Iterative Frequency Domain Equalization for Single-Carrier Wideband MIMO Channels

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***Abstract*— Single carrier frequency domain equalization (SC-FDE) is receiving considerable attention recently due to its comparable complexity and performance with OFDM. In this paper, an iterative SC-FDE with decision feedback in frequency domain is proposed for wideband multiple-input multiple-output (MIMO) channels and more general multipath vector channels as a generalization of previous works considering FDE with decision feedback for single-input single-output (SISO) systems. The proposed detector is based on iterative forward and backward filtering followed by a MIMO multi-stream detector that uses a priori log-likelihood ratios (LLR) of coded symbols. Forward and feedback filters are jointly optimized according to the minimum mean square error (MMSE) criterion to minimize both self inter-symbol-interference (ISI) and interference from other streams transmitted at different antennas. It has been observed that the proposed structure exploits the multi-path diversity sources of the channel effectively and a performance very close to MIMO-OFDM outage probability can be achieved. Therefore, our proposed iterative SC-FDE technique for MIMO wideband channels can be viewed as a strong alternative to MIMO-OFDM schemes with similar complexity.**

I. Introduction

While OFDM based schemes are well-recognized candidates for broadband wireless technology, single-carrier (SC)-based technology has also started to gain considerable attention due to its comparable complexity with OFDM. It has been shown in [1] that frequency domain equalization (FDE) can readily be applied to SC transmission to yield similar performance to OFDM. Since OFDM suffers from high peak-to-average power ratio (PAPR) problem, SC techniques leading to more efficient use of power amplifiers are more suitable for uplink channels [1], [2]. Due to the attractive features of SC-FDE, it has been viewed as a strong alternative to OFDM-based systems recently and its importance is clear for wideband channels.

Block iterative FDE was proposed for uncoded single-input single-output (SISO) multipath channels in [3] and, block iterative FDE was considered in [4] together with channel decoding. Iterative equalization schemes for wideband MIMO channels was considered in [5]. They consider MMSE type forward filtering and successive interference cancellation (SIC) to mitigate the interference in time domain. Turbo equalization with MMSE type filtering in frequency domain was studied in [6], but it does not consider the use of decision feedback filters or SIC operation.

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The contribution of the paper is threefold. We first show that SC-FDE with both forward and backward filters can be generalized from SISO to vector channels, which includes MIMO as a special case. We, furthermore, derive the jointly optimal forward and backward filters in the frequency domain so that the complexity advantage of FDE is not compromised. Since reliability of coded symbols from the decoding process are used in deriving the optimal forward and backward filters, the filters employed in this work have a different structure from that of previous interference-cancellation-based MIMO turbo equalizers, such as in [7], [8] and [9]. Third, the MIMO wideband channel can be quasi-parallelized with the help of our proposed space-frequency equalizer and so, the code construction techniques achieving optimal rate-diversity tradeoff given by the singleton bound for block-fading channels [10], [11] can be effectively used such that the proposed equalization scheme combined with this type of coding structures yields a very close performance to the MIMO-OFDM outage probability and hypothetical matched filter bound (MFB) [12].

This paper is organized as follows. In Section II, the system model is described. In Section III, iterative frequency domain equalization techniques with frequency domain decision feedback generalized to vector channels are discussed in detail. In Section IV, asymptotic performance analysis of the proposed iterative FDE is done. Finally, the code construction techniques, simulation results and concluding remarks are presented in Section V and Section VI, respectively.

II. System Model

The following notation is used throughout the paper. Boldface upper-case letters denote matrices and scalars are denoted by plain lower-case letters. The superscript $(\cdot)^*$ denotes the complex conjugate for scalars and $(\cdot)^H$ denotes the conjugate transpose for vectors and matrices. The $n \times n$ identity matrix is shown with $\mathbf{I}_n$. The autocorrelation matrix for a random vector $\mathbf{a}$ is $\mathbf{R_a} = E\{\mathbf{a}\mathbf{a}^H\}$ where $E\{\cdot\}$ stands for the expected value operator. The $(i,j)^{th}$ element of a matrix $\mathbf{A}$ is denoted by $\mathbf{A}(i,j)$ and the i^{th} element of a vector $\mathbf{a}$ is denoted by a^i.

This paper considers block-based transmission as in [4] and [3]. During the transmission of one block, the channel is assumed to be constant and it changes independently from block to block. Without dealing with the channel estimation problem, the channel is assumed to be perfectly known at each block transmission. Cyclic prefix (CP) is used to prevent inter-block interference and enable frequency domain equalization with length larger than or equal to maximum channel length (L) as explained in [13]. The signal for a transmitted block with CP is a sequence of vectors: $[\mathbf{x}_0, \mathbf{x}_1, \ldots, \mathbf{x}_{N-1}, \mathbf{x}_0, \ldots, \mathbf{x}_{L-1}]$.

Assuming symbol rate sampling, the discrete time baseband equivalent model of the point-to-point MIMO wideband channel with n_r receive antennas and n_t transmit antennas can be written as [14],

$$\mathbf{y}_k = \sum_{l=0}^{L-1} \mathbf{H}_l \mathbf{x}_{k-l} + \mathbf{n}_k, \qquad k = 0, 1, \ldots, N-1, \tag{1}$$

where $\mathbf{H}_l$'s, $l = 0, \ldots, L-1$, are complex channel matrices comprised of independent zero-mean circularly symmetric complex Gaussian (ZMCSCG) random variables with variance given by the power delay profile of each channel [15]. Block fading model is considered and thus the channel matrices are assumed to be constant during a coherence interval significantly larger than a duration needed for the transmission of one block [13] and channel state information at transmitter (CSIT) is not available. Noise vectors $\mathbf{n}_k$'s are also taken as ZMCSCG white (spatially and temporally) noise with variance N_0. Only BPSK modulation is considered during the simulation studies. Extension to other M-ary or M-PSK modulations is straightforward.

If we define the DFT operation as $A_k = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} a_n e^{-j2\pi nk/N}$ for $k = 0, \ldots, N-1$, where a_n and A_k are the time domain sequence and its frequency domain sequence, respectively, then after the DFT operation to each element of $\mathbf{y}_k$ in (1), we can obtain the following expression in the frequency domain as done in [9]

$$\mathbf{Y}_k = \overline{\mathbf{\Lambda}}_k \mathbf{X}_k + \mathbf{N}_k, \qquad k = 0, \ldots, N-1 \tag{2}$$

where $\overline{\Lambda}_k$ is an $n_r \times n_t$ matrix representing the channel frequency response at the k^{th} tone with the entries [9]

$$\lambda_m^i(k) = \sum_{l=0}^{L-1} \mathbf{H}_l(i,m) e^{-j2\pi kl/N}, \tag{3}$$

for $i = 1, \ldots, n_r$ and $m = 1, \ldots, n_t$ and $\mathbf{H}_l(i,m)$ is a scalar and defined as the $(i,m)^{th}$ element of the channel matrix $\mathbf{H}_l$. The expression in (2) is the frequency domain equivalent of the channel in (1) and will be frequently used in the remainder of the paper. Also, DFT operation is performed by using $q_n^m = \frac{1}{\sqrt{N}} e^{-j2\pi mn/N}$ for $m, n = 0, 1, \ldots, N-1$ hereafter.

III. Iterative Frequency Domain Equalization for Wideband MIMO channels

We consider iterative frequency domain equalization (FDE) with frequency domain decision feedback in this paper. Since both equalization and decoding process can be performed in each iteration, turbo principle can be applied as done in [4], [16]. In Fig. 1, an exemplary receiver structure is shown for the frequency domain decision feedback (FDDF) case.

As it will be observed in Section V, the combined multipath and space enriched diversity of the channel is exploited by the proposed space-frequency equalizer effectively such that the performance obtained by the matched filter bound (MFB) [12] is approximately achieved.

The iterative frequency domain equalizer with hard and soft decision feedback in the frequency domain is studied in [2], [3] and [4] for the SISO systems. We derive the filter matrices based on the MMSE criterion like the FDE with time domain decision feedback (FDE-TDDF) case.

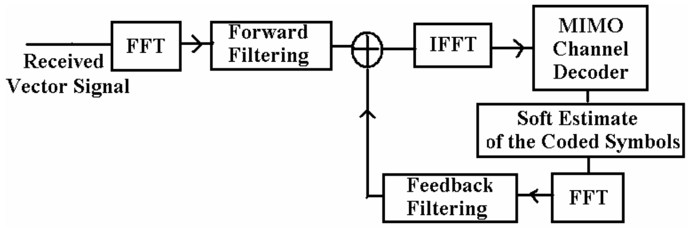

Fig. 1. FDE-FDDF: Iterative FDE with frequency domain decision feedback (FDDF)

Output of the FDE-FDDF for the k^{th} vector in the block (for the i^{th} iteration) can be expressed as,

$$\tilde{\mathbf{x}}_k^{(i)} = \sum_{j=0}^{N-1} (q_j^k)^* \left[(\mathbf{W}_j^{(i)})^H \mathbf{Y}_j - (\mathbf{C}_j^{(i)})^H \hat{\mathbf{X}}_j^{(i-1)} \right] \tag{4}$$

for $k = 0, \ldots, N-1$. $\mathbf{W}_j^{(i)}$'s and $\mathbf{C}_j^{(i)}$'s are forward and feedback filters both in frequency domain with sizes $n_r \times n_t$ and $n_t \times n_t$ respectively and $\hat{\mathbf{X}}_j^{(i-1)}$'s are the DFT's of soft decisions from the previous iteration.

It can be shown that one can find an optimum equalizer with time domain decision feedback which is equivalent to an equalizer with frequency domain decision feedback. The forward and backward filter matrices are jointly optimized and found according to the MMSE criterion in time domain given by $E\left\{\sum_{k=0}^{N-1} ||\tilde{\mathbf{x}}_k^{(i)} - \mathbf{x}_k||^2\right\}$ presented in [4], [3]. Since, the proposed FDE-FDDF and FDE-TDDF structures are equivalent, one can find a relation between time domain feedback filters $\mathbf{F}_j^{(i)}$ and frequency domain feedback filters $\mathbf{C}_j^{(i)}$. It can be shown that

$$(\mathbf{F}_k^{(i)})^*(m,n) = \sum_{l=0}^{N-1} (\mathbf{C}_l^{(i)})^*(m,n) e^{j2\pi kl/N} \tag{5}$$

for $m, n = 1, \ldots, n_t$ and $k = 0, \ldots, N-1$ where $(m,n)^{th}$ elements of $(\mathbf{C}_j^{(i)})^H$ and $(\mathbf{F}_j^{(i)})^H$ are defined as $(\mathbf{C}_j^{(i)})^*(m,n)$ and $(\mathbf{F}_j^{(i)})^*(m,n)$, respectively. Since the optimization problem for FDE-FDDF case is mathematically equivalent to FDE-TDDF with the constraint $\mathbf{F}_0^{(i)}(n,n) = 0,\ n = 1, \ldots, n_t$, we can set the constraint for frequency domain feedback filters from (5) as

$$(\mathbf{F}_0^{(i)})^*(n,n) = \sum_{l=0}^{N-1} (\mathbf{C}_l^{(i)})^*(n,n) = 0,\ n = 1, \ldots, n_t \tag{6}$$

With this constraint, one can avoid self-subtraction of the desired symbol by its previous estimate. The Lagrange multiplier method can be used to obtain optimal forward and backward frequency domain filters. Lagrangian vectors and the corresponding scalar constraints (Lagrangian function) can be written as

$$\mathbf{\Gamma}^{(i)} = \text{diag}\left[\Gamma_1^{(i)}, \ldots, \Gamma_{n_t}^{(i)}\right]_{(n_t \times n_t)},$$

$$Lagrangian(\mathbf{\Gamma}^{(i)}) = \sum_{n=1}^{n_t} \sum_{j=0}^{N-1} (\mathbf{C}_j^{(i)}(n,n))^* \Gamma_n^{(i)} \tag{7}$$

The mitigation of inter-stream interference originated from other antenna stream's spatial interference and inter symbol interference (ISI) resulted from frequency selectivity is done optimally with this structure. This differs from previous MIMO studies which use spatial interference suppression techniques based on successive interference cancellation (SIC).

Interleaving operation is used both in time and space, thus we can assume that,

$$E\{\mathbf{x}_k(\mathbf{x}_l)^H\} = E_s\mathbf{I}_{n_t}\delta_{kl}, \text{ for } k,l = 0,\ldots,N-1. \quad (8)$$

Some important correlation matrices used by the forward and feedback filters are defined for the i^{th} iteration as

$$\mathbf{P}^{(i)} = E\{\mathbf{x}_k(\hat{\mathbf{x}}_k^{(i-1)})^H\}, \quad \mathbf{B}^{(i)} = E\{\hat{\mathbf{x}}_k^{(i-1)}(\hat{\mathbf{x}}_k^{(i-1)})^H\} \quad (9)$$

for $k = 0,\ldots,N-1$. They can be found by using the soft feedback decisions, $\hat{\mathbf{x}}_k$'s, obtained from the decoder as done in [17]. The correlation matrices are updated in each iteration by using the soft information provided by the decoder. The forward and backward filters are shown to be independent of time index k and, so the block processing on each frequency bin can be implemented effectively.

After taking the gradient of the MMSE cost function and the Lagrangian with respect to the rows of $(\mathbf{W}_j^{(i)})^H$ and $(\mathbf{C}_j^{(i)})^H$, equating the gradients to the zero vector, taking expectations and combining vectors into single matrix equations for $n = 1,\ldots,n_t$, one can obtain the following matrix equations giving the optimal forward and backward filter matrices in the frequency domain

$$\mathbf{R}_{\mathbf{Y}_j}\mathbf{W}_j^{(i)} = \overline{\mathbf{\Lambda}}_j\left[E_s\mathbf{I}_{n_t} + \mathbf{P}^{(i)}\mathbf{C}_j^{(i)}\right] \quad (10)$$

$$\mathbf{B}^{(i)}\mathbf{C}_j^{(i)} = (\mathbf{P}^{(i)})^H\left[\overline{\mathbf{\Lambda}}_j^H\mathbf{W}_j^{(i)} - \mathbf{I}_{n_t}\right] - \mathbf{\Gamma}^{(i)} \quad (11)$$

for $j = 0,\ldots,N-1$, where

$$\mathbf{R}_{\mathbf{Y}_j} = E\{\mathbf{Y}_j(\mathbf{Y}_j)^H\} = \left(\overline{\mathbf{\Lambda}}_j\overline{\mathbf{\Lambda}}_j^H E_s + N_0\mathbf{I}_{n_r}\right) \quad (12)$$

and $\mathbf{\Gamma}^{(i)}$ can be obtained from the constraint

$$\sum_{j=0}^{N-1}\mathbf{C}_j^{(i)}(n,n) = 0,\ n = 1,\ldots,n_t. \quad (13)$$

By substituting $\mathbf{W}_j^{(i)}$'s into (11) and using the constraint (13), the Lagrangian terms given in (7) and backward filter matrices can be readily found after some calculus as

$$\Gamma_n^{(i)} = \frac{\left[\sum_{j=0}^{N-1}\mathbf{A}_j^{(i)}(n,:)\mathbf{D}_j^{(i)}(:,n)\right]}{\left[\sum_{j=0}^{N-1}\mathbf{A}_j^{(i)}(n,n)\right]},\ n = 1,\ldots,n_t \quad (14)$$

$$\mathbf{C}_j^{(i)} = \mathbf{A}_j^{(i)}\left[\mathbf{D}_j^{(i)} - \mathbf{\Gamma}^{(i)}\right], \quad (15)$$

where

$$\mathbf{A}_j^{(i)} = \left[\mathbf{B}^{(i)} - (\mathbf{P}^{(i)})^H\overline{\mathbf{\Lambda}}_j^H\mathbf{R}_{\mathbf{Y}_j}^{-1}\overline{\mathbf{\Lambda}}_j\mathbf{P}^{(i)}\right]^{-1}, \quad (16)$$

$$\mathbf{D}_j^{(i)} = (\mathbf{P}^{(i)})^H\overline{\mathbf{\Lambda}}_j^H\mathbf{R}_{\mathbf{Y}_j}^{-1}\overline{\mathbf{\Lambda}}_jE_s - (\mathbf{P}^{(i)})^H, \quad (17)$$

and $\mathbf{A}_j^{(i)}(n,:)$ is the $n-$th row of $\mathbf{A}_j^{(i)}$, $\mathbf{D}_j^{(i)}(:,n)$ is the $n-$th column of $\mathbf{D}_j^{(i)}$. Forward filters $\mathbf{W}_j^{(i)}$'s are obtained from (10) for $j = 0,\ldots,N-1$.

The computational complexity to obtain forward and feedback filters is considerably reduced for FDE-FDDF case, since only $n_r \times n_r$ and $n_t \times n_t$ matrix inversions are needed as can be seen from (14)-(17) and size of these matrices is independent of block length (N) just like OFDM-based systems.

As to the decoding stage, well known turbo decoding idea is used. The equalizer and decoder iteratively exchange soft information in terms of likelihood values of the transmitted data to improve their performance. The soft-in soft-out decoder produces likelihood information of each coded bit and it can be in the form of a convolutional, block or spacetime trellis decoder depending on the encoding structure. The equalizer coefficients are updated by using the likelihood information of transmitted data given by the decoder at each iteration.

IV. Asymptotic Performance Analysis of FDE-FDDF

At each iteration, forward and feedback filters approach the optimal coefficients in case of perfect feedback with the help of improved log a-posteriori probability (APP) ratio of each coded symbol obtained from the decoder. At later iterations, feedback decisions become more and more reliable and correlation matrices approach the asymptotic values: $\mathbf{P}^{(i)} \to E_s\mathbf{I}_{n_t}$ and $\mathbf{B}^{(i)} \to E_s\mathbf{I}_{n_t}$. Signal-to-interference-noise-ratio's (SINR) of each parallelized channel after equalization are evaluated for the asymptotic case and given as

$$SINR_m = \sum_{l=0}^{L-1}\sum_{i=1}^{n_r}|H_l(i,m)|^2\frac{E_s}{N_0}, \quad \text{for } m = 1,\ldots,n_t \quad (18)$$

The complete work can be found in ([18], Appendix).

It is seen from (18) that one can achieve the full diversity gain $(n_r \times L)$ at each of the parallelized channels. If transmit diversity schemes in the form of coding across antennas such as universal space-time codes [19] or other coding-multiplexing based techniques [11] are utilized, the maximum potential diversity gain of $(n_r \times n_t \times L)$ can be achieved by the proposed equalization scheme here.

It is well known that the SC-MMSE receiver reduces to a channel matched filter if the perfect a priori information of all transmitted symbols leading to ISI and inter-stream interference are available at the receiver and all the interference is cancelled [16]. Therefore, an upper bound to the packet error rate (PER) referred to as the matched filter bound (MFB) of the receiver can be obtained by assuming perfect decision feedback [12]. As long as the channel is not in outage, the feedback decisions approach to their true values and thus, the proposed receiver attains a very close performance to MFB.

V. Simulation Results

A. Outage Probability and MFB Calculations

In this section, we will compare the performance of our proposed equalizer with the hypothetical MFB performance and the corresponding constrained outage probability of MIMO-OFDM system. The constrained capacity can be found for the system model in (2) given the complex vector set χ of cardinality M^{n_t} (e.g., M-ary or M-PSK modulations) similar to the derivations for block fading channels in [11] as

$$C^{\chi}_{MIMO-OFDM} = \frac{1}{N}\sum_{j=0}^{N-1} I(\mathbf{X}_j; \mathbf{Y}_j|\overline{\mathbf{\Lambda}}_j) =$$
$$n_t \log_2(M) - \frac{1}{N}\sum_{j=0}^{N-1} E_{\mathbf{N}_j}\{\sum_{\mathbf{X}_k\in\chi} \frac{1}{M^{n_t}} \log_2 \sum_{\mathbf{X}_i\in\chi}$$
$$\exp\left(\frac{-\|\overline{\mathbf{\Lambda}}_j(\mathbf{X}_k - \mathbf{X}_i) + \mathbf{N}_j\|^2 + \|\mathbf{N}_j\|^2}{N_0}\right)\} \quad (19)$$

and the corresponding outage probability can be written as

$$P_{out}^{MIMO-OFDM,\,\chi}(R) = \mathbb{P}\left\{C^{\chi}_{MIMO-OFDM} < R\right\}. \quad (20)$$

Constrained outage probability and the MFB will be used for performance evaluation in the next part. We used powerful channel codes in our simulations to get a close performance to channel capacity, but similar observations are obtained when less powerful or high rate codes are applied. Again, the proposed receiver operates very close to the hypothetical MFB since this structure uses jointly optimal forward and feedback filtering in each equalizer iteration. Due to space limitations, we can not include these results.

B. Code Construction and Performance Results

The code construction used in our work is similar to the structure for random-like codes adapted to the block-fading channel based on blockwise concatenation and on bit-interleaved coded modulation (BICM) in [11]. The presented coded modulation construction in [11] systematically yields singleton-bound achieving turbo-like codes defined over an arbitrary signal set. As such, any other coding architecture that performs well in parallel block fading channels can be used in our system. We have used the same encoding and decoding structures as in [11] in simulations.

Here, concatenated convolutional codes are used. The outer code is a simple repetition code of rate $r = 1/n_t$ and the inner codes are rate-1 accumulators, which is referred to as repeat and blockwise accumulate (RBA) code [11]. Fig. 2 shows the performance of the proposed FDE-FDDF for a 4×4 MIMO system with the use of full block diversity attaining RBA code of rate $r = 1/4$. Channel model described in Section II is assumed and COST207 channel with exponential power delay profile for suburban and urban areas [15] is used (with 7 μ sec delay spread). BPSK modulation is used for simplicity, but other M-ary or M-PSK modulations combined with BICM [20] can be applied to our proposed structure. Symbol duration is taken as 1 μ second, and the channel length L equals 8. The first channel tap is taken to possess unity power. The information block length, i.e., the information bits entering the outer encoder is taken as $K = 250$, then the block length N is equal to $K/(r \cdot n_t) + 1 = 251$ including termination bits. Number of iterations inside the Turbo RBA decoder is set to 10 and the number of equalizer iterations at which the forward and backward filters are updated by using the reliability matrices is taken as 3.

It is seen from Fig. 2 that the performance of FDE-FDDF is 0.3 dB away from MFB. There is approximately 1.5 dB difference between the outage probability of the MIMO-OFDM at rate $R = n_t \cdot r = 1$ bits/sec/Hz and this gap from the outage is similar to the gaps obtained with RBA in parallel block fading channels in [11]. Then, one can say that ISI, spatial interference and, the error propagation problem in decision feedback are almost eliminated, since the perfect decision feedback performance (MFB) is approximately achieved. Moreover, it is seen that the performance of FDE-FDDF shows the same slope as MIMO-OFDM outage and so it is possible to attain the maximum diversity of the MIMO broadband channel by using the proposed space-frequency equalizer and coding across transmit antennas. Furthermore, SC-FDE-based schemes could be a promising candidate for wideband MIMO systems as an alternative to MIMO-OFDM schemes and if one takes the loss due to PAPR problem in OFDM based systems into consideration, the performance difference between SC-FDE-based MIMO schemes and the MIMO-OFDM systems will be more significant.

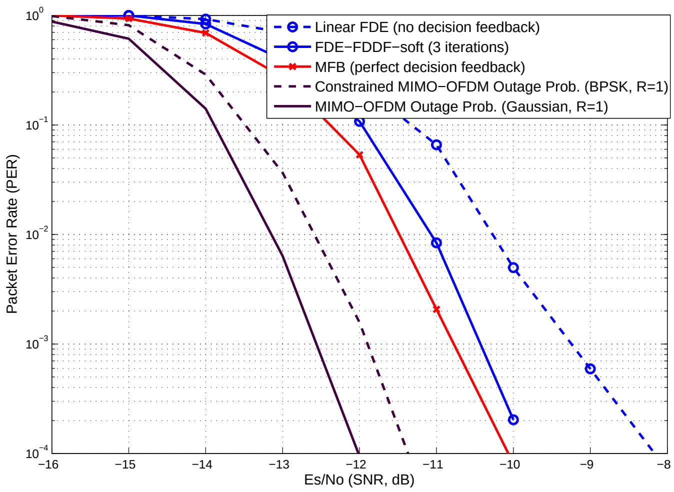

Fig. 2. Performance comparison of SC-FDE with MFB and outage for 4×4 MIMO, $T_s = 1\mu$ sec and COST 207 suburban channel

In Fig. 3, simulation results are depicted for code rate, $r = 1/2$. A full block diversity attaining blockwise concatenated code (BCC) is used for encoding as adapted from [11]. The outer code is a rate$-\frac{1}{2}$ convolutional code and the inner codes are n_t trivial rate-1 accumulators. The information block length K is taken as 248. Similar results are obtained and a close performance to MIMO-OFDM outage at rate $R = n_t \cdot r = 2$ bits/sec/Hz is achieved within 2 dB.

Our proposed SC-FDE can also be applied to classical SISO ISI channels. In Fig. 4, we compared the performance of iterative SC-FDE-FDDF-soft feedback with that of the outage of an OFDM scheme. A convolutional encoder with $r = 1/2$ serially concatenated (SC) with a rate-1 accumulator is used for information block length $K = 123$. At first glance, it is surprising to note that constrained OFDM outage probability is surpassed by the iterative FDE-FDDF, but as stated in [21] the

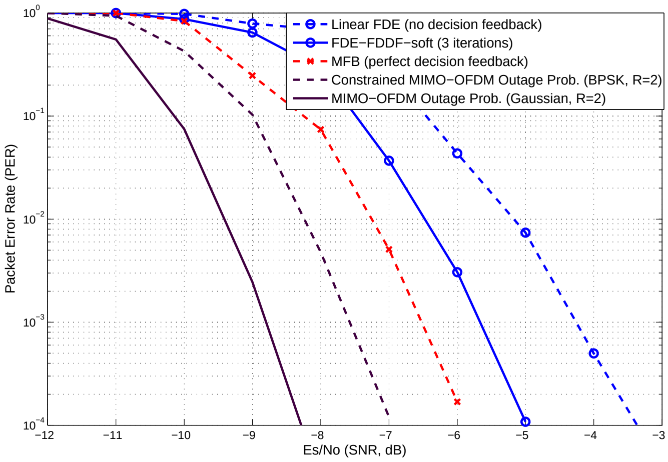

Fig. 3. Performance comparison of SC-FDE with MFB and outage for 4×4 MIMO, $T_s = 1\mu$ sec and COST 207 suburban channel

capacity of wideband channels under non-Gaussian alphabets is an open problem and OFDM is not the capacity achieving scheme for non-Gaussian input alphabets.

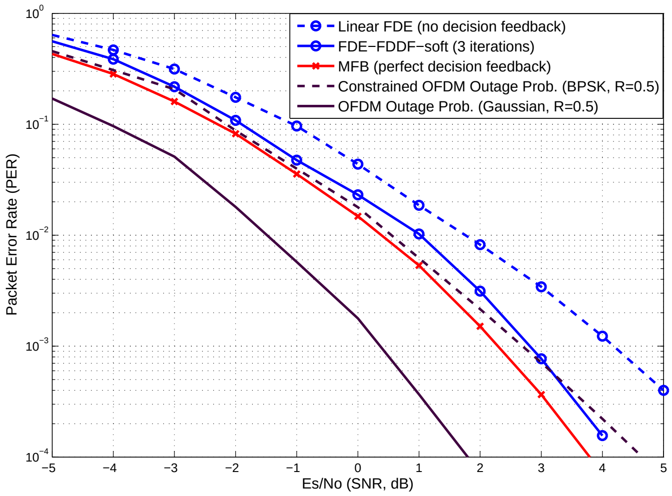

Fig. 4. Performance comparison of SC-FDE with MFB and outage for SISO system, $T_s = 0.5\mu$ sec, COST 207 typical suburban exponential channel, $L = 15$

Furthermore, it is interesting to note that the performance improvement of the FDE-FDDF scheme over the linear FDE without decision feedback is about 2 dB at PER=0.0001 for all simulation results. There is also a loss in diversity as observed in the reduced PER slope without decision feedback. One can say that the proposed space-frequency equalizer gains more diversity in comparison to linear FDE by a careful design of both the forward and backward filters.

VI. Conclusion

In this paper, we extended the SC-FDE mechanism from SISO channels to more general vector-based models which include MIMO as a special case. We have also shown that capacity-achieving jointly optimal forward and backward filtering operations can be effectively performed in the frequency domain. It is observed that error performance close to the outage probability can be attained by careful coding across transmit antennas without compromising computational complexity. Therefore, our proposed iterative SC FDE technique for MIMO wideband channels can be viewed as a strong alternative to MIMO-OFDM schemes with similar complexity. Future studies will include the effect of channel estimation error on the performance of vector channel SC-FDE systems.

References

[1] D. Falconer, S. L. Ariyavisitakul, A. Benyamin-Seeyar, and B. Eidson, "Frequency domain equalization for single-carrier broadband wireless systems," *IEEE Commun. Mag.*, vol. 40, no. 4, pp. 58–66, April 2002.

[2] N. Benvenuto and S. Tomasin, "On the comparison between OFDM and single carrier modulation with a DFE using a frequency-domain feedforward filter," *IEEE Trans. Commun.*, vol. 50, pp. 947–955, June 2002.

[3] N. Benvenuto and S. Tomasin, "Iterative design and detection of a DFE in the frequency domain," *IEEE Trans. Commun.*, vol. 53, no. 11, pp. 1867–1875, Nov. 2005.

[4] B. Ng, C. Lam, and D. Falconer, "Turbo frequency domain equalizer for single-carrier broadband wireless systems," *IEEE Trans. Wireless Comm.*, vol. 6, no. 2, pp. 759–767, Feb. 2007.

[5] K.Kansanen and T. Matsumoto, "A computationally efficient MIMO turbo-equaliser," *Proc. IEEE VTC*, vol. 1, pp. 277–281, 2003.

[6] M. S. Yee, M. Sandell, and Y. Sun, "Comparison study of single carrier and multi-carrier modulation using iterative based receiver for MIMO system," *Proc. IEEE VTC*, pp. 1275–1279, 2004.

[7] J. Karjalainen, N. Veselinovic, K. Kansanen, and T. Matsumoto, "Iterative frequency domain joint-over-antenna detection in multiuser MIMO," *IEEE Trans. Wireless Comm.*, vol. 6, no. 10, pp. 3620–3631, Oct. 2007.

[8] R. Visoz, A. O. Berthet, and S. Chtourou, "Frequency-domain block turbo-equalization for single-carrier transmission over MIMO broadband wireless channel," *IEEE Trans. Commun.*, vol. 54, no. 12, pp. 2144–2149, Dec. 2006.

[9] Y. Zhu and K. B. Letaief, "Single-carrier frequency-domain equalization with noise prediction for MIMO systems," *IEEE Trans. Commun.*, vol. 55, no. 5, pp. 1063–1076, May 2007.

[10] R. Knopp and P. Humblet, "On coding for block fading channels," *IEEE Trans. Inform. Theory*, vol. 46, no. 1, pp. 189–205, Jan. 2000.

[11] A. G. Fabregas and G. Caire, "Coded modulation in the block-fading channel: Coding theorems and code construction," *IEEE Trans. Inform. Theory*, vol. 52, no. 1, pp. 91–114, Jan. 2006.

[12] J. R. Barry, E. A. Lee, and D. G. Messerschmitt, *Digital Communication*, Springer, Third Edition.

[13] A. Goldsmith, *Wireless Communications*, Camridge University Press, 2005.

[14] E. Biglieri, R. Calderbank, A. Constantinides, A. Goldsmith, A. Paulraj, and H. Vincent Poor, *MIMO Wireless Communications*, Cambridge University Press, 2007.

[15] J. G. Proakis, *Digital Communications*, McGRAW-HILL, 2001.

[16] X. Wang and H. V. Poor, "Iterative (turbo) soft interference cancellation and decoding for coded CDMA," *IEEE Trans. Commun.*, vol. 47, pp. 1046–1061, July 1999.

[17] G. M. Guvensen and A. O. Yilmaz, "Iterative decision feedback equalization and decoding for rotated multidimensional constellations in block fading channels," *Proc. IEEE VTC*, vol. 1, pp. 1–6, April 2009.

[18] G. M. Guvensen, *Near capacity operating practical transceivers for wireless fading channels*, Thesis (M.S.): Middle East Technical University, Ankara, 2009.

[19] D. Tse and P. Viswanath, *Fundamentals of Wireless Communication*, Cambridge University Press, 2005.

[20] G. Caire, G. Taricco, and E. Biglieri, "Bit-interleaved coded modulation," *IEEE Trans. Inform. Theory*, vol. 44, no. 3, pp. 927946, May 1998.

[21] P. O. Vontobel, A. Kavcic, D. M. Arnold, and H. A. Loeliger, "A generalization of the blahutarimoto algorithm to finite-state channels," *IEEE Trans. Inform. Theory*, vol. 54, no. 5, pp. 1887–1918, May 2008.