Iterative Interference Cancellation Scheme with Pilot-Aided Space-Time Estimation in DS-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 add a pilot-aided space-time channel estimation in each iteration. The iterative structure has two advantages: the observation signal used for estimation contains less interference from one iteration to the following and soft estimates of coded bits are available for data-aided estimation.

I. 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 (TDD) are considered to deal with the available paired and unpaired frequency bands, the FDD mode currently receives more interest from European manufacturers. The firstly proposed receivers will probably be implemented in a very simple way, sufficient to face the presumed relatively low traffic coming from third generation first users. Subsequently, advanced receivers will have to be employed to deal with the expected increase of traffic and associated interference. Therefore, interference mitigation techniques have to be considered in W-CDMA receivers, namely in Base Stations, where additional computational complexity may be tolerated. Because of the large size of spreading sequences in FDD mode, interference cancellation techniques [1] seem appropriate to mitigate intra-cell interference with a reasonable complexity increase. The whole interference on a given user is rebuilt from a bank of Rake receivers and respreaders, producing the contribution of all other users. This interference is then subtracted from the received signal to produce a new clearer observation for the user of interest. This process may 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 be used to reduce this interference [4][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.

The trend in [6] is to include as many signal processing functions as possible in the iterative process to further improve the receiver performance. In this paper, we propose to integrate an explicit space-time channel estimation in each iteration of an iterative space-time interference cancellation scheme. Since the estimation is renewed in each iteration, it benefits from the iterative process. Another approach to cope with unknown channels is presented in [7]: the adaptive iterative space-time interference cancellation. A comparison between these two approaches is performed in [8].

The paper is organized as follows: In section II, the principle of an iterative space-time interference cancellation is presented for a Direct Sequence (DS) CDMA system with time-multiplexed pilot symbols. The space-time channel estimation is then inserted in the iterative process in section III. Performance results presented in section IV show that the proposed detection scheme almost achieves single-user performance even with the highly loaded simulated system. Promising simulation results of the iterative receiver in a uplink UMTS FDD mode, where pilots are multiplexed in quadrature and signatures, are also presented.

II. ITERATIVE SPACE-TIME INTERFERENCE CANCELLATION

Let us first consider a DS-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 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 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 the data sequence. The $N_D = N_c / 2 + N_p$ obtained symbols, $D_k(0), \ldots, D_k(N_D - 1)$, are finally spread by a user-specific signature over $N_{chips} = SF.N_D$ chips $s_k(n)$, where SF is the spreading factor, before transmission on user k space-time channel. Each user's channel contains P paths, each path p having a delay equal to $\tau_{k,p}$ chips, a complex Gaussian coefficient $h_{k,p} = \rho_{k,p}\exp(j\nu_{k,p})$

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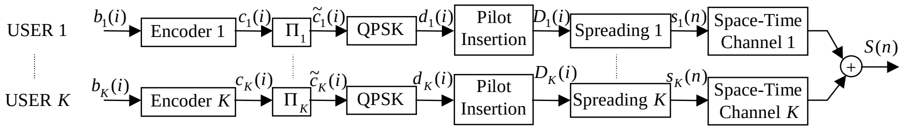

Fig. 1: Structure of Direct Sequence CDMA transmitters.

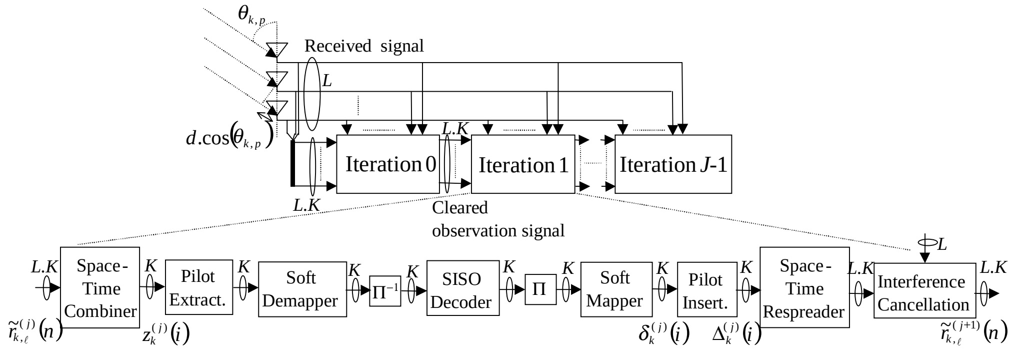

Fig. 2: Structure of the iterative space-time interference cancellation receiver.

and a direction of arrival (DOA) $\theta_{k,p}$.

Fig. 2 depicts an iterative space-time interference cancellation 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 $d.\cos(\theta_{k,p})$, 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

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

$$\text{where} \quad \xi_{k,\ell,p} = \nu_{k,p} + \varphi_{k,\ell,p} = \nu_{k,p} + (\ell-1).\varphi_{k,p}$$
$$= \nu_{k,p} + 2\pi \frac{d}{\lambda}(\ell-1)\cos(\theta_{k,p}) \tag{1}$$

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

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

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

In each receiver iteration j, a user-specific multi-antenna observation signal $\tilde{r}_{k,\ell}^{(j)}(n)$, where $\ell = 1,\ldots, L$, 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:

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

The space-time combiner is detailed on Fig. 3. Perfect estimation is assumed in this section. 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 $i = 0,\ldots, N_D-1$, which mainly contains the contribution of a single path. To cancel residual signals in non-desired directions, a beamforming combines signals from the L different antennas, using coefficients $\beta_{k,1,p},\ldots, \beta_{k,L,p}$:

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

Conventional beamforming is used here: $\beta_{k,\ell,p}=\exp(-j.\varphi_{k,\ell,p})$. Observations for all paths of user k are then combined using coefficients $\alpha_{k,1},\ldots, \alpha_{k,P}$ to provide us with a single observation $z_k{}^{(j)}(i)$ for each transmitted symbol $D_k(i)$:

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

Maximum Ratio Combining (MRC) is performed here: $\alpha_{k,p}=\rho_{k,p}\exp(-j.\nu_{k,p})$.

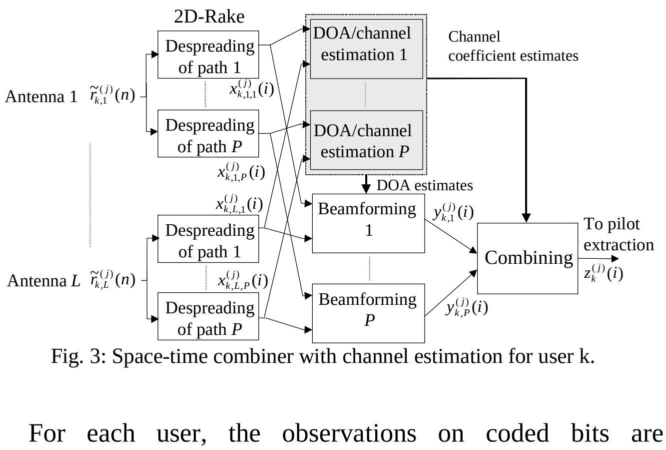

Fig. 3: Space-time combiner with channel estimation for user k.

For each user, the observations on coded bits are deinterleaved, after pilots have been extracted, and decoded by 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 reinserted to form the estimated sequence of transmitted symbols:

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

Symbols $\Delta_k^{(j)}(i)$ are respread and each user's contribution in the global interference is rebuilt on each antenna, by filtering the signal by the space-time channel impulse response. Using these contributions, interference cancellation is performed on the received signal to supply the following iteration with a new cleared user-specific 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.

III. Iterative Space-Time Interference Cancellation with Channel Estimation

As channels are unknown and pilots are available in the system, 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 accuracy will also improve, thus making the detection less erroneous. Hence, we expect a performance improvement in comparison with 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.

As depicted on Fig. 3, the space-time channel estimation individually computes the complex coefficient and the DOA for each path p of user k, *i.e.*, it works with signals $x_{k,\ell,p}^{(j)}(i)$ for $\ell = 1,\dots,L$ and $i = 0,\dots,N^{(j)}-1$. In iteration 0, the estimation is performed just using pilot symbols ($N^{(0)} = N_p$ and $\Delta_k^{(-1)}(i) = p_k(i)$ for $i = 0,\dots,N_p - 1$), whereas coded bit estimates are available to improve estimation in iterations $1,\dots,J-1$ ($N^{(j)} = N_D$). Two space-time channel estimation algorithms are considered: a root-MUSIC algorithm with spatial smoothing and a low-complexity estimator based on an approximation of the Maximum Likelihood (ML) criterion.

A. Root-MUSIC Estimator

We first apply a classical root-MUSIC algorithm to DOA estimation. A spatial smoothing [4] is added to allow it to distinguish coherent paths with same delay but different DOAs. Originally blind, the root-MUSIC algorithm is modified to benefit from the knowledge of pilot symbols. Indeed, for each user k, each path p and each iteration j, it performs the eigenvector decomposition of the spatial covariance matrix estimate

$$\hat{\mathbf{R}}_{k,p}^{(j)} = \frac{1}{N^{(j)}} \sum_{i=0}^{N^{(j)}-1} \mathbf{x}_{k,p}^{(j)}(i)\mathbf{x}_{k,p}^{(j)H}(i) \tag{7}$$

where $\mathbf{x}_{k,p}^{(j)}(i) = (x_{k,1,p}^{(j)}(i),\dots, x_{k,L,p}^{(j)}(i))^T$. $\bullet^T$ and $\bullet^H$ denote the transpose and the transpose-conjugate operations, respectively. In order to reduce the noise level on each covariance matrix element, we correlate the $(N^{(j)} \times L)$ received signal matrix $\mathbf{X}_{k,p}^{(j)} = (\mathbf{x}_{k,p}^{(j)}(0),\dots, \mathbf{x}_{k,p}^{(j)}(N^{(j)}-1))^T$ with the pilot sequence $\Delta_k^{(j-1)} = (\Delta_k^{(j-1)}(0),\dots, \Delta_k^{(j-1)}(N^{(j)}-1))^T$:

$$\mathbf{y}_{k,p}^{(j)T} = \frac{1}{N^{(j)}} \Delta_k^{(j-1)H} \mathbf{X}_{k,p}^{(j)}(i) \tag{8}$$

From vector $\mathbf{y}_{k,p}^{(j)}$ with size L, we obtain a new spatial covariance matrix estimate,

$$\hat{\mathbf{R}}'^{(j)}_{k,p} = \mathbf{y}_{k,p}^{(j)}\mathbf{y}_{k,p}^{(j)H} \tag{9}$$

the elements of which are less noisy. It can thus be used in the root-MUSIC algorithm to improve the accuracy of the DOA estimate $\hat{\theta}_{k,p}^{(j)}$. Finally, the complex channel coefficient estimate $\hat{\rho}_{k,p}\exp(j\hat{\nu}_{k,p})$ is computed at the beamforming output, by simple averaging.

B. Low-Complexity Estimator

Since the root-MUSIC algorithm is quite complex and as we aim at performing estimation in each iteration, we also consider a low-complexity estimator with the restrictive assumption that paths from different DOAs arrive with different delays. This estimator is derived from an approximation of the 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

$$x_{k,\ell,p}^{(j)}(i) = h_{k,\ell,p} D_k(i) + w'_{k,\ell,p}(i) \tag{10}$$

where $w'_{k,\ell,p}(i)$ is the noise sample on symbol i for antenna ℓ, user k and path p. For a given user and a given path, all noise samples $w'_{k,\ell,p}(i)$, $\forall \ell = 1,\dots, L$, $\forall i = 0,\dots, N^{(j)}-1$, are Gaussian and independent. To perform ML estimation of $\rho_{k,p}$, $\nu_{k,p}$ and $\theta_{k,p}$, we must minimize the quadratic Euclidean distance $d^{(j)}{}_{k,p}{}^2$ between the received sequence and the expected one:

$$d_{k,p}^{(j)2} = \sum_{i=0}^{N^{(j)}-1} \sum_{\ell=1}^{L} \left| x_{k,\ell,p}^{(j)}(i) - h_{k,\ell,p}\Delta_k^{(j-1)}(i) \right|^2 \tag{11}$$

To simplify the minimization, we perform it separately on each antenna to find the estimate $\hat{\xi}_{k,\ell,p}^{(j)}$: (12)

$$\begin{aligned}\hat{\xi}_{k,\ell,p}^{(j)} &= \underset{\xi}{\operatorname{argmin}}\, d_{k,\ell,p}^{(j)\,2} = \underset{\xi}{\operatorname{argmin}} \left(\sum_{i=0}^{N^{(j)}-1} \left| x_{k,\ell,p}^{(j)}(i) - \rho e^{j\xi}\Delta_k^{(j-1)}(i) \right|^2 \right) \\ &= \arg\left(\Delta_k^{(j-1)H} \mathbf{x}'^{(j)}_{k,\ell,p} \right) \in \left]-\pi;\pi\right]\end{aligned}$$

where $\mathbf{x}'_{k,\ell,p}{}^{(j)} = (x_{k,\ell,p}^{(j)}(0),\dots, x_{k,\ell,p}^{(j)}(N^{(j)}-1))^T$ and $\arg(a)$ is the argument of the complex number a. Since the antenna array is a ULA, a phase unwrapping followed by a linear regression yields the estimates $\hat{\theta}_{k,p}^{(j)}$ and $\hat{\nu}_{k,p}^{(j)}$ from $\hat{\xi}_{k,\ell,p}^{(j)}$,

$\ell = 1,\ldots, L$. Finally, $\hat{\rho}_{k,p}^{(j)}$ is obtained by minimization of $d^{(j)}{}_{k,p}{}^2$:

$$\hat{\rho}_{k,p}^{(j)} = \frac{\sum_{\ell=1}^{L} \mathrm{Re}\left\{ e^{-j.\tilde{\xi}_{k,\ell,p}^{(j)}} .\Delta_k^{(j-1)H} .\mathbf{x}'^{(j)}_{k,\ell,p} \right\}}{L.\Delta_k^{(j-1)H} .\Delta_k^{(j-1)}} \tag{13}$$

where $\tilde{\xi}_{k,\ell,p}^{(j)} = \hat{\nu}_{k,p}^{(j)} + 2\pi \dfrac{d}{\lambda}(\ell - 1)\cos(\hat{\theta}_{k,p}^{(j)})$.

These channel estimates, obtained either by the root-MUSIC or the low-complexity estimator, are then employed in the space-time combiner including conventional *i.e.*, MRC, beamforming and MRC path combining.

IV. Simulation Results

Simulation results are first presented for a highly interfered system, which was taken as a first simulation step to validate the detection algorithm. The iterative receiver is subsequently tested for a 78 kbps UMTS FDD uplink scenario with various amplitudes of pilots.

The first considered uplink transmission deals with the signals of 9 users, spread by a factor 5, each user transmitting on a distinct 2-path channel. A rate ½ (7,5) convolutional channel coding is applied and 5 antennas compose the ULA at the receiver side. Blocks of 240 QPSK symbols, including 17 % of pilots, are transmitted, which the space-time channel is assumed stationary over. For each delay, a channel coefficient and a single DOA in a 120° sector, are randomly chosen in each transmission block. All users have same power and perfect power control is assumed, which explains why single-user performance on AWGN channel is taken as a reference, with a shift in Signal to Noise Ratio (SNR) due to the pilot overhead. The $10\log_{10} L$ dB gain due to multiple antennas explains the very low SNR values. To demonstrate the interference mitigation capability of the receiver, the average Bit Error Rate (BER) over all users is drawn versus SNR for different configurations on Fig. 4. With only 4 iterations, almost the whole multiuser interference is removed thanks to the efficient iterative channel estimation using root-MUSIC with spatial smoothing, modified to deal with pilot symbols, and the interference mitigation scheme. When the space-time channel is only estimated in iteration 0 and the same estimate is used in all the following iterations, a performance loss of 0.8 dB appears after 4 iterations and for a BER equal to 10^{-4}. We find similar results for the low-complexity estimator. The loss observed by estimating the channel in iteration 0 instead of estimating it in all iterations is equal to 2.4 dB for a 10^{-2} BER. The estimator's low complexity is translated into a loss of 0.9 dB at a 10^{-3} BER as compared to the root-MUSIC performance.

In the second simulation scenario, 16 users transmit 78 kbps data in the uplink of a UMTS system in FDD mode [10]. A block of 1560 information bits is transmitted over 2 frames of 10 ms. 8 parity check bits are added before rate 1/3 (557,663,711) convolutional encoding. Each frame is divided into 15 timeslots. Each timeslot contains 2560 QPSK chips *i.e.*, 160 coded bits, spread on the in-phase signal by a factor 16, and 10 control bits, including 6 pilot bits, spread on the quadrature signal by a factor 256. Data bits form the Dedicated Physical Data Channel (DPDCH) and control bits form the Dedicated Physical Control Channel (DPCCH). After despreading and a multiplication by $-j$ for control bits' observation, we obtain separately complex data observations and complex control observations. Theses observations may be used in the space-time channel estimation. Note that pilot symbols, used as a reference by the estimation algorithm, are now BPSK symbols. Low-complexity channel estimation is performed, with a 2-antenna ULA.

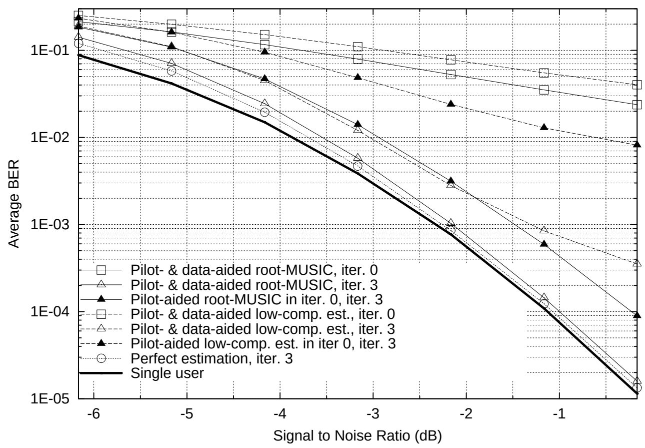

Fig. 4: Comparison of different estimation schemes for the general scenario.

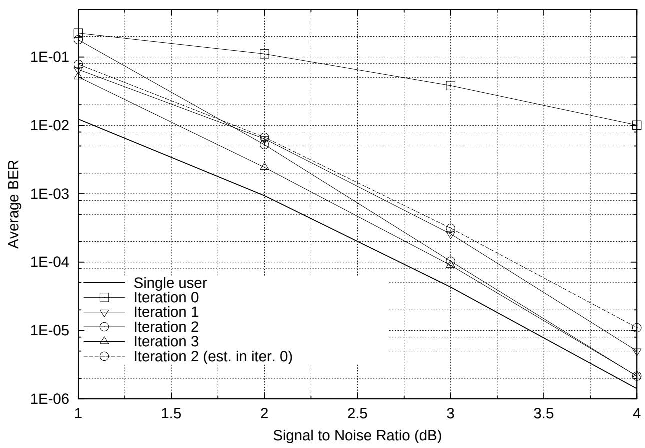

Fig. 5: Influence of the number of iterations in the 78kbps FDD UMTS scenario with equal DPDCH and DPCCH amplitudes.

Fig. 5 shows the iterative receiver performance results, when in-phase (DPDCH) and quadrature (DPCCH) signals have the same amplitude. Thanks to the high pilot energy as compared to the data energy, the space-time channel estimation is very accurate and only 3 iterations are necessary to achieve single-user performance. However the single-user performance is quite bad, because of the high overhead from DPCCH. A performance loss of 0.4 dB to achieve a BER equal to 10^{-5} is encountered when the channel estimation is only performed in iteration 0, which is still reasonable. The high quality of this first estimate is due to the high pilot energy.

To really measure the performance improvement brought by the receiver, we have to simulate a more critical scenario, in which the amplitude of DPCCH equals 4/15 of the DPDCH amplitude: the power difference between DPDCH and DPCCH equals –11.5 dB. The corresponding results are presented on Fig. 6. Since the overhead from DPCCH is smaller than for Fig. 5, the single-user performance results are improved. Because of the low pilot energy, the first estimate is not sufficiently accurate anymore and a considerable performance loss is observed when estimation is only performed in iteration 0. This performance loss with respect to single-user performance can be reduced to 0.3 dB at a 10^{-5} BER after 5 iterations, when channel estimation is performed in each iteration.

V. 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, for both general and UMTS FDD uplink scenarios. The structure including channel estimation in each iteration is particularly interesting in UMTS FDD uplink when DPCCH has a weak amplitude with respect to DPDCH amplitude.

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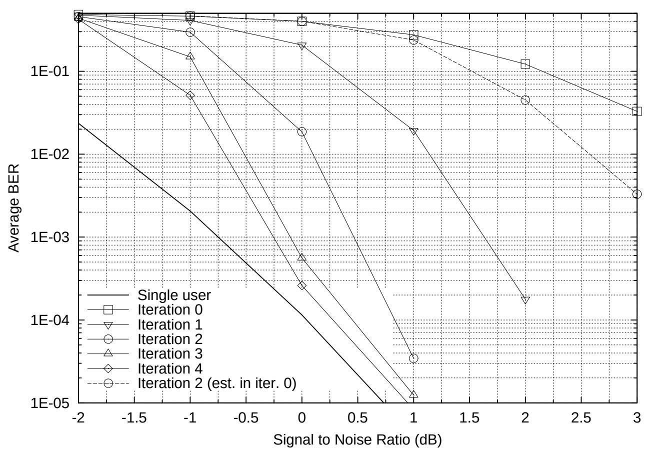

Fig. 6: Influence of the number of iterations in the 78kbps FDD UMTS scenario with DPCCH amplitude equal to 0.27 DPDCH amplitude.