Performance of Joint Space-Time Equalization and Decoding Techniques for Wireless Systems

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Abstract- **This paper shows how performance gain in a mobile radio environment can be achieved by using joint space-time equalization and decoding. We apply different joint procedures with two types of space-time equalization techniques. As a matter of fact, the joint techniques are able to provide a considerable gain with none to small additional computational cost.**

I. Introduction

In mobile radio environments, it is well known that intersymbolic and co-channel interference (ISI and CCI) are among the major impairments to achieve higher capacity and data rates. These negative effects can be mitigated with the aid of an antenna array, which works in the spatial domain to form beams in the direction of arrival (DOA) of the desired signal and suppress CCI. Additionally, the antenna array is able to provide array gain and make use of the spatial diversity, if available, to compensate the decrease of the signal-to-noise ratio, due to fading. It is also possible to mitigate ISI, but this can demand a high number of antennas, since mobile radio-channels present a rich multipath environment [1].

On the other hand, in order to mitigate ISI, it is possible to make use of temporal equalizers. These equalizers can consist of a finite impulse response (FIR) filter, an infinite impulse response (IIR) filter, or a maximum likelihood sequence estimator (MLSE). Moreover, the use of fractionally spaced, instead of symbol spaced equalizers, makes it possible to reduce CCI too. However, due to fundamental limitations [1], noise enhancement may occur, leading to unsatisfactory performance.

Therefore, space-only and time-only processing cannot efficiently mitigate both CCI and ISI. The combination of both space and time processing leads us to the space-time (ST) processing, which enables fully exploitation of spatial and temporal characteristics of the mobile radio channel. This is the enabling key to suppress both CCI and ISI and so improve network capacity, coverage and quality.

Besides, error correction codes are usually used to improve system performance. The use of such codes makes it possible to achieve very small error rates which enables, for example, the transmission of data through wireless systems.

Nevertheless, equalization and decoding techniques are commonly investigated and used separately. However, a joint utilization of both techniques can provide more robustness and, therefore, additional performance, with none to small additional computational cost. Some references dealing with joint techniques can be found in the literature [4-7]. In a recent work [8], we introduced a new approach to be applied in SISO (single input – single output) temporal equalizers. In the present paper, we aim to employ such new joint technique in a space-time framework. We compare the performance of our proposition with previous obtained results by means of computer simulations, taking as a reference the more general structure presented in [5].

The paper is organized as follows. In section 2, space-time equalization techniques are presented. The joint equalization and decoding procedures are briefly explained in section 3, where the proposed solution is posed. The system model and the simulation results are respectively presented in section 4 and 5. Finally, the conclusions are stated in section 6.

II. Space-time Equalization

Space-time equalization has received a lot of attention from researchers along last years. In this paper, we deal with two space-time techniques. The first one will be named conventional, since it uses the typical configuration of antenna array with finite impulse response (FIR) temporal filters to form the space-time front-end. Aiming more degrees of freedom and more robustness, one can use a decision-feedback equalizer (DFE) or a MLSE in the output of this ST front-end, as shown in figure 1.

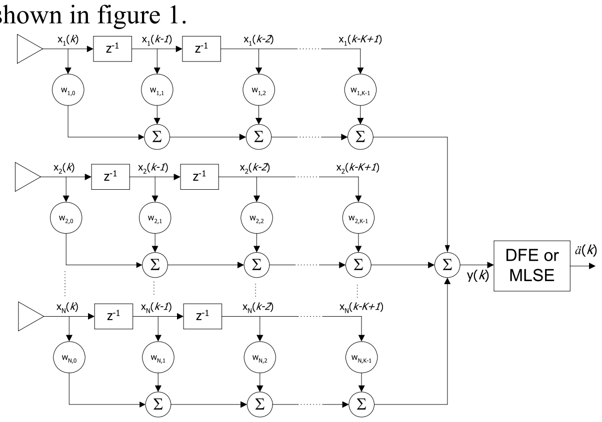

Fig. 1: Conventional space-time filtering structure

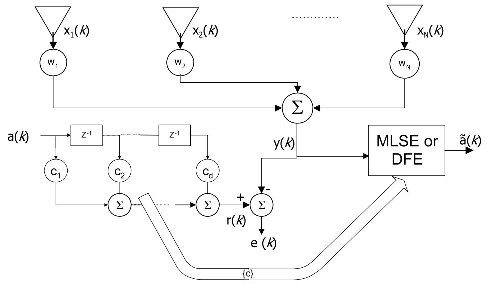

Fig. 2: Decoupled space-time processing technique

The second space-time technique to be investigated comes from a modification of the conventional antenna array. In order to improve the degree of freedom without increasing the number of antennas, some authors (e.g. [2]) have proposed an alternative configuration we have called decoupled space-time (D-ST) processing [3]. The corresponding scheme is presented in figure 2.

In this structure, the role of the antenna array is to cancel only the CCI. To accomplish this task, it is trained with the modified sequence $r(k)$ that contains the ISI estimation. This enables the array to ignore the desired user multipaths and provide more degrees of freedom to cancel the CCI. The transversal filter that modifies the training sequence is adapted with the error $e(k)$, obtained by comparing the filter and the array outputs. The array output $y(k)$ contains the remained ISI, to be eliminated by the MLSE or DFE temporal equalizer [2]. It is also possible to use the coefficients c_i (i = 1, … , d) of the transversal filter as a channel estimator that can be directly employed into the temporal equalizer [3].

III. Joint Equalization and Decoding Procedures

Usually equalization and channel decoding are treated separately. However, recent works ([4] and [6]) have shown that the joint use of equalization and channel decoding can lead to additional performance gain.

Clearly, the great variety of equalization and error correction leads to a number of possible joint techniques. In this paper, we are going to focus our attention to three similar procedures based on the DFE and trellis-coded modulation (TCM).

The first procedure is described by [4] and is depicted in figure 3. In this technique, the feedback filter (FBF) is divided in two parts: FBF1 and FBF2. The FBF1 receives the hard or soft decisions. The FBF2 is fed by regenerated symbols, obtained from the TCM decoder with a tentative decision delay (*TD*). These regenerated symbols are more reliable then the symbols obtained from the soft or hard decision device, so that the effect of error propagation could be mitigated. It is worth noting that there is a trade off in this structure, since for high values of *TD*, the symbols to be fed into FBF2 are more reliable, but the output of FBF1 becomes less reliable due to the filtering of more decided symbols without the reliability of the code.

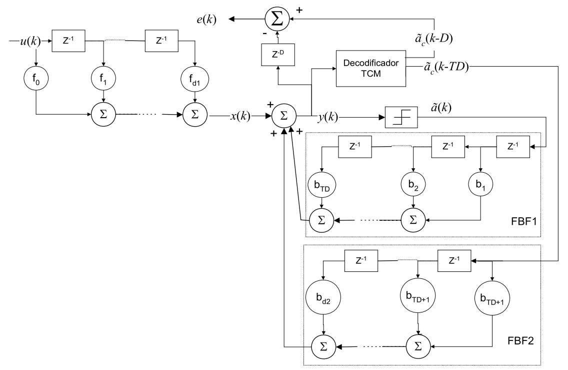

Fig. 3: Joint technique proposed in [4]

In fact, due to the memory characteristic of TCM system, some symbol transitions are not allowed. Hence, even for *TD* = 0, the regenerated symbols are more reliable than the decision device. This is illustrated in figure 4 by plotting the symbol error rate *vs.* the symbol energy-to-noise ratio (E_s/N_o) for hard decision and for some values of *TD*.

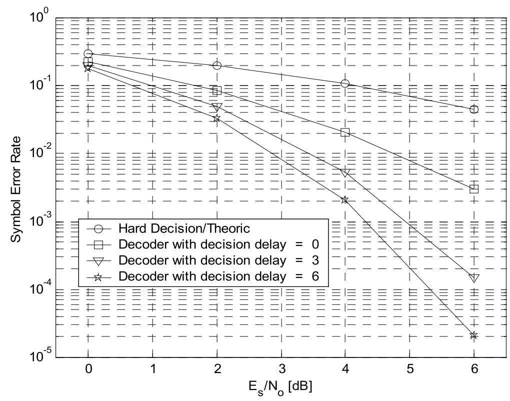

Fig. 4: Hard decision vs. decoder performance

Based on this result, we propose to remove the FBF1 part and use the survivor path of the TCM decoder to take benefit of the more reliable values of *TD*. This is the feature of the structure depicted in figure 5.

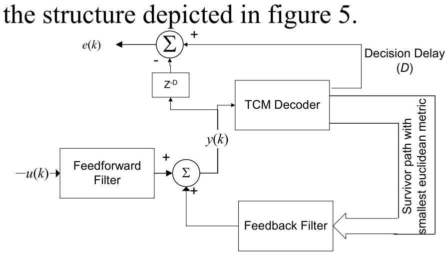

Fig. 5: New joint DFE & TCM decoder procedure

In addition, another method can be used to mitigate error propagation. It was proposed by [5] and is called delayed decision-feedback sequence estimator (DDFSE). It was originally introduced as a technique to reduce the computational effort of the MLSE, using an FBF to reduce the channel memory effect. Moreover, reference [5] has also realized that this structure can be used as a joint equalization and decoding technique. The DDFSE with TCM is depicted in figure 6.

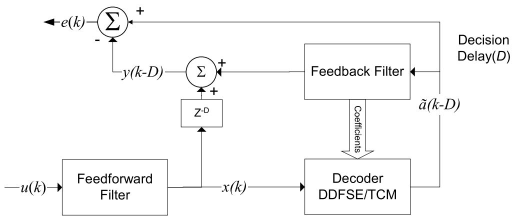

Fig. 6: Delayed Decision-Feedback Sequence Estimator

By comparing figures 5 and 6, it can be noted that our new joint DFE and decoding technique can be seen as a simplified version of the DDFSE one, but with a significant gain in the computational cost.

The different solutions described in this section will be compared by simulations. Afterwards, it is worth posing some characteristics to be assumed in the system model.

IV. System Model

The TCM system used in the simulations consists in a convolutional encoder with a [64 74] (octal) sequence generator and a $\pi/4$-QPSK modulation, with gray code. We do not use interleaving in this process due to limitations imposed by the joint equalization and decoding techniques. We have assumed a TDMA system with a time slot composed by a training sequence of 40 symbols and a data sequence of 225 symbols. We consider the last 12 data symbols as tail symbols. The symbol rate is 270.833 ksymbols/s.

After the training sequence the equalizer is switched to the decision-directed mode in order to track the channel variations. In our system model, we are also considering that the CCI, if present, is symbol and slot synchronized with the desired user. We also assume that perfect symbol synchronization is achieved.

The antenna array for all equalization structures is disposed in an uniform linear arrangement. In this case, the phase difference between two consecutive antennas associated to the n^{th} received wave is given by:

$$\phi_n = \frac{2\pi d \sin(\theta_n)}{\lambda}, \qquad (1)$$

where θ_n is the DOA (direction of arrival) of the n^{th} wave, d is the distance between the antennas in wavelengths, and λ is the carrier wavelength. It is assumed that the first antenna has a null phase reference value. By considering $d = \lambda/2$, it is possible to define the antenna array response vector as:

$$\mathbf{f}(\theta) = \begin{bmatrix} 1 & e^{j\pi\sin(\theta)} & \cdots & e^{j(M-1)\pi\sin(\theta)} \end{bmatrix}^T \qquad (2)$$

The following equation describes the Jakes [9] model for a space-time flat fading environment:

$$\boldsymbol{\alpha}(t) = N^{-1/2} \sum_{n=1}^{N} e^{j[2\pi f_d \cos(\phi_n)t + \Phi_n]} \mathbf{f}(\theta_n) \qquad (3)$$

where N is the number of received waves, assumed to be 80; Φ_n and ϕ_n are random phases related to the n^{th} wave's delay and incident angle respectively, considered to be uniformly distributed between 0 and 2π; θ_n is a uniformly distributed random variable, which can assume the values $[\theta - \Delta/2, \theta + \Delta/2]$, where θ is the path's DOA and Δ is the angle spread. The variable f_d is the maximum Doppler frequency.

The mobile radio channel is usually modeled by a sum of delayed paths from a given transmitter, mobile or radio station. Thus, it is possible to represent the channel impulse response as:

$$\mathbf{h}(t) = \sum_{i=1}^{paths} \boldsymbol{\alpha}_i(t) p(t) \delta(t - t_i), \qquad (4)$$

where t_i is the path delay; $\boldsymbol{\alpha}_i$ is the space-time fading of the i^{th} path; and $p(t)$ is the shaping pulse which is a raise cosine with roll-off factor $\alpha = 0.35$.

For simulation purposes, the channel model employed uses two types of configuration [10]: the typical urban (TU) and the hilly terrain (HT) profiles, with 6 paths each one. The CCI, when present, has the same profile used for the desired user.

For signal processing purposes, we have truncated the channel impulse response. Hence, in our simulations, eq. (4) obeys the following restriction:

$$\mathbf{h}(t) = \begin{cases} \mathbf{h}(t), & if \quad -2T \le t \le 8T \\ 0, & otherwise \end{cases} \qquad (5)$$

Thus, by considering a single-user case in a single-input multiple-output context, the signal model for the output vector of the antenna array, $\mathbf{x}$, is easily written as:

$$\mathbf{x}(t) = \sum_{k=-\infty}^{\infty} a(k)\mathbf{h}(t - kT) + \mathbf{n}(t), \qquad (6)$$

where $\mathbf{n}(t)$ is the vector with additive white Gaussian noise and $a(k)$ is the desired user symbol.

V. Simulation Results and Performance Assessment

For all simulations, we have used the RLS algorithm to guarantee convergence during the training period. The forgetting factor is equal to 0.93. After training, we have switched to the NLMS algorithm with a step-size equal to 0.1. The NLMS has shown to be efficient to track the slow channel variations. In all simulations, the adopted Doppler frequency is $f_d = 40$ Hz ($f_d T = 1.48\text{x}10^{-4}$) and the decoding delay is 12.

In order to well access the difference between the approach in [4] and our proposal, we show first a result with only temporal equalizers, before dealing with the space-time configurations.

We have used a channel with tree paths with delays 0, $1.5T$ and $6.8T$, and power 0 dB, –6 dB and –8 dB respectively and $f_d = 40$ Hz. The technique in [4] is named incomplete feedback, since it uses partial feedback of the TCM decoder. We have used in this structure both hard and soft decision devices. The soft decision device is a simplification of the MAP criterion [4] and is given by:

$$\tilde{a}(k) = \tanh\left(\gamma \operatorname{Re}\{y(k)\}\right) + i \tanh\left(\gamma \operatorname{Im}\{y(k)\}\right), \quad (7)$$

where γ is a scalar that depends on the channel configuration. Its value is obtained by simulation and, in our example, we set $\gamma = 3$.

The conventional DFE corresponds to the case where there is no feedback from the decoder to the DFE.

Our new proposal is called full feedback, since it uses the whole survivor path with regenerated symbols as input vector of the FBF. The perfect DFE is fed and trained by correct symbols during all time slots. The feedforward has 3 coefficients and the feedback filter has 8 coefficients in all structures. The training delay is equal to 2 symbols periods.

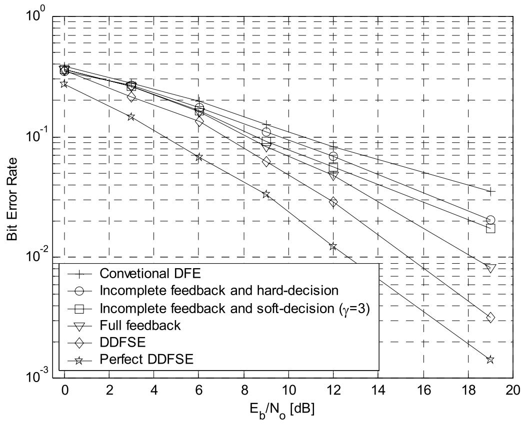

Fig. 8: Performance of the joint techniques

As we can see, our new proposal has better performance than [4]'s technique, even with soft decision. Moreover, the DDFSE has the best performance. The DDFSE is less sensible to error propagation effect and, therefore, it is closer to the perfect DFE, which has not error propagation. However, it is worth noting that the DDFSE has a higher computational cost if compared with the other techniques. On the contrary, reference [4] and our new proposal have almost the same computational complexity of the conventional DFE. Then, for the space-time framework in the sequel, the incomplete feedback solution will be discarded. We refer our structure as the space-time DFE with decoder feedback.

For the following simulations, we have used two antennas in both decoupled and conventional structures, unless otherwise specified. In all simulations we have assumed spatial diversity, i.e., the angle spread is equal to 360°.

The decoupled space-time technique has another use of the survivor path with the smallest euclidean metric. It is also used in the filter that modifies the training sequence in our new technique and in the DDFSE. This may guarantee better tracking performance. The decoupled technique has 10 coefficients in the filter that modifies the training sequence and the coefficient c_2 was set to 1 in order to avoid the null solution [2]. It uses a DFE as temporal equalizer with 2 feedforward coefficients and 8 feedback coefficients, obtained from the filter that modifies the training sequence, by using the MMSE solution and a delay of 2 symbols.

In the conventional space-time (ST-DFE) structure, the ST front-end has 3 coefficients per antenna and the feedback filter attached to this front-end has 8 coefficients. The training delay is equal to 2 symbols periods.

In figures 9 to 11, we present the results for the TU channel.

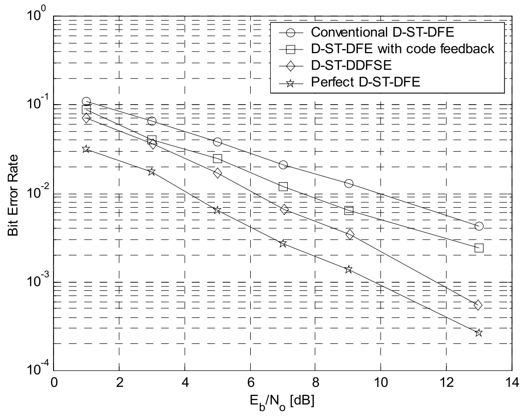

Fig. 9: Performance of the D-ST structure for the TU channel

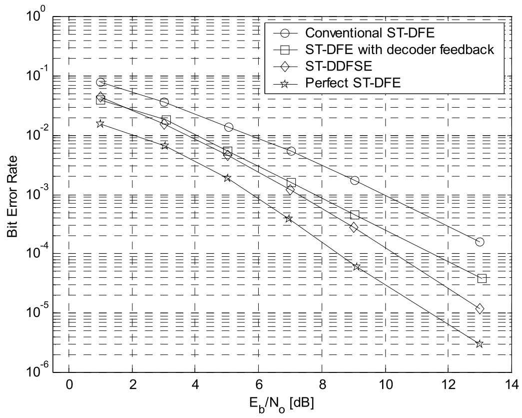

Fig. 10: Performance of the ST structure for the TU channel

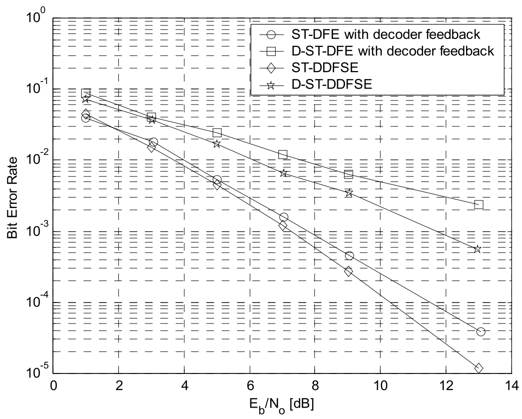

Fig. 11: Performance comparison for the TU channel

As we can see, the performance of the joint techniques is always better than the conventional techniques, with a gain of at least 2 dB. The DDFSE is closer to the performance of the perfect DFE in both conventional and decoupled ST structures. It is worth noting that the performance difference between our new technique and the DDFSE is much higher at the D-ST structure. This may occur because the D-ST structure cannot exploit the temporal diversity as the ST structure does. This makes the D-ST structure much more susceptible to error propagation in our new technique. This also explains why the ST structure has so much better performance than the decoupled technique in this case.

In figures 12 to 14 we show the results for the HT channel. The performance of both techniques is almost the same in this case. This happens because the ST structure cannot make use of temporal diversity in this HT profile, due to insufficient number of taps per antenna. This does not allow the ST front-end to work as a Rake receiver. Also, in this channel, there is a small difference between our proposal and the DDFSE proposal. Both techniques have almost the same performance of the perfect DFE.

In the next simulations, we use two interferers, each one with a HT profile and $f_d = 40$ Hz. In these simulations the number of antennas in both space-time structures were set to 3. The E_b/N_o was fixed in 10 dB. Both interferers have the same mean power, which were changed to attain the desired signal-to-interference-ratio.

The performance of the decoupled structure outperforms the conventional structure in this case, due to the increasing in degree of freedom that solution provides. It should be expected that the ST-DFE/DDFSE perform better when more taps per antenna are used, since this may give more degrees of freedom to this structure and, additionally, allow it to make use of the most delayed multipaths. However, slower convergence, higher stationary error and the difficulty in finding the optimum training delay are additional problems that should be taken into account.

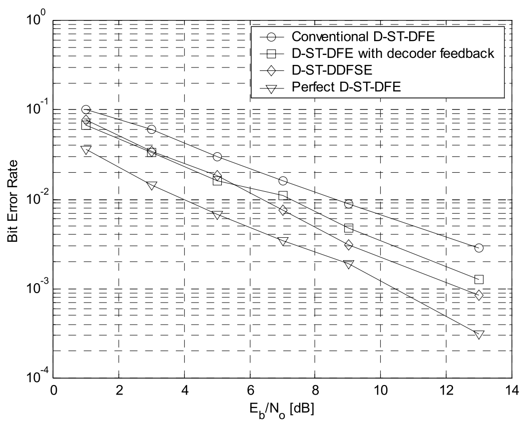

Fig. 12: Performance of the D-ST structure for the HT channel

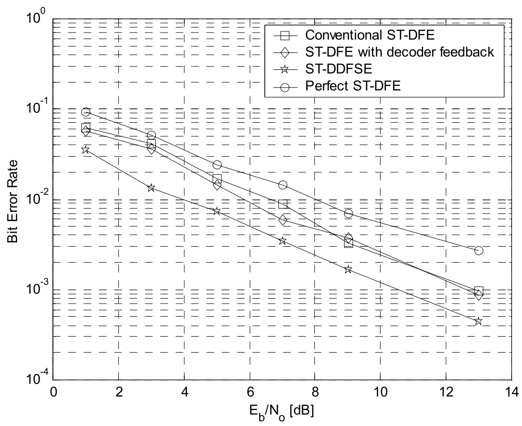

Fig. 13: Performance of the ST structure for the HT channel

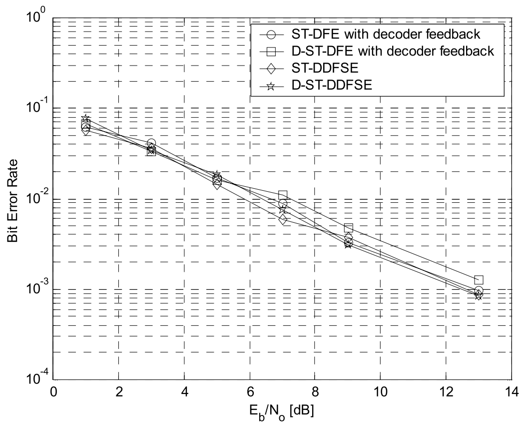

Fig. 14: Performance comparison for the HT channel

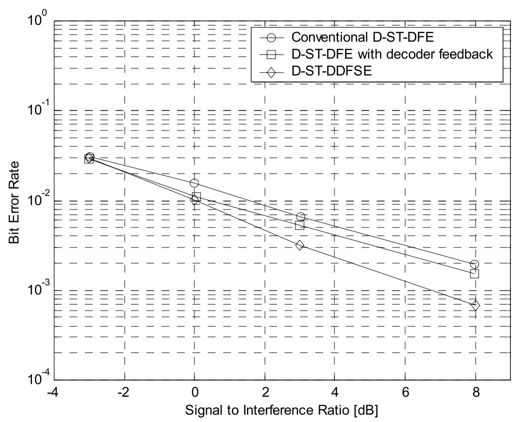

Fig. 15: Performance of the D-ST structure for the HT channel in the presence of CCI

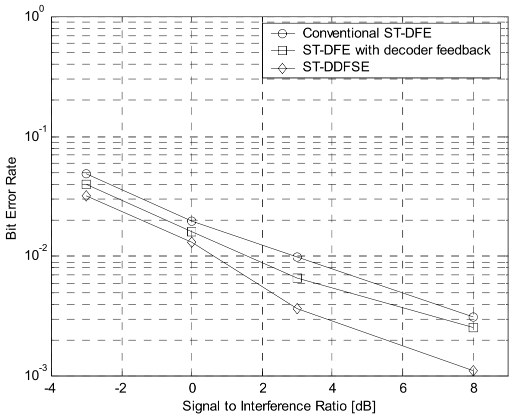

Fig. 16: Performance of the ST structure for the HT channel in the presence of CCI

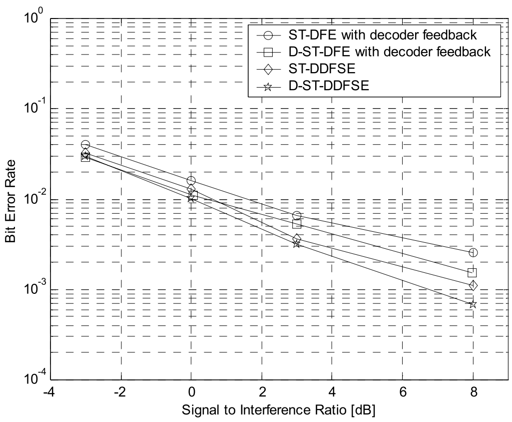

Fig. 17: Performance comparison in the presence of CCI

VI. Conclusions & Perspectives

We have shown how performance gain in a mobile radio environment can be achieved by using joint space-time equalization and decoding. Furthermore, the use of space-time processing is also fundamental to deal with typical impairments of the wireless environment, as high delay spread, fading and co-channel interference.

Among the three joint equalization and decoding techniques presented in this paper, we have proposed a new technique, which outperforms [4]'s technique. In fact, the so-called DDFSE has the best overall performance yet but in the expenses of a significantly higher computational cost.

On the other hand, our technique is limited to TCM systems. However, block and convolutional codes can be used, but in an iterative manner like turbo-equalization and techniques such as [6], which demands higher computational costs. Further studies will consider such aspects.

Acknowledgment

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