoint structure, which is composed of an adaptive MMSE beamforming filter followed by an adaptive MMSE path combining filter. This enables to deal with two short filters (instead of one larger filter for the joint approach) that have better convergence speed properties. Besides, to keep a low complexity, we choose the simple Normalised - Least Mean Square (N-LMS) adaptive algorithm [11] for both space and time combiners.

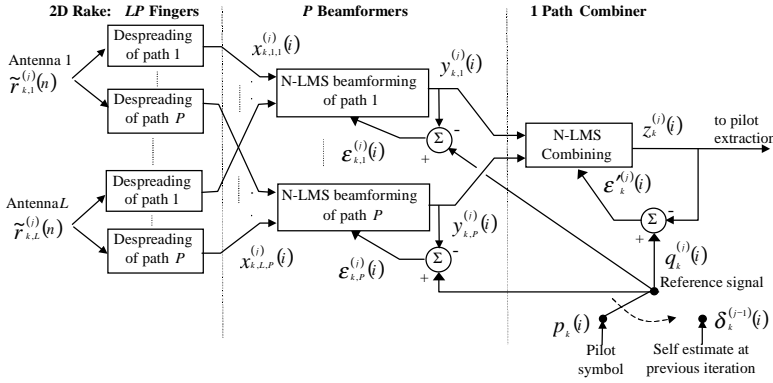


Fig. 4: Adaptive structure of the space-time filter for user k .

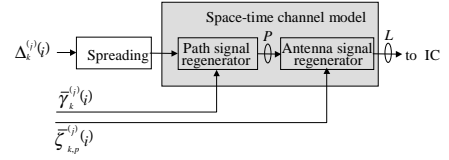


Fig. 5: Structure of the space-time re-spreader for user k .

As represented on Fig. 4, we get a user-specific adaptive space-time combining structure, which relies on a joint optimisation of beamforming and combining filters thanks to a common reference signal. The beamforming coefficients in (4) are updated according to the N-LMS algorithm, which generates vector $\bar{\beta}_{k,p}^{(j)}(i+1)$ as follows:

$$\bar{\beta}_{k,p}^{(j)}(i+1) = \bar{\beta}_{k,p}^{(j)}(i) + \mu_{k,p}^{(j)}(i) \cdot \epsilon_{k,p}^{(j)}(i) \cdot \bar{x}_{k,p}^{(j)*}(i) \quad (10)$$

where $\epsilon_{k,p}^{(j)}(i)$ 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. Initialisation of vector $\bar{\beta}_{k,p}^{(0)}(0)$ ensures an omnidirectional beamforming, *i.e.*, $\bar{\beta}_{k,p}^{(0)}(0) = (1,0,0,\dots,0)$. A classical way to generate the error signal $\epsilon_{k,p}^{(j)}(i)$ would be to consider a reference signal taking into account the complex channel coefficient [12] since this perturbation has not been corrected yet. However, as this method would require explicit channel estimation, we prefer for complexity reasons the following error signal definition:

$$\epsilon_{k,p}^{(j)}(i) = q_k^{(j)}(i) - y_{k,p}^{(j)}(i) \quad (11)$$

where $q_k^{(j)}(i)$ is a reference signal, either a pilot symbol $p_k(i)$ or a self estimate $\delta_k^{(j-1)}(i)$. Applying (11) in (10) forces the beamformer to compensate the channel phase rotation $v_{k,p}(i)$. Hence, this first filtering process can be considered as a “rotated beamforming”.

Path combining is processed adaptively for each user k and outputs the decision variable $z_k^{(j)}(i)$ thanks to the combining coefficient vector $\bar{\alpha}_k^{(j)}(i)$, which is generated using another N-LMS update rule as follows:

$$\bar{\alpha}_k^{(j)}(i+1) = \bar{\alpha}_k^{(j)}(i) + \mu_k^{(j)}(i) \cdot \epsilon_k^{(j)}(i) \cdot \bar{y}_k^{(j)*}(i) \quad (12)$$

where $\epsilon_k^{(j)}(i)$ is the error signal leading the convergence of the algorithm and $\mu_k^{(j)}(i)$ is the step size of the N-LMS algorithm. Initialisation of vector $\bar{\alpha}_k^{(0)}(0)$ ensures an equal gain path-combining, *i.e.*, $\bar{\alpha}_k^{(0)}(0) = (1,1,\dots,1)$. The error signal $\epsilon_k^{(j)}(i)$ is generated using the same reference signal $q_k^{(j)}(i)$ as in (11):

$$\epsilon_k^{(j)}(i) = q_k^{(j)}(i) - z_k^{(j)}(i) \quad (13)$$

Due to channel decoding, the received signals are processed block by block in each iteration. At the end of the processing of a given block at iteration j , vectors $\bar{\alpha}_k^{(j)}(N_D - 1)$ and $\bar{\beta}_{k,p}^{(j)}(N_D - 1)$ 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. Besides, using the same reference signal in equations (11) and (13) allows to jointly optimise 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 finalises this correction. Particularly when the structure has converged, beamforming integrally compensates the channel phase rotation and path combining only weights each path with a distinct real-valued coefficient. Finally, this adaptive space-time combiner benefits from the iterative structure in two ways. On one hand, the reference signal $q_k^{(j)}(i)$ can be either a pilot symbol or a self-estimate, which becomes more reliable with the number of iterations. On the other hand, the initialisation of the 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.

4.2. Interference regeneration filter

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. So the signal estimate of each user must be re-spread by the associated spreading sequence and filtered by a model of the space-time channel, as represented on Fig. 5. 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 $\bar{\beta}_{k,p}^{(j)}(i)$ and $\bar{\alpha}_k^{(j)}(i)$ do not explicitly contain the separated influence of the multipath channel and of the DOAs on the ULA. Therefore, we have to model the space-time channel in two parts: a path signal regenerator and an antenna signal regenerator.

The vector of P coefficients $\bar{\gamma}_k^{(j)}(i)$ used to perform path signal regeneration is derived from the vector of P coefficients $\bar{\alpha}_k^{(j)}(i)$ issued from the path-combining filter, which is assumed to converge to MRC:

$$\bar{\gamma}_k^{(j)}(i) = \bar{\alpha}_k^{(j)*}(i) \quad (14)$$

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

$$\zeta_{k,\ell,p}^{(j)}(i) = \frac{\beta_{k,\ell,p}^{(j)*}(i)}{|\beta_{k,\ell,p}^{(j)}(i)|} \quad (15)$$

This regeneration process assumes that the adaptive MMSE space-time combining structure proposed in section 4.1 converges to a space-time matched filter as the number of IC iterations increases. Indeed, if IC performs successfully and additive noise is omnidirectional, MMSE and MRC beamformers have the same antenna diagram.

5. Simulation Results

Simulation results are presented for a highly interfered system, which is taken as a first simulation step to validate the detection algorithms before testing them with real UMTS FDD uplink scenarios for the ASILUM project. The considered uplink transmission deals with the signals of 9 users, each user transmitting on a distinct 2-path channel and using a spreading factor equal to 5. A (7,5) rate $\frac{1}{2}$ convolutional channel coding is considered and a ULA with 5 antennas is used at the receiver side. Blocks of 240 QPSK symbols are transmitted, with 17 % of pilots, which the space-time channel is assumed stationary over. Channels are randomly chosen from one block to another, which can be considered as a poor case for adaptive systems. For each delay, a single DOA is randomly chosen in a 120° sector. Perfect power control is assumed, which explains why single-user performance on AWGN channel is taken as a reference. The $10\log_{10} L$ dB gain due to multiple antennas justifies the very low SNR values. To demonstrate the interference mitigation capability of the receiver, the average BER over all users is drawn versus SNR for different configurations. All users have same power. Fig. 6 shows that, with only 4 iterations, almost the whole multiuser interference is removed thanks to the efficient iterative channel estimation using modified root-MUSIC with spatial smoothing and the interference mitigation scheme. In Fig. 7, performance results with the adaptive scheme are presented. In order to initiate the convergence of the beamforming and the combining filters, the two first iterations only use pilot symbols with large step sizes. As the iteration number increases, smaller step sizes are employed and soft decoded bits from the previous iteration are used as pilots. With 5 iterations, the fully adaptive interference cancellation scheme with very low complexity adaptive N-LMS algorithms experiences a 0.5 dB loss to achieve a BER of 10^{-4} , as compared with the single-user reference. Finally, Fig. 8, which presents results in the 5th iteration with different estimation configurations, underscores the usefulness of both the iterative and data-aided aspects in the estimation process. Without these two features, the performance loss due to

estimation errors is not negligible. It is worth noting that the performance degradation induced by the complexity reduction in the non-adaptive scheme with simplified ML estimation and in the adaptive scheme are not prohibitive.

6. Conclusions

The iterative scheme including interference cancellation, beamforming and SISO decoding exhibits near single-user performance even when actual pilot- and data-aided space-time channel estimation is performed. This excellent performance is obtained thanks to the iterative structure of the receiver. Furthermore, we have shown that an equivalent adaptive structure could reduce the complexity with a reasonable loss in performance.

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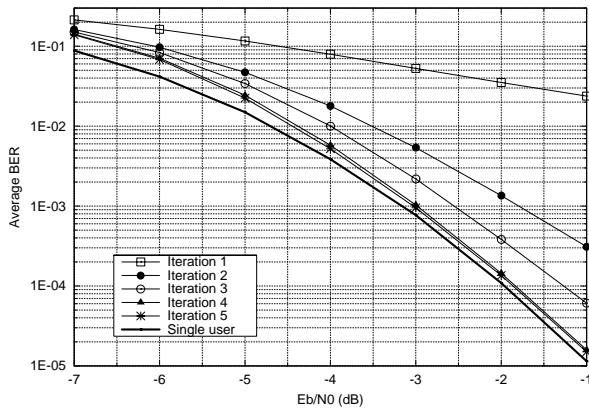


Fig. 6: Influence of the number of iterations with iterative space-time channel estimation.

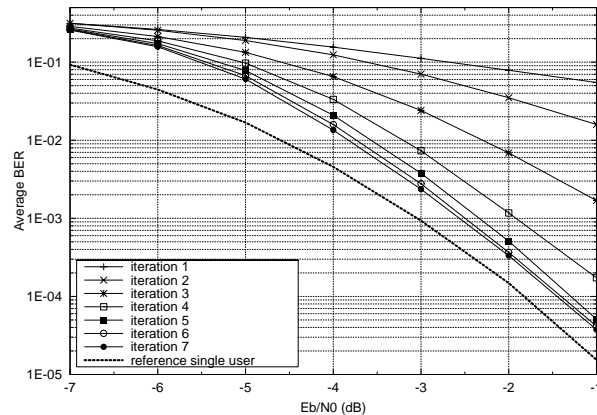


Fig. 7: Influence of the number of iterations with adaptive interference cancellation.

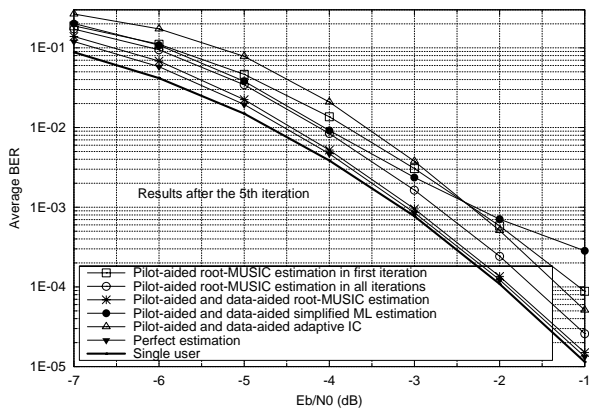


Fig. 8: Performance comparison of adaptive and non-adaptive interference cancellation schemes.

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