Turbo Decoding of Error Control Coded and Unitary Precoded OFDM *

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***Abstract*— This paper addresses the design of a novel scheme that combines error control coding (EC) and unitary precoding (UP) in order to obtain high diversity gains in OFDM systems with low complexity. The overall diversity of the proposed system is shown to be the product of the individual diversities achievable by the error control coding and by the unitary precoding, while the complexity is just a linear multiple of the sum of their individual complexities. In a practical HiperLan/2 setup, the proposed system achieves a gain of 4dB, (3.3dB), at a bit error rate (BER) of 10^{-3}, relative to conventional OFDM systems that only deploy convolutional (turbo), coding.**

I. INTRODUCTION

Orthogonal Frequency Division Multiplexing (OFDM) converts a frequency selective fading channel into parallel flat fading subchannels, thus making equalization at the receiver very simple. Unfortunately, uncoded OFDM suffers from loss of diversity; it has diversity order one, though an L^{th} order frequency selective channel is capable of supplying diversity up to order $L+1$; see e.g., [11]. This is a manifestation of the fact that uncoded OFDM cannot guarantee symbol detectability in the presence of channel nulls located on the Fast Fourier Transform (FFT) grid. Convolutionally coded OFDM (CC-OFDM) is deployed in practical systems to alleviate this problem at the cost of bandwidth expansion, where maximum likelihood (ML) decoding is applied using Viterbi's Algorithm [4]. Turbo Coded OFDM (TC-OFDM) has also been proposed to improve performance of CC-OFDM at the expense of increased complexity [6].

An alternative recent solution to mitigating channel nulls is the use of linear or unitary precoding [7, 11]. Unitary constellation precoded (UP) transmissions can guarantee symbol detectability and maximum diversity at high SNR without bandwidth expansion. Although coding is effective in combating errors irrespective of how they arise, linear precoding has been established to be very effective in combating errors induced by fading [11]; therefore, systems employing only coding may not perform well in fading environments.

With either channel coding or precoding alone, both CC-OFDM and UP-OFDM have inherent limitations to the amount of diversity they can achieve. In the former case, the diversity is equal to the minimum free distance d_{free} of the underlying error control code. A code of higher constraint length is required to obtain a high d_{free}, but in such cases ML decoding becomes prohibitive. In the latter case, the diversity of the system is limited by the precoder size that is also limited by the computational complexity for optimal decoding.

In this paper, we propose and study the performance of a novel strategy that combines *error control coding* and *unitary precoding* in OFDM, with the aim of combating both fading-induced and noise induced errors. We term our proposed system as EC-UP-OFDM, where EC can be replaced by either CC (convolutional coding) or TC (turbo coding). The overall diversity of the system is shown to be the product of the individual diversities of UP-OFDM and EC-OFDM. To facilitate the equalization, we employ precoders of small size, and develop an iterative (turbo) receiver relying on soft-in soft-out (SISO) modules which exchange extrinsic information of the coded / information bits between the equalizer and the decoder. The complexity of the proposed iterative receiver is just a linear multiple of the sum of the individual complexities of UP-OFDM and EC-OFDM.

In a practical HiperLan/2 setup, simulation results show that CC-UP-OFDM achieves a 4dB gain over CC-OFDM, while TC-UP-OFDM has a 3.3 dB gain over TC-OFDM, at a bit error rate of 10^{-3}. Hence, the combined use of both error control coding and precoding is an effective fading countermeasure. The improved performance comes at the cost of moderate increase in complexity.

II. COMBINED CODING WITH UNITARY PRECODING

In this section, we describe the system model of error control coded and unitary precoded OFDM (EC-UP-OFDM) and analyze its performance.

A. EC-UP-OFDM system model

With reference to Figure 1, we consider a wireless link employing error control coded transmissions over Rayleigh fading frequency selective channels. The information binary bit stream $\{d_i\}$ is encoded by a channel encoder. The output bit stream $\{b_i\}$ is modulated to obtain the symbol sequence $\{s_i\}$ which after interleaving by the interleaver Π_1 yields the symbol stream $\{s_n\}$. Symbol blocks $\{\bar{s}_n\}$ of size N are formed by using a serial to parallel (S/P) converter which are then linearly precoded by multiplying them by a non-redundant unitary matrix Θ_N, of size $N \times N$, with entries in the complex field. To reduce the complexity we choose the precoder Θ_N to be block diagonal; $\Theta_N = diag(\Psi_K, \Psi_K, \cdots, \Psi_K)$, where Ψ_K is of size $K \ll N$. The precoded symbols are interleaved by a interleaver Π_2. The precoded blocks are

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OFDM modulated by taking the Inverse FFT at the transmitter, and inserting the cyclic prefix (CP). Without unitary precoding, CC-OFDM with CP is employed in the current HiperLAN2 standards [3]. Our distinct feature at the transmitter side is the combination of unitary precoding along with error control coding (CC or TC).

OFDM converts the multipath frequency-selective channel into a set of equivalent flat fading channels. The received samples after the deinterleaver Π_2^{-1} can be written as

$$y_n = \alpha_n \bar{s}_n + \eta_n \tag{1}$$

where $\bar{s}_n$ is a precoded symbol (not a symbol block). In the presence of perfect interleaving, the multiplicative coefficients $\{\alpha_n\}$ are independent. Perfect interleaving can be obtained by having the size of the interleaver Π_2 large so that the precoded symbols are spread over a time period which is much more than the coherence time of the channel. Perfect interleaving can also be obtained if the channel taps are independent and the symbols are assigned independent subchannels as in GUP-OFDM [7]. In this case however, we require the size of Π_1 to be large (i.e., the symbols are spread over a time period which is much larger than the coherence time). Thus one of the purposes of Π_1 and Π_2 is to approximate perfect interleaving as much as possible which offers us the uncorrelated fading coefficients α_n in (1). From now on we assume that the system has achieved perfect interleaving for simplicity of system design and performance analysis. Practical channel setups with correlated α_n will be tested mainly through simulations.

Unitary precoding spreads each information symbol to multiple (K) independent subchannels, which increases the diversity of our transmissions. For low-complexity equalization, we prefer selecting a small K. In practice, we propose to choose $K = 2$ or $K = 4$, with the following two precoders:

$$\Psi_2 = \frac{1}{\sqrt{2}}\begin{bmatrix} 1 & e^{j\frac{\pi}{4}} \\ 1 & e^{j\frac{5\pi}{4}} \end{bmatrix}, \Psi_4 = \frac{1}{2}\begin{bmatrix} 1 & e^{j\frac{\pi}{8}} & e^{j\frac{2\pi}{8}} & e^{j\frac{3\pi}{8}} \\ 1 & e^{j\frac{5\pi}{8}} & e^{j\frac{10\pi}{8}} & e^{j\frac{15\pi}{8}} \\ 1 & e^{j\frac{9\pi}{8}} & e^{j\frac{18\pi}{8}} & e^{j\frac{27\pi}{8}} \\ 1 & e^{j\frac{13\pi}{8}} & e^{j\frac{26\pi}{8}} & e^{j\frac{39\pi}{8}} \end{bmatrix}$$

These two precoders have been designed using algebraic number theory [7], and have been shown to achieve maximum coding gains in the class of linear precoders with the same diversity. Another interesting property is that the absolute value of each element of the precoder is $1/\sqrt{K}$, a fact which we will shortly use in the performance analysis.

B. Performance Analysis

We will study the performance of EC-UP-OFDM, under the assumption of soft ML decoding. This analysis has also been done for more general codes in [10]. Here we will investigate the system in which a (n_0, k_0) convolutional code is employed. We use the union bound to bound the error by calculating the pairwise error probability (PEP) between two coded sequences. Consider an error event of weight w wherein the coded symbol sequence $\mathbf{S} = (..., s_0, s_1, s_2, ...)$ is wrongly decoded as $\tilde{\mathbf{S}} = (..., \tilde{s}_0, \tilde{s}_1, \tilde{s}_2, ...)$. Denote s_{i_m} as the interleaved symbol corresponding to s_m. Then $\tilde{s}_{i_m} \neq s_{i_m}, m = 1, 2, ..., w$, are distinct symbols in the sequences obtained by interleaving $\mathbf{S}$ and $\tilde{\mathbf{S}}$. The interleaver Π_1 is designed such that $|i_j - i_m| > K$ for $j, m \in (1, 2, ...w)$, for small w's. This guarantees that the symbols $\{s_{i_m}\}$ enter different size K precoders. The pairwise error probability between the two sequences $\mathbf{S}$ and $\tilde{\mathbf{S}}$ is dependent on the distance between the two respective received sequences. We first note that single error in $\mathbf{S}$ causes K precoded samples at the receiver to be different. Thus, there will be Kw non identical precoded samples between the two received sequences constituting the error event, in the absence of noise. Specifically, the symbols s_{i_m} and $\tilde{s}_{i_m}$ belong to the blocks $\mathbf{s}_{j_m}$ and $\tilde{\mathbf{s}}_{j_m}$, respectively, where $j_m = \lfloor i_m/K \rfloor$. Hence, we have $y_{j_m K+k} \neq \tilde{y}_{j_m K+k}, \forall k = 0, \ldots, K-1$, corresponding to $s_{i_m} \neq \tilde{s}_{i_m}$. Let $\phi_{i,j}$ denote the $(i,j)^{th}$ entry of the precoding matrix Ψ_K. The distance between the two received sequences conditioned on the given channel $\mathbf{h}$ is then given by

$$\begin{aligned} d^2(\mathbf{S}, \tilde{\mathbf{S}}|\mathbf{h}) &= \sum_{m=1}^{w}\sum_{k=1}^{K} |y_{j_m K+k} - \tilde{y}_{j_m K+k}|^2 \\ &= \sum_{m=1}^{w}\sum_{k=1}^{K} |\alpha_{j_m K+k}\phi_{k,g_m}(s_{i_m} - \tilde{s}_{i_m})|^2, \end{aligned}$$

where $g_m = (i_m \bmod K) + 1$. Let $\delta = \min\{|s_i - s_j| : s_i \neq s_j\}$ and s_i, s_j are the constellation mapped symbols. Then clearly

$$d^2(\mathbf{S}, \tilde{\mathbf{S}}|\mathbf{h}) \geq \frac{\delta^2}{K} \sum_{m=1}^{w}\sum_{k=1}^{K} |\alpha_{j_m K+k}|^2. \tag{2}$$

Therefore, the pairwise error probability for this error event is bounded by :

$$P(\mathbf{S} \to \tilde{\mathbf{S}}|\mathbf{h}) \leq Q\left(\sqrt{\frac{\delta^2/K \sum_{m=1}^{w}\sum_{k=1}^{K} |\alpha_{j_m K+k}|^2}{2N_0}}\right) \tag{3}$$

where $Q(x) = (1/\sqrt{2\pi})\int_x^\infty e^{-t^2/2}dt$ with $x \geq 0$. To further simplify (3), we use the Chernoff Bound, $Q(x) \leq (1/2)e^{-x^2/2}$. Since $\{\alpha_{j_m K+k}\}$ are independent and Rayleigh distributed, we can average out to get the bound:

$$P_{pep}^{w} = P(\mathbf{S} \to \tilde{\mathbf{S}}) \leq \frac{1}{2}\left(1 + \frac{\delta^2/K}{4N_0}\right)^{-wK} \tag{4}$$

where P_{pep}^{w} is the pairwise error probability between two coded sequences which differ by weight w. Using the union bound the total error probability can be bounded by the sum of the error probabilities of every individual error event. Thus, the probability of an error event P_e can be bounded as

$$P_e \leq \sum_{w=d_{free}}^{\infty} \frac{1}{2} A_w \left(1 + \frac{\delta^2/K}{4N_0}\right)^{-wK}, \quad (5)$$

where A_w is the multiplicity of error events of weight w. Thus

$$P_e \leq A(Z)|_{Z=(1+\frac{\delta^2/K}{4N_0})^{-K}}, \quad (6)$$

where $A(Z) = \sum_{w=d_{free}}^{\infty} A_w Z^w$ is the error event weight enumerating function. Similarly let $B(W, D) = \sum_{w=d_{free}}^{\infty} \sum_{d=1}^{\infty} B_{w,d} W^w D^d$ be the bit-error weight enumerating function, with $B_{w,d}$ is the total number of divergent paths that constitute an error event of weight w, which are generated by d information bits different from the all zero bit sequence. Following this notation the bit error rate P_b can be bounded as in [2]

$$P_b \leq \sum_{w=d_{free}}^{\infty} \sum_{d=1}^{\infty} \frac{d}{k_0} B_{w,d} P_{pep}^{w}. \quad (7)$$

Therefore, if we define $B_w = \sum_{d=1}^{\infty} (d/k_0) B_{w,d}$ (This quantity is known as *bit multiplicity* and the pairs (B_w, w) form the *bit distance spectrum* of the convolution code; see [2]) and the function $B(Z) = \sum_{w=d_{free}}^{\infty} B_w Z^w$, the bit error rate can also be bounded as

$$P_b \leq B(Z)|_{Z=(1+\frac{\delta^2/K}{4N_0})^{-K}}. \quad (8)$$

The bounds in (5) and (8) are tight at high SNR. More importantly it can be seen that the diversity of the system is $d_{free}K$, the *product* of the individual diversities achievable by the channel coding and the unitary precoding. It means that the slope of the BER rate at high SNR can be increased tremendously by addition of small size precoders. In order to achieve the diversity of the joint scheme, ML decoding is required, which may be computationally demanding. However, sub-optimal iterative (turbo) decoding scheme is available with reasonable complexity.

A few remarks are in order. Firstly, the analysis relies on the assumption that the multiplicative coefficients $\{\alpha_n\}$ are independent for error events of low weight w. Indeed, a large size of Π_2 will decorrelate the equivalent channels. The separation induced by this interleaver must be more than the coherence time of the channel. However, for the case when the channel is varying very slowly, one cannot use interleavers of large size, for hardware implementation, or, delay reasons. In this case we can use a block or a random interleaver Π_1 of moderate size, which sends the error events to as widely spaced precoded blocks as possible. The errors are then mapped to independent subchannels by the use of the second interleaver Π_2. The reason that the precoder size is chosen to be 2 or 4 is twofold. Firstly, it facilitates low complexity decoding. Secondly, it has been shown in [10] that the gain in SNR introduced by precoding is limited to 3.5dB when $K \to \infty$, and a gain of 2.7dB is already obtained when $K = 4$. Thus, there is little need to consider a joint system employing a precoder of larger size.

C. Hard Decoding

Let $\mathbf{C}$ and $\tilde{\mathbf{C}}$ be two blocks that are precoded by the precoder Ψ_K. It was proved in [7] that the pairwise error probability that a sent block $\mathbf{C}$ is wrongly decoded as $\tilde{\mathbf{C}}$ is bounded by

$$P_{bler} = P(\mathbf{C} \to \tilde{\mathbf{C}}) \leq (G_c \gamma)^{-G_d}, \quad (9)$$

where G_c is the coding gain, G_d is the diversity gain of the system, γ is the SNR and P_{bler} is the block error rate. A hard decoder outputs the hard decisions on the coded bits for the equalization of the unitary precoding. These hard decisions are then fed to the channel decoder which uses ML decoding for deciding the information bits. We assume BPSK modulation, and let the probability of bit error due to the hard decisions be p_0. Clearly, $P_{bler} = 1 - (1 - p_0)^K$. The block error rate should be greater than the bit error rate. Thus from (9) we get, $p_0 \leq P_{bler} \leq (G_c\gamma)^{-G_d}$. On the other hand, every block error must have at least one bit error, so that bit error rate must be at least equal to P_{bler}/K. Hence, $p_0 \geq P_{bler}/K$. With a given p_0, the bit error rate is bounded as [2]

$$P_b \leq B(Z)|_{Z=\sqrt{4p_0(1-p_0)}}. \quad (10)$$

Thus, we can see that at high SNR for which we can neglect the term $(1 - p_0)$, the BER is bounded by the value of the input-output weight enumerating function $B(\cdot)$ at the value $\sqrt{4p_0}$. In fact, the bit error rate may be approximated at high SNR as $P_b \approx B_{d_{free}} 2^{d_{free}} p_0^{d_{free}/2}$. Since $p_0 \geq P_{bler}/K$ and bound in (9) is tight at high SNR, we get that the P_b is lower bounded by the value of the input-output weight enumerating function $B(\cdot)$ at the value $2\sqrt{G_c/K}\gamma^{-G_d/2}$. Comparing with (8) we can see that hard decoding incurs a loss in diversity by up to a factor of 2. The diversity in this case is at most $d_{free}K/2$. To approach optimal performance with reasonable complexity, we next develop an iterative (turbo) decoding scheme, which is based on passing soft extrinsic information between the equalizer, and the channel decoder.

III. Turbo Decoding of EC-UP-OFDM

Our receiver consists of FFT processing each noisy block of N samples obtained after discarding the cyclic prefix and sampling the receive filter output. The discrete-time baseband equivalent matrix-vector input-output relationship is given by:

$$\mathbf{r}_N = \mathbf{D}_H \Theta_N \mathbf{s}_N + \boldsymbol{\eta}_N, \qquad (11)$$

where $\mathbf{D}_H$ is the diagonal matrix consisting of the FFT coefficients of the channel taps, $\mathbf{s}_N$ is the transmitted symbol block of size N, and $\boldsymbol{\eta}_N$ is additive white Gaussian noise vector of covariance matrix $\sigma^2\mathbf{I}$. Thanks to our design, we can view (11) as the concatenation of N/K sub-blocks of K symbols each obeying:

$$\mathbf{r}_K^b = \mathbf{D}_{H,K}^b \Psi_K \mathbf{s}_K^b + \boldsymbol{\eta}_K^b, \qquad b \in [1, 2, \ldots, N/K]. \qquad (12)$$

Denote the i^{th} symbol of the symbol vector $\mathbf{s}_K^b$ as $s_K^b(i)$. Then the optimal equalizer outputs the *extrinsic information* about the symbol $s_K^b(i)$ as follows. With the definition of

$$f(\mathbf{s}_K^b) = \exp\left(\frac{-\|\mathbf{r}_K^b - \mathbf{D}_{H,K}^b \Psi_K \mathbf{s}_K^b\|^2}{2\sigma^2}\right), \qquad (13)$$

we calculate the extrinsic information for $s_K^b(i)$ as:

$$\lambda\{s_K^b(i) = s\} = \bar{c}_K^b \cdot \sum_{\mathbf{s}_K^b : s_K^b(i) = s} f(\mathbf{s}_K^b) \left(\prod_{j \neq i} P[s_K^b(j)]\right), \qquad (14)$$

where $P[s_K^b(j)]$ is the *A Priori Probability* (APrP) of the symbol $s_K^b(j)$, and $\bar{c}_K^b$ is the normalization constant. The extrinsic information is the residual or additional information that is gleaned out of the structure imposed by the equalizer, when the knowledge of the APrP's of the symbols are available. This approach to calculate the *extrinsic information* was also used in [9]. The extrinsic information calculated using (14) is de-interleaved and passed as APrP to a SISO channel decoding module which outputs soft information about the information and the coded bits. An optimal BCJR algorithm [1], or the suboptimal SOVA [5], may be employed. The channel decoder output in the form of extrinsic information is then fed back to the equalizer after interleaving, and the whole decoding process is repeated for a number of iterations.

IV. Simulated Performance

In our simulations, we compared the performance of CC-UP-OFDM with CC-OFDM [4]; and TC-UP-OFDM with TC-OFDM [6, 8]. We used BPSK modulation and $K = 4$. The simulations were run till either 1,000 errors were detected, or, over 500 channels and data points with at least 100 errors collected. For CC, we used a generator polynomial [133 171] in octal, while for TC we used a recursive systematic code with the same generator polynomial but having a feedback [133]. In both cases, the code was punctured to rate 3/4. We considered the HiperLAN2 channel B setup with $L = 15$ [3]. The channel is assumed to be constant over a time period of one OFDM symbol ($4\mu s$), but varies from symbol to symbol according to Jakes' fading model. We assume a carrier frequency of 5.2 GHz, and a terminal speed of 3m/s. The channel thus varies very slowly with the coherence time 3.4ms. We assume perfect knowledge of the channel at the receiver.

When the interleaver size is small, we did not observe much performance improvement by combining UP with CC. However, by employing a random interleaver, and increasing its size to 128×128 we observed significant performance improvement. We observe from Figure 2 that CC-UP-OFDM outperforms CC-OFDM considerably, with a gain of about 4dB at bit error rate of 10^{-3}. With such an interleaver, the total delay is calculated as $128 \times 128 \times 50 \times 1.25 \approx$ 1ms, which is tolerable for most applications. On the other hand, a block interleaver of dimension 128×128 does not offer much turbo gain. This is because the errors were spread by a distance of only 128 precoded symbols. Thus separation by 128 precoded symbols, i.e., just 2 OFDM symbols cannot guarantee independent channels. However the random interleaver has a good spreading factor, and hence the precoded symbols propogate through channels that are less correlated. The performance improvement of CC-UP-OFDM comes with a moderate increase of decoding complexity, which is about 5 times as complex as that of MAP decoding of CC-OFDM.

In Fig. 3, we replace the CC by a TC, to show that turbo coded OFDM can also benefit from unitary precoding. At a BER of 10^{-3}, TC-UP-OFDM outperforms TC-OFDM by about 3.3dB. Unitary precoding thus dramatically improves the system performance for coded OFDM systems.

V. Conclusions

In this paper we proposed and analyzed a novel scheme that combines *coding* and *constellation precoding* with an aim to reap the benefit of both. The diversity of the resulting system was seen to be the product of the individual diversities of the component coder and precoder, while the decoding complexity of the system was just a linear multiple of the sum of their individual complexities. This allows us to design systems that have high diversity, but with decoding complexity much less than purely coded systems or purely precoded systems having the same diversity. We further showed that hard decoding entails a loss in diversity by a factor of 2. Finally, we have established through simulations that EC-UP-OFDM can achieve a much better performance than EC-OFDM if sufficient interleaving is present.

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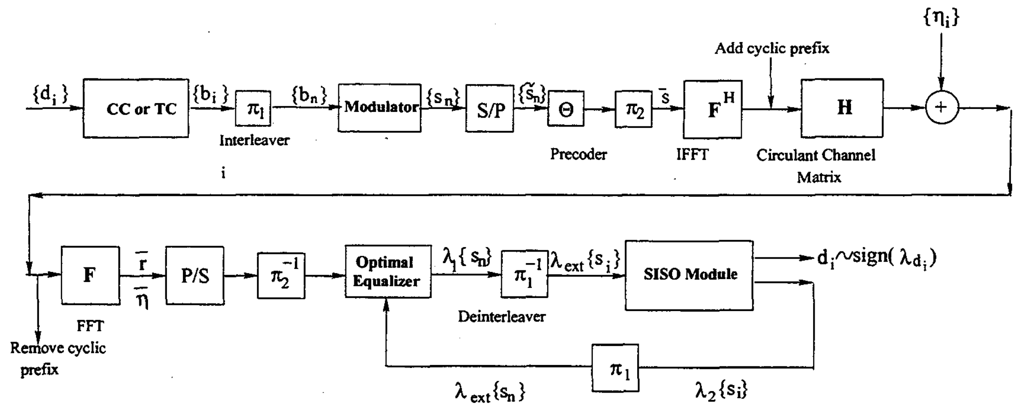

Fig. 1. Discrete-time baseband equivalent system model

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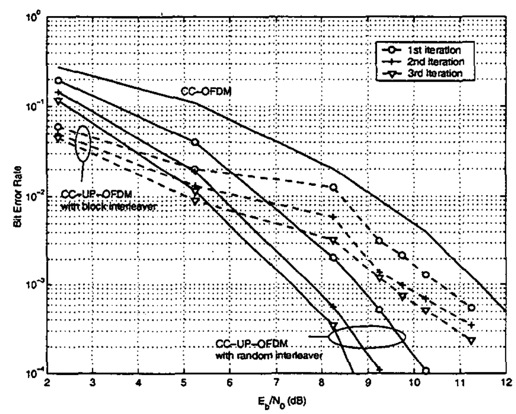

Fig. 2. CC-UP-OFDM vs. CC-OFDM

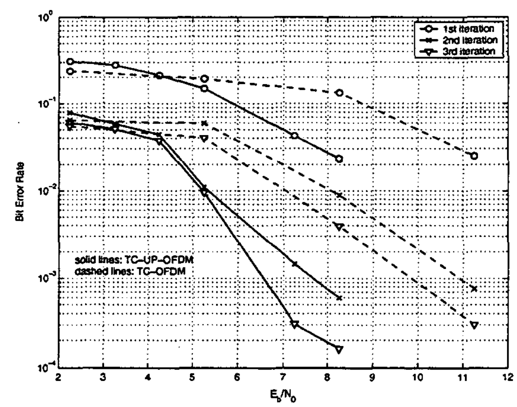

Fig. 3. TC-UP-OFDM vs. TC-OFDM