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Performance Analysis of Subspace Methods Used in Blind Channel Identification

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Abstract –– A subspace-based method is proposed for estimating the channel responses of single-input-multiple-output (SIMO) Orthogonal Frequency Division Multiplexing (OFDM) system. Our technique relies on minimum noise subspace (MNS) decomposition to obtain noise subspace in a parallel structure from a set of pairs (combinations) of system outputs that form a properly connected sequence (PCS). The developed MNS-OFDM estimator is more efficient in computation than subspace (SS)- OFDM estimator, although the former is less robust to noise than the later. To maximise the MNS-OFDM estimator performance, a symmetric version of MNS is implemented. We present simulation results demonstrating the channel identification performance of the corresponding OFDM-based SIMO systems employ cyclic prefixing approach.

IndexTerms ––– Blind Channel Identification, Equalisation, MNS, OFDM

1 INTRODUCTION

OFDM is a multi-carrier digital modulation technique that facilitates the transmission of high data rates with a limited bandwidth. It is an effective technique for several applications such Digital Audio Broadcasting (DAB) and terrestrial Digital Video Broadcasting (DVB). In addition, OFDM forms the basis for the physical layer in upcoming standards for broadband Wireless Local Area Network (WLAN) , i.e. ESTI-BRAN HiperLAN/2 , IEEE 802.11a and Multimedia Mobile Access Communication Systems (MMAC) and for Fourth Generation (4G) broadband wireless systems that will perform multimedia transmission to mobiles and portable personal communications devices, i.e. European MEMO project and for IEEE 802.16.

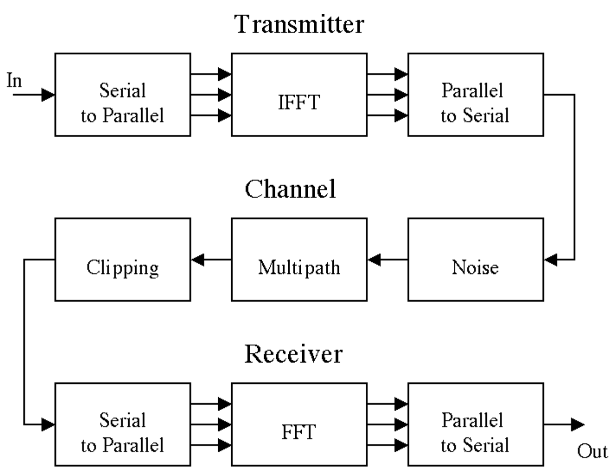

Fig 1- Block Diagram Of OFDM

The transmitter first converts the input data from a serial stream to parallel sets. Each set of data contains one symbol, Si, for each subcarrier. For example, a set of four data would be [S0 S1 S2 S3].

Before performing the Inverse Fast Fourier Transform (IFFT), this example data set is arranged on the horizontal axis in the frequency domain as shown in Figure 2. This symmetrical arrangement about the vertical axis is necessary for using the IFFT to manipulate this data.

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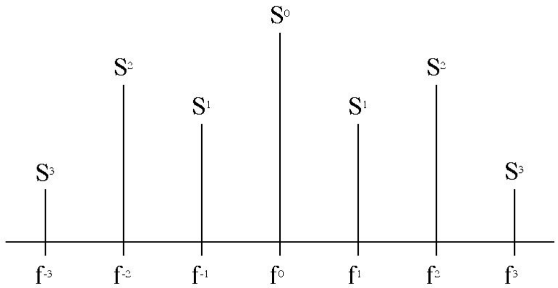

Fig 2- Frequency Domain Distribution Of Symbols

An inverse Fourier transform converts the frequency domain data set into samples of the corresponding time domain representation of this data. Specifically, the IFFT is useful for OFDM because it generates samples of a waveform with orthogonal frequency components.Then, the parallel to serial block creates the OFDM signal by sequentially outputting the time domain samples.

The channel simulation will allow examination of the effects of noise, multipath, and clipping. By adding random data to the transmitted signal, simple noise can be simulated. Multipath simulation involves adding attenuated and delayed copies of the transmitted signal to the original. This simulates the problem in wireless communication when the signal propagates on many paths. For example, a receiver may see a signal via a direct path as well as a path that bounces off a building. Finally, clipping simulates the problem of amplifier saturation. This addresses a practical implementation problem in OFDM where the peak to average power ratio is high.

The receiver performs the inverse of the transmitter. First, the OFDM data are split from a serial stream into parallel sets. The Fast Fourier Transform (FFT) converts the time domain samples back into a frequency domain representation. The magnitudes of the frequency components correspond to the original data. Finally, the parallel to serial block converts this parallel data into a serial stream to recover the original input data.

Due to increase in the normalised delay spread, multipath fading becomes a major concern as systems with high data rate are more liable to intersymbol interference (ISI). Classically, ISI is eliminated by employing a cyclically extended time domain guard interval (GI). Thus, each OFDM symbol is preceded by a periodic extension of the symbol itself. This GI is also known as cyclic prefix (CP) and the system CP-OFDM. Recently, zero-padding OFDM (ZP-OFDM), which pre-pends each OFDM symbol with zeros rather than replicating the last few samples, has been proposed. ZP-OFDM not only has all the advantages of the CP-OFDM, but also guarantees symbol recovery and ensures finite impulse response (FIR) equalisation. However, the implementation of a ZP-OFDM system involves transmitter modifications and complicates the equalizer.

To maximise the performance advantage of OFDM system, reliable identification of Single Input Multiple Output (SIMO) channels is desired. Currently, the channel identification and equalisation technique used requires a major fraction of the channel capacity to send a training sequence over the channels. There are practical situations where it is not feasible to utilize a training sequence such as in fast varying channels. To save this fraction of channel capacity, blind identification is an attractive approach. Using the blind channel identification techniques, the OFDM-based SIMO receiver can identify the channel characteristics and equalises the channel all based on the received signal, and no training sequence is needed, which hence saves the channel capacity.

Blind identification and equalisation of SIMO channels have been a very active area of research during the past few years. Among the various known algorithms, Second Order Statistics (SOS)-based algorithms are the most attractive due to their special properties. It was, for a while, believed that the subspace (SS)-based method was the only key to the surprising success among the existing SOS-based techniques. The SS-based method applies the MUltiple SIgnal Classification (MUSIC) concept to a relation between the channel impulse responses and the noise subspace associated with a covariance matrix of the system output.

One of the important advantages of SS-based method is its deterministic property. That is, the channel parameters can be recovered perfectly in the absence of noise, using only a finite set of data samples, without any statistical assumptions over the input data. More recently, the use of the SS-based method has been suggested to accomplish blind SIMO channel identification in OFDM systems. Despite their high identification efficiency, SS-based methods are computationally very intensive, which may be unrealistic or too costly to implement in real time, especially for large sensor array systems. The main reason is that they require non-parallelisable eigen-value-decomposition (EVD) of a large dimensional matrix to extract (estimate) the noise or signal subspace.

In this paper, using a minimum noise subspace (MNS) decomposition concept, we introduce several techniques for blind identification and equalisation of OFDM-based SIMO systems. Our techniques compute the noise subspace via a set of noise vectors (basis of the noise subspace) that can be computed in parallel from a set of pairs (combinations) of system output, without using reference or pilot symbols. Therefore, an EVD for smaller covariance matrices is required to extract noise subspace. Ideally, this approach, which relies on the known structure of the received OFDM symbols, provides a perfect channel estimate in the absence of noise. It is believed to have inspired all the subsequent developments which have taken place to accomplish unknown parameter identifications in a wide range array signal processing applications. Furthermore, the developed techniques significantly reduce receiver complexity in wireless broadband multi-antenna systems.

2 PROPOSED WORK

In this section, OFDM-based SIMO is introduced by using CP and ZP techniques.

2.1 Standard CP-OFDM System

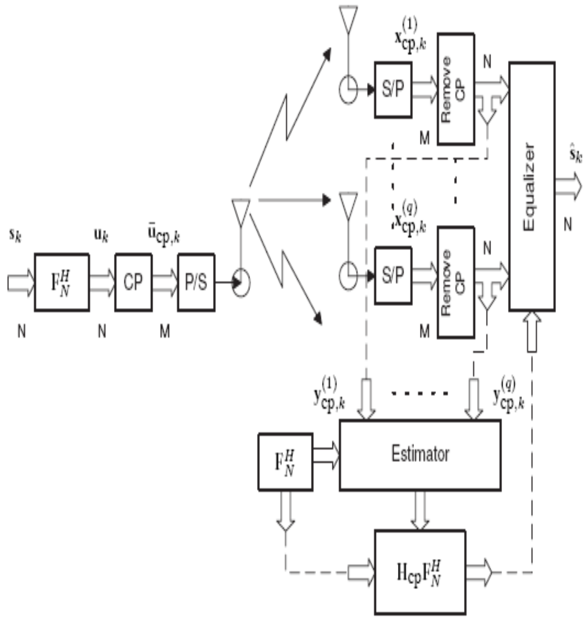

Fig. 3 - CP-OFDM system: transmitter and receiver.

Figure 3 depicts the baseband discrete-time block equivalent model of a standard CP-OFDM system. The transmitted symbols are parsed into blocks of size N: $\mathbf{s}_k = [s_k(0), s_k(1), \ldots, s_k(N-1)]^T$ where k = 0, 1, 2 . . .,K − 1. The elements of $\mathbf{s}_k$ are considered to be independent and identically distributed (i.i.d). We regard these elements to be in the frequency domain. The symbol block $\mathbf{s}_k$ is then modulated and converted into time domain using the IFFT matrix $\mathbf{F}^H{}_N$, where $\mathbf{F}_N$ has entires

$$f_{n,d} = \frac{1}{N}\exp\left(\frac{j2\pi nd}{N}\right)$$

and d, n = 0, . . . , N − 1. The data vector $\mathbf{u}_k = \mathbf{F}^H{}_N\mathbf{s}_k = [u_k(0), u_k(1), \ldots, u_k(N-1)]^T$ is then appended with a CP of length L, resulting in a size M = N + L signal vector: $\mathbf{u}_{cp,k} = \mathbf{T}_{cp}\mathbf{u}_k = [u_k(N-L), \ldots, u_k(N-1), u_k(0), \ldots, u_k(N-1)]^T$. We consider $\mathbf{T}_{cp}$ is a concatenation of the last L rows of an N×N identity matrix $\mathbf{I}_N$(that we denote as $\mathbf{I}cp$) and the identity matrix itself $\mathbf{I}_N$, i.e., $\mathbf{T}_{cp} = [\mathbf{I}^T{}_{cp}, \mathbf{I}^T{}_N]^T$. CP makes the OFDM appear periodic over the time span of interest. The channel response is denoted by $h^{(r)}(l)$ where l = 0, 1, . . . , $L_{(r)}$, and r = 1, 2, . . . , q. To avoid ISI, as indicated previously, the CP length L is selected to be equal to or greater than the channel order, i.e., $L_{(r)} \le L$. We consider the upper bound of the SIMO channel order $L_{(r)}$ as a CP length L. The received k-th block at r-th output for n = 0, 1, . . .,M − 1, is given by

$$x^{(r)}_{\mathrm{cp},k}(n) = \sum_{l=0}^{L} h^{(r)}(l)\, u_{\mathrm{cp},k}(n-l) + v^{(r)}_{\mathrm{cp},k}(n) \tag{1}$$

where $u_{cp,k}(n-l)$ and the AWGN, $v^{(r)}{}_{cp,k}(n)$, is assumed to be mutually uncorrelated and stationary. Using the following notations

$$\begin{aligned} \mathbf{x}_{\mathrm{cp},k}(n) &= \left[x^{(1)}_{\mathrm{cp},k}(n), x^{(2)}_{\mathrm{cp},k}(n), \ldots, x^{(q)}_{\mathrm{cp},k}(n)\right]^T \\ \mathbf{v}_{\mathrm{cp},k}(n) &= \left[v^{(1)}_{\mathrm{cp},k}(n), v^{(2)}_{\mathrm{cp},k}(n), \ldots, v^{(q)}_{\mathrm{cp},k}(n)\right]^T \\ \mathbf{h}(l) &= \left[h^{(1)}(l), h^{(2)}(l), \ldots, h^{(q)}(l)\right]^T \end{aligned} \tag{2}$$

we can rewrite the input-output relation (1) in vector matrix as

$$\mathbf{x}_{\mathrm{cp},k}(n) = \sum_{l=0}^{L} \mathbf{h}(l)\, u_{\mathrm{cp},k}(n-l) + \mathbf{v}_{\mathrm{cp},k}(n). \tag{3}$$

2.2 ZP-OFDM System

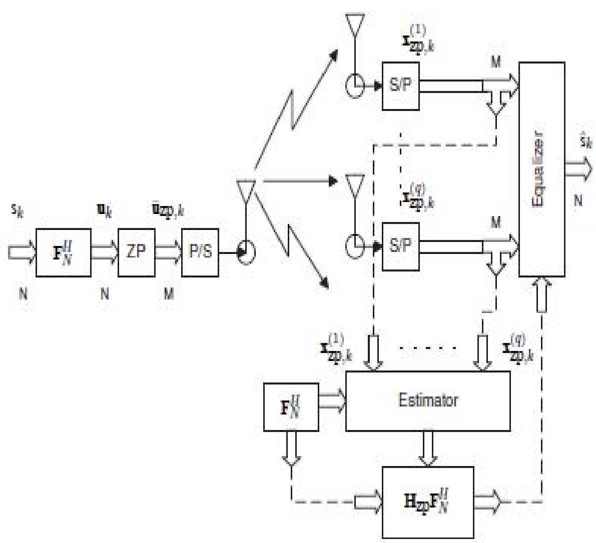

Fig. 4 - ZP-OFDM system: transmitter and receiver.

Figure 4 depicts the baseband discrete-time block equivalent model of a standard ZP-OFDM system. The only difference between ZP-OFDM and CP-OFDM is that the CP is replaced by L trailing zeros that are padded at each precoded block $\mathbf{u}_k$ to yield $\mathbf{u}_{zp,k} = \mathbf{T}_{zp}\mathbf{u}_k = [u_k(0), u_k(1), \ldots, u_k(N-1), 0, \ldots, 0]^T$ where $\mathbf{T}_{zp} = [\mathbf{I}^T_N, \mathbf{0}^T_{L\times N}]T$. We can write the received block symbol $\mathbf{x}_{zp,k}$ as

$$\begin{aligned}\mathbf{x}_{zp,k} &= \mathbf{H}_0\bar{\mathbf{u}}_{zp,k} + \overbrace{\mathbf{H}_1\bar{\mathbf{u}}_{zp,k-1}}^{\text{ISI}} + \mathbf{v}_{zp,k}\\ &= \mathbf{H}_0\mathbf{T}_{zp}\mathbf{F}_N^H\mathbf{s}_k + \underbrace{\mathbf{H}_1\mathbf{T}_{zp}\mathbf{F}_N^H\mathbf{s}_{k-1}}_{\text{ISI}} + \mathbf{v}_{zp,k}\end{aligned}$$

Where,

$$\mathbf{H}_0 = \begin{bmatrix} \mathbf{h}(0) & \mathbf{0} & \ldots & \ldots & \mathbf{0} \\ \vdots & \ddots & \ddots & \ddots & \vdots \\ \mathbf{h}(L) & \ldots & \mathbf{h}(0) & \ldots & \mathbf{0} \\ \mathbf{0} & \mathbf{h}(L) & \ldots & \mathbf{h}(0) & \vdots \\ \vdots & \ddots & \ddots & \ddots & \vdots \\ \mathbf{0} & \ldots & \mathbf{h}(L) & \ldots & \mathbf{h}(0) \end{bmatrix}$$

$$\mathbf{H}_1 = \begin{bmatrix} \mathbf{0} & \ldots & \mathbf{h}(L) & \ldots & \mathbf{h}(1) \\ \vdots & \ddots & \ddots & \ddots & \vdots \\ \vdots & \ddots & \ddots & \ddots & \mathbf{h}(L) \\ \mathbf{0} & \ldots & \ldots & \ldots & \mathbf{0} \\ \vdots & \ddots & \ldots & \ddots & \vdots \\ \mathbf{0} & \ldots & \ldots & \ldots & \mathbf{0} \end{bmatrix}.$$

$$\begin{aligned}\mathbf{x}_{zp,k} &= \left[\mathbf{x}^T_{zp,k}(0), \mathbf{x}^T_{zp,k}(1), \ldots, \mathbf{x}^T_{zp,k}(M-1)\right]^T\\ \mathbf{v}_{zp,k} &= \left[\mathbf{v}^T_{zp,k}(0), \mathbf{v}^T_{zp,k}(1), \ldots, \mathbf{v}^T_{zp,k}(M-1)\right]^T(\end{aligned}$$

and the key advantage of ZP-OFDM lies in the all-zero L ×N matrix **0** which eliminates the ISI, since $\mathbf{H}_1\mathbf{T}_{zp}\mathbf{F}^H_N = \mathbf{0}$.Forming the qM ×N matrix $\mathbf{H}_{zp}$ from the first N columns of matrix $\mathbf{H}_0$, can be expressed as

$$\mathbf{x}_{zp,k} = \mathbf{H}_{zp}\mathbf{F}_N^H\mathbf{s}_k + \mathbf{v}_{zp,k}$$

Where,

$$\begin{aligned}\mathbf{x}_{zp,k} &= \left[\mathbf{x}^T_{zp,k}(0), \mathbf{x}^T_{zp,k}(1), \ldots, \mathbf{x}^T_{zp,k}(M-1)\right]^T\\ \mathbf{v}_{zp,k} &= \left[\mathbf{v}^T_{zp,k}(0), \mathbf{v}^T_{zp,k}(1), \ldots, \mathbf{v}^T_{zp,k}(M-1)\right]^T(\end{aligned}$$

and **H**zp is, a block-Toeplitz matrix, defined as

$$\mathbf{H}_{zp} = \begin{bmatrix} \mathbf{h}(0) & \mathbf{0} & \ldots & \mathbf{0} \\ \vdots & \ddots & \ddots & \vdots \\ \mathbf{h}(L) & \ddots & \ddots & \vdots \\ \mathbf{0} & \ddots & \ddots & \mathbf{h}(0) \\ \vdots & \ddots & \ddots & \vdots \\ \mathbf{0} & \ldots & \mathbf{0} & \mathbf{h}(L) \end{bmatrix}.$$

Corresponding to the first N columns of **H**zp, the **H**0 submatrix is block-Toeplitz and is always guaranteed to be invertible, which assures symbol recovery (perfect detectability in the absence of noise) regardless of the channel zero locations.

3 SS-BASED METHOD

The desire for a more efficient algorithm led to the development of subspace(SS) methods for the blind estimation of multi-channel FIR filters [12]. The basic idea behind these methods consists of estimating the unknown parameters by exploiting the orthogonality of subspaces of certain matrices obtained by arranging in a prescribed order the second order statistics of the observation. This scheme shares many similarities with well-known techniques for direction-of-arrival (DOA) estimation in a narrow-band array processing context. The existence of such SS-based methods for blind estimation was brought to light by Gurelli and Nikias [9] and Moulines et al [12] (see also Hua [11] and Abed-Meraim [1], [5],[6], [10]).

Blind channel estimation is particularly important for OFDM applications where severe ISI can arise from the time-varying multipath fading that commonly exists in a mobile communication environment. The varying channel characteristics must be identified and equalised in real time to maintain the correct flow of information. The use of SS-based methods to accomplish blind SIMO channel estimation for OFDM has been proposed for frequency-flat fading channels in [7], [8]. The extension of it to the general MIMO case has been successfully introduced by Zeng et al [15]. More recently, some SS-based methods have been proposed for single-user OFDM systems [13], [14]. The method in [14] can be applied to OFDM systems without CP and, therefore, leads to higher data-rate.

4 MNS-BASED METHOD

Unfortunately, a widely acknowledged problem with the aforementioned techniques is its extensive computational complexity due to the EVD of a 'large' dimensional matrix and rather slow convergence with respect to the number of block symbols. In fast changing environments, such as in cellular communications, their applications may be limited. This problem is alleviated by the MNS decomposition approach proposed by Abed-Meraim et al [4]. Based on this contribution, it is easy to show that, only q − 1 properly chosen noise eigenvectors are just as efficient as using the

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whole noise subspace range(**G**cp) (or range(**G**zp)) to yield a consistent estimate of **H**cp for the CPOFDM system (or **H**zp for ZP-OFDM system). Furthermore, each of the q−1 noise eigenvectors can be found by using EVD of a 'small' dimensional covariance matrix corresponding to the (distinct) pairs of channel outputs given by a *properly connected sequence* (PCS) defined as follows :

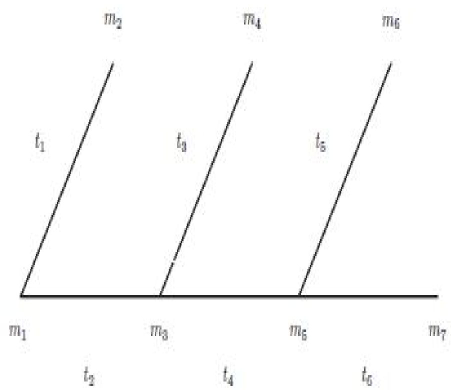

Fig. 5 - Tree that connects q = 7 channel outputs as its notes.

Definition 1: Denote the q system outputs by a set of members m1, . . . , mq. A combination of two (q ≥ 2) members (ti = mi1,mi2) is called a pair. A sequence of q−1 pairs is said to be properly connected if each pair in the sequence consists of one member shared by its preceding pairs and another member not shared by its preceding pairs.

Example 1: Consider a system with one input and seven outputs. The following sequence of pairs has minimum redundancy (six pairs) and spans all system outputs m1, . . . , m7.

t1 = (m1,m2), t2 = (m1,m3), t3 = (m3,m4)
t4 = (m3,m5), t5 = (m5,m6), t6 = (m5,m7)

Figure 5 demonstrates an example of PCS with q = 7. In the Tree pattern, the notes m2, m4, m6, and m7 are ending nodes while the nodes m1, m3, and m5 are branching nodes.

Remarks:

• MNS-based method can be applied to applications relating to source localisation and array calibration [4].

• In practice a PCS is easy to construct, however, it is not a necessary condition to give the MNS.

• A PCS exploits the diversity of the system outputs with minimum redundancy. This follows, since a sequence has less than q − 1 pairs or a pair in the sequence has less than two members, then the sequence does not give the required number of independent noise vectors.

• A set of q − 1 pairs span all the system outputs are not necessarily sufficient to give the required MNS.

5 SIMULATION RESULT

l	0	1	2	3	4
$h^{(1)}(l)$	-0.049+0.359i	0.482-0.569i	-0.556+0.587i	1.0000	-0.171+0.061i
$h^{(2)}(l)$	0.443-0.0384i	1.0000	0.921-0.194i	0.189-0.208i	-0.087-0.054i
$h^{(3)}(l)$	-0.211-0.322i	-0.199+0.918i	1.0000	-0.284-0.524i	0.136-0.190i
$h^{(4)}(l)$	0.417+0.03i	1.0000	0.873 + 0.145i	0.285+ 0.309i	-0.049+0.161i

Table 1: channel set :Impulse response

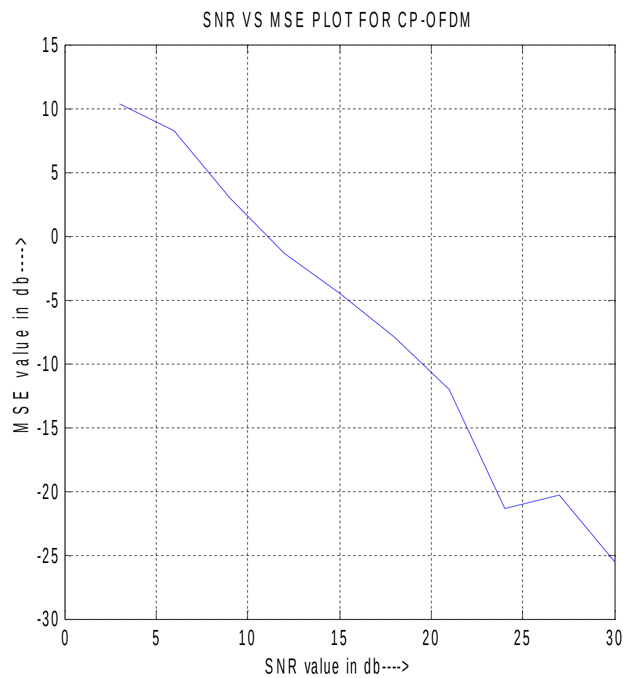

Fig .6 - Performance analysis of CPOFDM systems using SS-based method: SNR Vs MSE

Simulation Example 1: We first investigated the influence of weighting and performance of CP-OFDM receivers through the implementation of the SS-based method in terms of their estimation capabilities. We fixed the number of OFDM symbols to K = 1000, and varied the SNR from 5 to 30 dB.L=4; M=4; N=5; d=M+N; L: antenna . M: channel length. N: smoothing. d: equalization delay

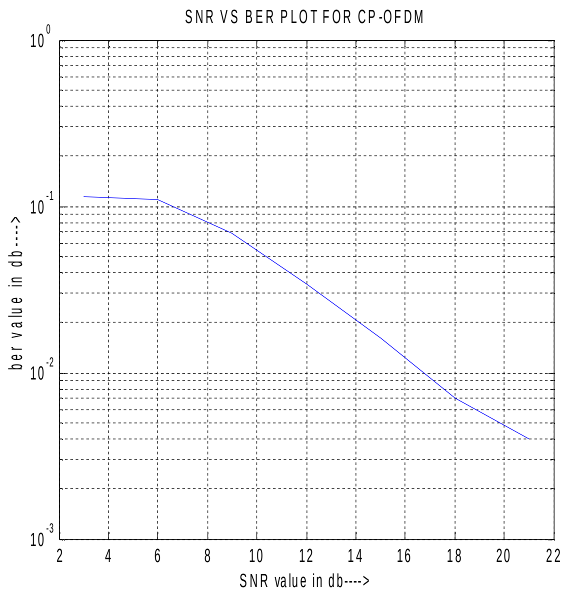

Fig .7 - Performance analysis of CPOFDM systems using SS-based method: SNR Vs BER

Simulation Example 2: The overall BER performance of the proposed SS-based method for the CP-OFDM systems Corresponding to SNR range of 2-20 dB. In order to check the equaliser gain.

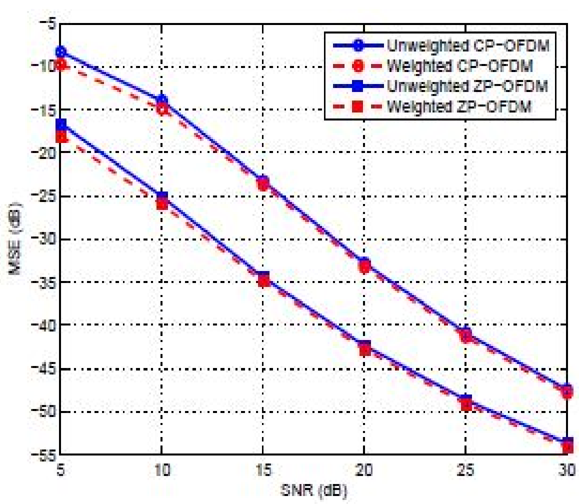

Fig . 8 - Performance comparison of CP-OFDM and ZP-OFDM systems using the SS-based method: SNR vs MSE

Simulation Example 3: We first investigated the influence of weighting and compare the CP-OFDM and ZP-OFDM receivers through the implementation of the SS-based method in terms of their estimation capabilities. We fixed the number of OFDM symbols to K = 1000, and varied the SNR from 5 to 30 dB. We simulated the output of a SIMO with q = 2 FIR channels of maximum order L = 4. The generated symbols are transmitted through 20 sub-carriers and so the size of the FFT / IFFT was N = 20.

6 PROPERTIES OF THE PROPOSED TECHNIQUES

In all the aforementioned SOS-based methods for OFDM based SIMO systems, the focus has been on channel identification and equalisation. In this section we give some comments on the above proposed techniques. For uniformity, we subsequently drop the subscripts m/n and express the covariance matrices as **R**(i) (corresponding to SS-based method) and **R**(i) (corresponding to MNS / SMNS-based method). Moreover, we drop the subscripts cp/zp and express the block-Circulant matrix **H**cp,(i) and block-Toeplitz matrix **H**zp,(i) as **H**(i) . For the sake of simplicity, we consider

ϱ =2N CP-OFDM
=2M ZP-OFDM

ν =qN CP-OFDM
=qM ZP-OFDM.

• The proposed channel estimators can be made to exploit the signal subspace regardless of the noise subspace and therefore the minimisation problem can be recast as a maximisation problem [5], [1], [12]. The solution of maximisation problem is considered more favorable to the minimisation problem, as there are fundamental limitations on the relative accuracy with which the smallest eigenvalues of the matrix can be computed, and they are more difficult to compute than the large ones. However, it is shown in [12] that the noise SS-based method exhibits better performance than the signal SS-based method.

• The main advantage of the MNS-based methods is that the large matrix EVD is avoided and the noise vectors are computed in parallel as the least eigenvectors of a smaller size covariance matrices **R**(i), i = 1, . . . , q − 1, which requires only O(ϱ^2) flops (in contrast with the O(ν^3) flops required for the computation of **R**). Comparatively, the SMNS- and MNS-based methods have almost the same order of computational cost for SIMO systems .

• The proposed ZP-OFDM estimator requires the EVD of a data correlation matrix of size m × m to extract the orthogonal subspace. In contrast, the proposed CP-OFDM estimator requires the EVD of data correlation matrix of size n×n. Since m > n, the proposed ZP-OFDM estimator is computationally more complex than the CPOFDM estimator.

• The CP-OFDM estimator is sensitive to channel zeros that are closed to the sub-carriers, whereas, the ZPOFDM estimator guarantees symbol recovery and offers a superior BER performance.

• The proposed CP-OFDM based SIMO system rely on the usual insertion of CP as in standard OFDM systems. Therefore, it does not require transmitter modification and is applicable to all standardised OFDM systems. In contrast, the ZP-OFDM based SIMO system presented requires transmitter modification to introduce ZP redundancy by a filter-bank precoder. Note that ZP is used in the DVB standard in the form of guard bits.
• By using the optimal weighting matrix **W**(i), the identification procedure becomes quite insensitive to the ill conditioning problem. In fact, if a pair of channels have close common zeros, the corresponding block-channel Matrix **H**(i) becomes nearly singular and consequently **W**(i) becomes large. Therefore, the inverse of the weighting matrix,**W**(i)−1, will qualitatively provide 'more weighting' (i.e., larger weighting coefficients) to the noise eigen vectors associated with a well conditioned **H**(i) than to those corresponding to ill conditioned block channel matrix. However, weighting MNS (WMNS)-based method often incurs high complexity and involves large decoding delay, and does not trade well for the accuracy improvement [5].

7 CONCLUSION

This paper presents original reformulation of the SS-based estimation procedure for the blind identification of OFDM based SIMO FIR channels. It fully exploits the relations between the noise subspace of a certain covariance matrix formed from the observed signals. This reformulation provides some additional insights into the existing subspace algorithms. More importantly, it allows one to analyse the second order statistics of the output signals for the case of CP-OFDM and ZPOFDM receivers. This technique, although reliable and robust in some scenarios, require a computationally expensive and non-parallelisable EVD to extract the noise subspace. In fast changing environments, such as in cellular communications, their application may be limited and impossible (too costly) to implement. These problems are alleviated by a MNS based method which exploits a minimum number of noise eigenvectors for multi-channel identification. This technique of MNS, and especially the concept of PCS, turns out to be a powerful tool that can be applied to other OFDM-based SIMO systems. The proposed MN-based method is much more computationally efficient than the standard SS-based method at the price of a slight loss of estimation accuracy. However, better estimates of FIR channels can be obtained by a symmetric version of MNS with the same order of computational cost. Simulations have shown that the proposed method are effective and robust.

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