Channel Estimation for MIMO-OFDM Systems Based on Data Nulling Superimposed Pilots

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***Abstract*—This paper proposes a new channel estimation algorithm based on data nulling superimposed pilots for the spatial multiplexing multiple-input-multiple-output (MIMO) orthogonal frequency division multiplexing (OFDM) systems. In the proposed method each OFDM data symbol of each transmit antenna is spread over all subcarriers by using a spreading matrix then nulls are introduced at certain subcarriers to cancel the mutual interference between data symbols and superimposed pilots. At receiver accurate channel estimation can be easily acquired based on the superimposed pilots. Then the superimposed pilots are removed from the received signal and simple iterative data detection scheme is used to compensate the distortion which occurred in the data symbols. The simulation results of the proposed algorithm show improvement in the estimation accuracy, bit error rate (BER) and computational complexity compared to that of the conventional superimposed pilot technique. The simulation results also show that the performance of the proposed technique approaches that of the frequency division multiplexed pilots technique while having higher data rate and some excess in the receiver complexity.**

I. INTRODUCTION

Orthogonal frequency division multiplexing (OFDM) offers high robustness to frequency selective-fading channels, high data rates, simple channel estimation and equalization methods [1]. Spatial multiplexing of multiple-input-multiple-output (MIMO) system offers high spectral efficiency and can be used to support high capacity demands [2-3]. These introduce spatial multiplexing MIMO-OFDM as an attractive scheme in nowadays standards [4-5]. However, the good performance of MIMO systems is conditioned on an accurate channel estimation at the receiver, so channel estimation is considered as a bottleneck for good performance MIMO systems.

In MIMO-OFDM systems, channel estimation is commonly performed based on pilot-assisted techniques [6] where frequency division multiplexed (FDM) pilots are multiplexed with data symbols. This technique can easily acquire accurate channel estimates, but the inserted FDM pilots decrease the data rate and the spectral efficiency due to the large number of FDM pilots required [7], especially in fast varying channels in which the number of pilot symbols is usually increased to track the channel variations accurately. Another channel estimation approach is the blind techniques where no pilots are transmitted, and the estimation process is based on higher order statistics of the received signals [8]. The main drawback of this method is that the data sequence should be long enough in order to get an accurate channel estimation [7] which, may be impractical especially, in fast varying channels. On the other hand, semi-blind channel estimation methods [8] depend on both the transmission of pilot symbols and the statistical properties of the received signal which results in a smaller number of pilot symbols and hence increases the spectral efficiency. Another attractive technique which is introduced in single-input-single-output (SISO) systems is the channel estimation based on superimposed pilots, in which superimposed pilots sequence known to the receiver is algebraically added to the data symbols and then used for channel estimation [9].This technique increases the spectral efficiency, but yet it has a limited performance due to the mutual interference between data symbols and the superimposed pilots. Therefore, iterative methods are introduced [10] to mitigate the mutual interference between data symbols and the superimposed pilots. Hence, offer better performance at the expense of increased receiver complexity. So, channel estimation based on superimposed pilots is considered as an attractive method in MIMO systems as it will provide high data rates and high spectral efficiency [11-12].

A new channel estimation method based on superimposed pilots is introduced in [13] for the SISO-OFDM system named data nulling superimposed pilots (DNSP) where data symbols are pre-coded among all subcarriers, and then a nulling matrix is used to introduce nulls at certain subcarriers that correspond to the superimposed pilots' active subcarriers. This method showed a promising improvement in terms of channel estimation accuracy and BER.

This paper proposes a channel estimation algorithm based on DNSP for the spatial multiplexing MIMO-OFDM system. The main objective of the proposed algorithm is to introduce an accurate channel estimation technique that increases the spectral efficiency compared to channel estimation techniques based on FDM pilots. In the proposed method each OFDM data symbol of each transmit antenna is pre-coded by using a spreading matrix that spreads data symbols on all subcarriers then nulls are introduced using a nulling matrix at certain subcarriers positions. Then orthogonal superimposed pilots are added in the frequency domain for each transmit antenna. This scheme eliminates the mutual interference between the data symbol and the superimposed pilots consequently it offers accurate channel estimation. After the superimposed pilots are removed from the frequency domain received signal and equalization process is performed, an iterative detection method is used to restore the OFDM data symbols.

The rest of the paper is organized as follows: the system model of the proposed algorithm is described in Section II, and

Section III presents the proposed MIMO-OFDM channel estimation technique. In Section IV, the MIMO-OFDM data detection process is discussed. In Section V, the complexity analysis is discussed and, the simulation results are presented in Section VI. Finally, some concluding remarks are discussed in Section VII.

In this paper, bold letters are used for matrices and vectors, the superscripts $(.)^*$, $(.)^{\mathrm{H}}$ and $(.)^{-1}$ are used to denote complex conjugate, matrix Hermitian transpose and matrix inversion respectively, and $\mathrm{E}\{.\}, \mathrm{Tr}\{.\}$ stand respectively for the expected value and the matrix trace operator. And $\mathrm{diag}[a_1, a_2, \ldots, a_{K-1}]$ is the $K \times K$ diagonal matrix whose diagonal entry is a_n. The discrete Fourier transform (DFT) of a $K \times 1$ vector $\mathbf{x}$ is denoted by $\mathbf{F}_K\mathbf{x}$, where $\mathbf{F}_K$ matrix of size $K \times K$ has (m, n) entry $1/\sqrt{K}e^{-2\pi\sqrt{-1}mn/K}, m, n = 0, .., K-1$, and $\mathbf{F}_{T,L}$ denotes the leading $T \times L$ submatrix of $\mathbf{F}_T$. Finally $\mathbf{I}_K$ denotes a $K \times K$ identity matrix.

II. SYSTEM MODEL

We consider a MIMO-OFDM system with K subcarriers and N_T transmit, N_R receive antennas where ($N_R \geq N_T$).The initial superimposed pilots of length K for the first transmit antenna denoted by $\mathbf{p}^1$ results from the K-point DFT of a periodic time-domain vector denoted by $\mathbf{g} = [\mathrm{g}_0, \mathrm{g}_1, \ldots, \mathrm{g}_{K-1}]^T$of period T where g_n defined as follows [10]:

$$\mathrm{g}_n = \sigma_g e^{j\frac{\pi}{T}\left(n_{[T]}\,(n_{[T]}+b)\right)} \qquad n = 0, 1 \ldots . K-1, \quad (1)$$

where $b = 1$ if T is odd, $b = 2$ if T is even, σ_g^2 denotes the power of the superimposed pilot and $n_{[T]}$ is the residue of n modulo T. Since $\mathbf{g}$ is periodic with period T then after performing N-point DFT its energy will be concentrated at T active subcarriers with spacing equals $C = K/T$ as follows:

$$\mathbf{p}^1 = \mathbf{F}_K\mathbf{g} = [\mathrm{p}_0^1, \mathrm{p}_1^1, \ldots, \mathrm{p}_{K-1}^1]^T$$

$$\text{where , } |\mathrm{p}_\mathrm{n}^1|^2 = \begin{cases} C{\sigma_g}^2 & n = tC, t = 0,1 \ldots T-1 \\ 0 & elsewhere \end{cases}. \quad (2)$$

The superimposed pilots for the i-th transmit antenna can be simply acquired by circularly shifting the initial superimposed pilots ($\mathbf{p}^1$) for $(i-1)f$ subcarriers where $f = C/N_T$ and f is assumed to be an integer number as follows:

$$\mathrm{p}_n^i = \mathrm{p}_{(n+f(i-1))[K]}^1 \qquad 1 \leq i \leq N_T. \quad (3)$$

The orthogonality among superimposed pilots of different transmit antennas is preserved owing to the circular shift of the initial superimposed pilots.

The proposed transmitter is shown in Fig.1(a). The OFDM data symbols vector of each transmit antenna of length K denoted by $\mathbf{s}^i = [\mathrm{s}_0^i, \mathrm{s}_1^i, \ldots, \mathrm{s}_{K-1}^i]^T$ is initially spread among all subcarriers by using a $K \times K$ unitary matrix denoted by $\mathbf{W}$ as follows:

$$\mathbf{q}^i = \mathbf{W}\mathbf{s}^i \qquad 1 \leq i \leq N_T . \quad (4)$$

To cancel the mutual interference between superimposed pilots and data symbols a distortion is introduced by using a nulling matrix such that nulls are introduced at TN_T subcarrier positions which are the positions of the superimposed pilots. The nulling matrix used denoted by ($\mathbf{I}_K - \mathbf{J}$), where $\mathbf{J}$ is $K \times K$ diagonal matrix with diagonal entries defined as follows:

$$[\mathbf{J}]_{ee} = \begin{cases} 1 & e = af \\ 0 & elsewhere \end{cases} \quad where\ a = 0,1,2, \ldots TN_T - 1. \quad (5)$$

The output vector from the nulling block can be expressed as follows:

$$\mathbf{x}^i = (\mathbf{I}_K - \mathbf{J})\,\mathbf{q}^i \qquad 1 \leq i \leq N_T . \quad (6)$$

Then the superimposed pilots are added to the spread and distorted data symbol vector ($\mathbf{x}^i$) to result in the frequency domain transmitted vector defined as follows:

$$\bar{\mathbf{z}}^i = (\mathbf{I}_K - \mathbf{J})\,\mathbf{W}\mathbf{s}^i + \mathbf{p}^i \qquad 1 \leq i \leq N_T . \quad (7)$$

Then $\mathbf{z}^i$ is passed to a K-point IDFT resulting in the time domain vector as follows:

$$\mathbf{z}^i = \mathbf{F}_K^{\mathrm{H}}\,\bar{\mathbf{z}}^i = [\mathrm{z}_0^i, \mathrm{z}_1^i, \ldots, \mathrm{z}_{K-1}^i]^T \qquad 1 \leq i \leq N_T . \quad (8)$$

After that a cyclic prefix (CP) of length ($\bar{L} \geq L$) is appended to the time domain vector of each transmit antenna where L is the length of the multipath channel which is modeled as time-variant finite impulse response (FIR) filter with order L. The channel is assumed to be quasi-static which means that the channel

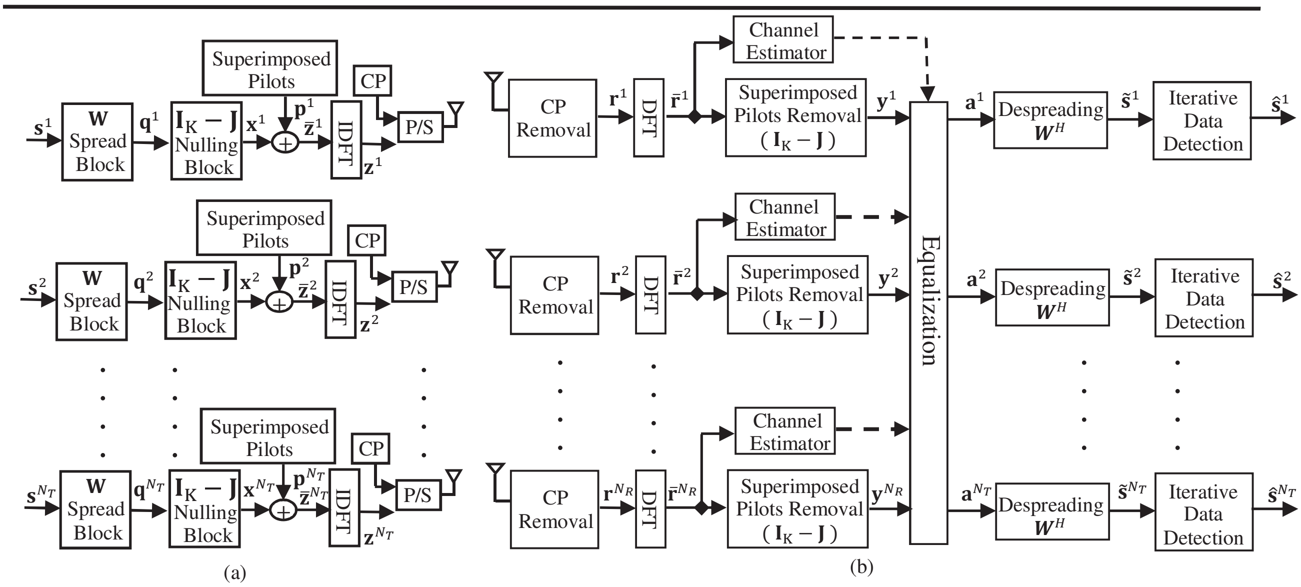

Fig. 1. a- Transmitter block diagram b- Receiver block diagram

coefficients remain constant within one OFDM block but can vary from one OFDM block to another.

The Channel Impulse Response (CIR) vector between the m-th receive antenna and the i-th transmit antenna is denoted by $\mathbf{h}^{mi} = [\mathrm{h}_0^{mi}, \mathrm{h}_1^{mi}, \dots \dots \mathrm{h}_{L-1}^{mi}]^{\mathrm{T}}$. The channel taps are assumed to be statistically independent, and $\sigma_l^2 = \mathrm{E}\left[\,|\mathrm{h}_l^{mi}|^2\right]$ where $m = 1,2, \dots N_R$, $i = 1,2 \dots$, N_T is the average power of the l-th channel tap. The power of all channel taps is normalized such that $\sum_{l=0}^{L-1} \sigma_l^2 = 1$.

The received time domain signal of the m-th receive antenna after removing the CP is defined as follows:

$$\mathrm{r}_n^m = \sum_{i=1}^{N_T} \sum_{l=0}^{L-1} \mathrm{h}_l^{mi}\, \mathrm{z}_{(n-l)_{[K]}}^i + \mathrm{v}_n^m \qquad 1 \le m \le N_R\,, \tag{9}$$

where v_n^m is the additive white gaussian noise (AWGN) with a variance of σ_{v}^2.

The proposed receiver block diagram is shown in Fig.1(b) where the received signal vector $\mathbf{r}^m=[\mathrm{r}_0^m, \mathrm{r}_1^m, \dots, \mathrm{r}_{K-1}^m]^T$is passed to the K–point DFT to generate the frequency domain received vector as follows:

$$\bar{\mathbf{r}}^m = \mathbf{F}_K\, \mathbf{r}^m = \sum_{i=1}^{N_T} \mathbf{H}^{mi}\, \bar{\mathbf{z}}^i \; + \bar{\mathbf{v}}^m, \tag{10}$$

where $\bar{\mathbf{r}}^m = [\bar{\mathrm{r}}_0^m, \bar{\mathrm{r}}_1^m, \dots, \bar{\mathrm{r}}_{K-1}^m]^T$denotes the received frequency domain vector at the m-th antenna, $\bar{\mathbf{v}}^{\,m}$denotes the K–point DFT of the AWGN vector $\mathbf{v}^m$=$[\mathrm{v}_0^m, \mathrm{v}_1^m, \dots, \mathrm{v}_{K-1}^m]^T$of covariance matrix $\sigma_{\mathrm{v}}^2 \mathbf{I}_K$ and $\mathbf{H}^{mi}$ is the Channel Frequency Response (CFR) matrix of size $K \times K$ between the i-th transmit and m-th receive antenna. Since the multipath channel is assumed to be quasi-static channel then $\mathbf{H}^{mi}$ is a diagonal matrix defined as follows:

$$\mathbf{H}^{mi} = \mathrm{diag}\;[\mathbf{F}_K \mathbf{h}^{mi}] = \mathrm{diag}\;[\mathrm{H}_0^{mi}, \mathrm{H}_1^{mi}, \dots, \mathrm{H}_{K-1}^{mi}]$$
$$\text{where} \quad 1 \le i \le N_T \quad , \quad 1 \le m \le N_R \tag{11}$$

III. MIMO –OFDM Channel Estimation

Since superimposed pilots' active subcarriers of each transmit antenna is free of data symbols interference in the frequency domain, therefore multipath channel can be easily estimated if $T \ge L$ .The CFR estimation at superimposed pilots' active subcarriers can be done easily using the least square (LS) estimation as follows:

$$\widehat{\mathrm{H}}_{b_t}^{mi} = \frac{\bar{\mathrm{r}}_{b_t}^m}{\mathrm{p}_{b_t}^i} \tag{12}$$
$$where \quad b_t = tC + (i-1)f \quad \& \; t = 0,1,..T-1.$$

Now CFR matrix $\widehat{\mathbf{H}}^{mi}$ can be computed using three steps. The first step is implementing a T-point IDFT on the estimated CFR vector denoted by $\mathbf{u}^{mi} = [\widehat{\mathrm{H}}_{b_0}^{mi}, \widehat{\mathrm{H}}_{b_1}^{mi}, \dots, \widehat{\mathrm{H}}_{b_{T-1}}^{mi}]^T$ as follows:

$$\tilde{\mathbf{h}}^{mi} = \frac{1}{\sqrt{T}} \mathbf{F}_{T,L}^{\mathrm{H}} \mathbf{u}^{mi} = [\tilde{\mathrm{h}}_0^{mi}, \tilde{\mathrm{h}}_1^{mi}, \dots \tilde{\mathrm{h}}_{L-1}^{mi}]^T, \tag{13}$$

where $\tilde{\mathbf{h}}^{mi}$ is the estimated CIR multiplied by a certain phase shift which results from using circular shift superimposed pilots in the frequency domain. Consequently, the second step is to remove this phase shift which depends on the shift of the superimposed pilots of each transmit antenna with respect to the initial superimposed pilots. So, this phase shift can be computed and removed easily to result the estimated CIR as follows:

$$\hat{\mathrm{h}}_l^{mi} = \tilde{\mathrm{h}}_l^{mi} e^{\left(\frac{2\pi\sqrt{-1}(i-1)fl}{K}\right)}$$
$$\text{where } 1 \le i \le N_T \qquad , \quad l = 0,1 \dots L-1. \tag{14}$$

Finally, the third step to compute the CFR matrix is to perform the K-point DFT on the estimated CIR vector denoted by $\hat{\mathbf{h}}^{mi} = [\hat{\mathrm{h}}_0^{mi}, \hat{\mathrm{h}}_1^{mi}, \dots \dots \hat{\mathrm{h}}_{L-1}^{mi}]^T$as follows:

$$\widehat{\mathbf{H}}^{mi} = \mathrm{diag}\;[\mathbf{F}_{\mathrm{K}}\, \hat{\mathbf{h}}^{mi}] = \mathrm{diag}\big[\widehat{\mathrm{H}}_0^{mi}, \widehat{\mathrm{H}}_1^{mi}, \dots \dots \widehat{\mathrm{H}}_{K-1}^{mi}\big]. \tag{15}$$

The MSE of the estimated CIR ($\hat{\mathbf{h}}^{mi}$) is as follows:

$$\mathrm{MSE}\big(\,\hat{\mathbf{h}}^{mi}\big) = \varepsilon_{\,\hat{\mathbf{h}}^{mi}} = \frac{1}{\mathrm{L}} \mathrm{Tr}\{\mathrm{E}[\;(\hat{\mathbf{h}}^{mi} - \mathbf{h}\,)\;(\,\hat{\mathbf{h}}^{mi} - \mathbf{h}\,)^{\mathrm{H}}]\}$$
$$\varepsilon_{\,\hat{\mathbf{h}}^{mi}} = \frac{\sigma_{\mathrm{v}}^2}{\mathrm{LT}} \mathrm{Tr}\{\mathbf{F}_{T,L}^{\mathrm{H}} \mathbf{G}^{-1} \mathbf{F}_{T,L}\}, \tag{16}$$

where $\mathbf{G} = diag(|\mathrm{p}_{b_0}^i|^2, |\mathrm{p}_{b_1}^i|^2, , , |\mathrm{p}_{b_{T-1}}^i|^2)$,which corresponds to the power of superimposed pilots' active subcarriers so the MSE is given by :

$$\varepsilon_{\,\hat{\mathbf{h}}^{mi}} = \frac{\sigma_{\mathrm{v}}^2}{K\sigma_g^2} \tag{17}$$

It is clear from (17) that the OFDM data symbol does not affect the performance of channel estimation which is expected as there is no mutual interference between data symbols and superimposed pilots which will lead to an accurate channel estimation results as it will be shown in the simulation results .

IV. MIMO-OFDM Data Detection

Firstly, the superimposed pilots are removed from the received frequency domain vector as follows:

$$\mathbf{y}^m = (\,\mathbf{I}_K - \mathbf{J}\,)\bar{\mathbf{r}}^m = \sum_{i=1}^{N_T} \mathbf{H}^{mi}\mathbf{x}^i \; + (\,\mathbf{I}_K - \mathbf{J}\,)\bar{\mathbf{v}}^m$$
$$= [\mathrm{y}_0^m, \mathrm{y}_1^m, \dots, \mathrm{y}_{K-1}^m]^T \tag{18}$$

where $(\,\mathbf{I}_K - \mathbf{J})^2 \; = (\,\mathbf{I}_K - \mathbf{J}\,)$ .The removal of superimposed pilots makes the noise vector colored by the nulling matrix and its covariance matrix is changed to be $\sigma_{\mathrm{v}}^2(\,\mathbf{I}_K - \mathbf{J}\,)$. The equalization process will be done individually on each subcarrier. Define the received vector of the N_R receive antennas at the k-th subcarrier by $\tilde{\mathbf{y}}_k = [\mathrm{y}_k^1, \mathrm{y}_k^2, \dots, \mathrm{y}_k^{N_R}]^T$. Define the MIMO channel of size $N_R \times N_T$ at the k-th subcarrier as follows:

$$\widetilde{\mathbf{H}}_k = \begin{bmatrix} \widehat{\mathrm{H}}_k^{11} & \widehat{\mathrm{H}}_k^{12} & \cdots & \widehat{\mathrm{H}}_k^{1N_T} \\ \widehat{\mathrm{H}}_k^{21} & \widehat{\mathrm{H}}_k^{22} & \cdots & \widehat{\mathrm{H}}_k^{2N_T} \\ \vdots & \vdots & \ddots & \vdots \\ \widehat{\mathrm{H}}_k^{N_R 1} & \widehat{\mathrm{H}}_k^{N_R 2} & \cdots & \widehat{\mathrm{H}}_k^{N_R N_T} \end{bmatrix}. \tag{19}$$

Then minimum MSE equalization of the received vector at the k-th subcarrier will be as follows:

$$\mathbf{b}_k = [\gamma(k)\mathbf{I}_{N_T} + \widetilde{\mathbf{H}}_k{}^H\widetilde{\mathbf{H}}_k]^{-1}\widetilde{\mathbf{H}}_k{}^H\tilde{\mathbf{y}}_k = [\mathrm{b}_k^1, \mathrm{b}_k^2, \dots, \mathrm{b}_k^{N_T}]^T$$

where $\gamma(k) = \begin{cases} 0 & k = af \\ \sigma_v^2 & else \end{cases}$ $\quad where\ a = 0,1,2, \dots TN_T - 1.$ (20)

The next step after the equalization process has been done on all subcarriers is the despreading. Define the equalized OFDM symbol of the i-th transmit antenna by:

$$\mathbf{a}^i = [\mathrm{b}_0^i, \mathrm{b}_1^i, \dots, \mathrm{b}_{K-1}^i]^T \approx (\mathbf{I}_K - \mathbf{J})\,\mathbf{W}\mathbf{s}^i + \mathbf{v}', \quad (21)$$

where $\mathbf{v}'$ is the noise vector of length K after performing the equalization process. Now, the despreading process takes place as follows:

$$\tilde{\mathbf{s}}^i = \mathbf{W}^{\mathrm{H}}\mathbf{a}^i \approx \mathbf{s}^i - \mathbf{W}^{\mathrm{H}}\mathbf{J}\mathbf{W}\mathbf{s}^i + \mathbf{W}^{\mathrm{H}}\mathbf{v}', \quad (22)$$

where $\mathbf{W}^{\mathrm{H}}\mathbf{W} = \mathbf{I}_K$. Symbol by symbol detection is initialized by treating $\mathbf{W}^{\mathrm{H}}\mathbf{J}\mathbf{W}\mathbf{s}^i$ as an extra additive noise and considering $\tilde{\mathbf{s}}^i$ as a soft decision of $\mathbf{s}^i$. So, the initial hard decision of $\mathbf{s}^i$ is given by:

$$(\hat{\mathbf{s}}^i)^{(0)} = \lfloor \tilde{\mathbf{s}}^i \rceil \quad , \quad 1 \le i \le N_T. \quad (23)$$

For the next iteration, the detected symbols from the previous hard decision process is used to compute $\mathbf{W}^{\mathrm{H}}\mathbf{J}\mathbf{W}\mathbf{s}^i$ and compensate for the data distortion. So, the detected symbol for the j-thiteration is given by:

$$(\hat{\mathbf{s}}^i)^{(j)} = \lfloor \tilde{\mathbf{s}}^i + \mathbf{W}^{\mathrm{H}}\mathbf{J}\mathbf{W}(\hat{\mathbf{s}}^i)^{(j-1)} \rceil \quad , 1 \le i \le N_T. \quad (24)$$

V. Complexity analysis

In this section, we analyze the computational complexity of the proposed DNSP scheme in terms of the required number of real multiplications and real additions for channel estimation and data detection processes per one OFDM symbol, then compare it with that of two other schemes: the first scheme is the conventional superimposed pilots scheme [11] which firstly acquire an initial inaccurate channel estimation based on the superimposed pilots then uses an iterative joint channel estimation and data detection algorithm to mitigates the mutual interference between superimposed pilots and OFDM data symbols hence improve the performance in terms of BER and MSE. The second scheme is the FDM scheme for the precoded OFDM [13] that spread OFDM data symols over all subcarriers then extra TN_T subcarriers is inserted as pilots for channel estimation at the cost of increasing bandwidth due to the FDM pilots overhead. In this analysis, one complex multiplication is counted as 4 real multiplications and 2 real additions. The CFR estimation at superimposed pilots' active subcarriers (12) needs 6 real multiplications and 2 real additions for single active subcarrier. Thus, for $N_T \times N_R$ MIMO system with superimposed pilots of T active subcarriers it needs $6TN_TN_R$ real multiplications and $2TN_TN_R$ real additions.The IDFT operation in (13) needs LTN_TN_R complex multiplications and $L(T-1)N_TN_R$ complex additions so it needs $4LTN_TN_R$ real multiplications and $2L(2T-1)N_TN_R$ real additions. The CIR computation in (14) needs $4LN_TN_R$ real multiplications and $2LN_TN_R$ real additions. So the overall number of operations for CIR estimation for $N_T \times N_R$ MIMO system based on the proposed DNSP scheme is $2N_T N_R (3T + 2L(T+1))$ real multiplications and $2T(2L+1)N_TN_R$ real additions.

TABLE I. Complexity analysis comparison

Scheme	**CIR Estimation process**
	Number of real multiplication operations
DNSP	$2N_T N_R (3T + 2L(T+1))$
FDM	$2N_T N_R (3T + 2L(T+1))$
Conv.SP at jth iteration	$2N_TN_R \begin{bmatrix} N_TT^2(j+1)(4K + N_TT + 3) \\ +2jK(T+1) + 2KT \end{bmatrix}$
	Number of real addition operations
DNSP	$2N_TN_RT(2L+1)$
FDM	$2N_TN_RT(2L+1)$
Conv.SP at jth iteration	$2N_TN_R \begin{bmatrix} N_TT^2(j+1)(N_TT\ (K+1) + 2K) \\ +2(j+1)TK + 3jK - T(j+1) \end{bmatrix}$
Scheme	**Data Detection process**
	Number of real multiplication operations
DNSP at jth iteration	$2KN_T[N_T(N_T + N_R + 5) + 2N_R + 2K(j+2)]$
FDM	$2KN_T[N_T(N_T + N_R + 5) + 2N_R + 4K]$
Conv.SP at jth iteration	$(j+1) \begin{bmatrix} 8KN_TN_R(N_T+1) \\ +2KN_T^2(N_T+3) \end{bmatrix}$
	Number of real addition operations
DNSP at jth iteration	$KN_T \begin{bmatrix} N_T(2N_T + 3N_R + 4) + 4N_R \\ +6K + 2j(2K-1) - 8 \end{bmatrix}$
FDM	$KN_T \begin{bmatrix} N_T(2N_T + 3N_R + 4) + 4N_R \\ +6K - 8 \end{bmatrix}$
Conv.SP at jth iteration	$(j+1) \begin{bmatrix} KN_TN_R(8N_T+6) \\ +2KN_T(N_T^2 - 1) \end{bmatrix}$

The complexity of the MMSE equalization of one OFDM symbol, as performed in (20), is based on an efficient method for matrix inversion introduced in [14] is $2KN_T^2[N_T + N_R + 5] + 4KN_T N_R$ real multiplications and $KN_T^2[2N_T + 3N_R + 4] + 4KN_T(N_R - 1)$ real additions. The despreading process for each transmit antenna, as performed in (22), needs $4K^2$ real multiplications and $2K^2 - 2K$ real additions so for the all transmit antennas it will need $4K^2N_T$ real multiplications and $[2K^2 - 2K]N_T$ real additions. Finally, the iterative data detection in (23) needs $4K^2N_T$ real multiplications and $[4K^2 - 2K]N_T$ real additions for each iteration. So the overall number of operations for jth iteration data detection process is real $2KN_T[N_T(N_T + N_R + 5) + 2N_R + 2K(j+2)]$ real multiplications and $KN_T[N_T(2N_T + 3N_R + 4) + 4N_R + 6K + 2j(2K-1) - 8]$ real additions. Comparison between DNSP scheme, FDM pilot scheme and conventional superimposed pilots scheme in terms of computational complexity is summarized in Table I. Also, a numerical illustration of the complexity of the three schemes under comparison will be calculated and discussed at the end of Section VI.

VI. Simulation results

In the simulation we consider an OFDM scheme of $K = 2048$ subcarriers with CP of length 64 samples and quadrature phase shift keying (QPSK) modulation, the channel is randomly generated and assumed to be uncorrelated Rayleigh fading channel with $L = 8$, their powers are given by the exponential delay profile $E[|\mathrm{h}_l|^2] = e^{-0.2l}$. The power of all transmit antennas are set to be equal and the total transmitted power is normalized such that $\sum_{t=1}^{N_T} \sigma_s^2 = 1$ where σ_s^2 is the power of each transmit antenna. So that the radiated power is independent on N_T [15]. Superimposed pilots of $T = 8$ and superimposed pilot to data power ratio denoted by $\alpha = \sigma_g^2/\sigma_s^2$ of 0.2 are used. We compare the MSE of the channel estimation and BER performance of DNSP MIMO scheme with two other schemes the first scheme is the conventional superimposed pilots scheme [11] which uses an iterative joint channel estimation and data detection algorithm with the same superimposed pilot to data power ratio of 0.2 and The second scheme is the FDM scheme for the precoded OFDM [13].

Fig. 2. Shows the MSE of $N_T = N_R = 4$ MIMO system for the DNSP scheme, conventional superimposed pilot scheme at iteration=0, 1 and 2 and FDM scheme. It is clear that the DNSP MSE is much better than the conventional superimposed pilots scheme even after 2 iterations and the MSE of DNSP is the same as that of FDM scheme although FDM scheme requires extra TN_T dedicated subcarriers for channel estimation process.

Fig. 3. Shows the BER of $N_T = N_R = 4$ MIMO system for the DNSP scheme at iteration=0, 1 and 2, the conventional superimposed pilot scheme at iteration=0, 1 and 2 and FDM scheme. It is found that the BER of DNSP at all iterations are much lower than that of the conventional superimposed pilots and approaches that of the FDM scheme. Also, it is clear that most of the gain in data symbols detection in DNSP scheme is obtained in the first iteration.

Fig. 4. Shows the BER of $N_T = N_R = 8$ MIMO system for the three schemes under comparison. Although the performance of the DNSP scheme is still much better than conventional superimposed pilots and close to that of the FDM but by comparing Fig.3. and Fig4 we can obtain an important observation that the gap in BER performance between DNSP and FDM of the $N_T = N_R = 8$ MIMO system increased compared to that of the $N_T = N_R = 4$ MIMO system. This increased gap is attributed to the fact that when the number of transmit antennas increases the distortion introduced to OFDM data symbol will increase so the performance of iterative data detection will deteriorate.

The computational complexity in terms of number of real multiplications and additions operations of the three schemes mentioned before were calculated according to the simulation parameters for $N_T = N_R = 4$ MIMO system at the first iteration and illustrated in Table II. It is found that the overall computational complexity of the DNSP scheme is higher than that of FDM scheme and lower than that of the conventional superimposed pilots because in conventional superimposed pilots channel estimation process and data detection process are iteratively done but in DNSP only the data detection process is iteratively done.

TABLE II. COMPLEXITY ANALYSIS COMPARISON FOR 4×4 MIMO SYSTEM AT THE FIRST ITERATION

Scheme	CIR Estimation process	Data Detection process	All processing
Number of real multiplications operations			
DNSP	5.4×10^3	2×10^8	2×10^8
FDM	5.4×10^3	1.35×10^8	1.35×10^8
Conv.SP	1.4×10^8	3.55×10^6	1.4×10^8
Number of real additions operations			
DNSP	4.4×10^3	1.7×10^8	1.7×10^8
FDM	4.4×10^3	10^8	10^8
Conv.SP	1.15×10^9	3×10^6	1.15×10^9

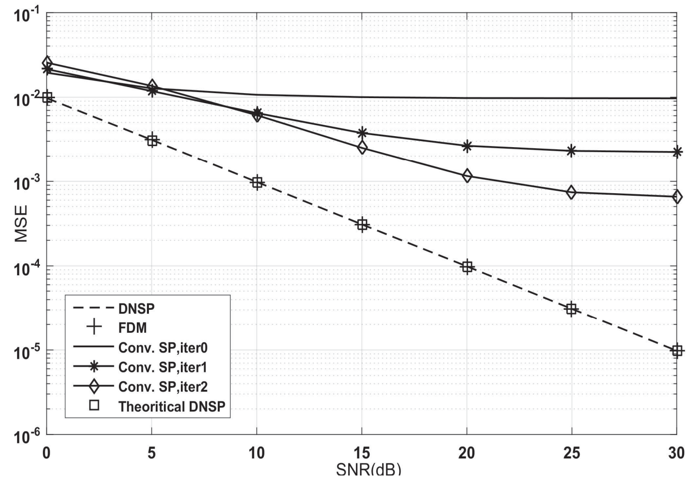

Fig. 2. MSE of the DNSP scheme, Conventional superimposed pilot scheme and FDM scheme for $N_T = N_R = 4$ MIMO system

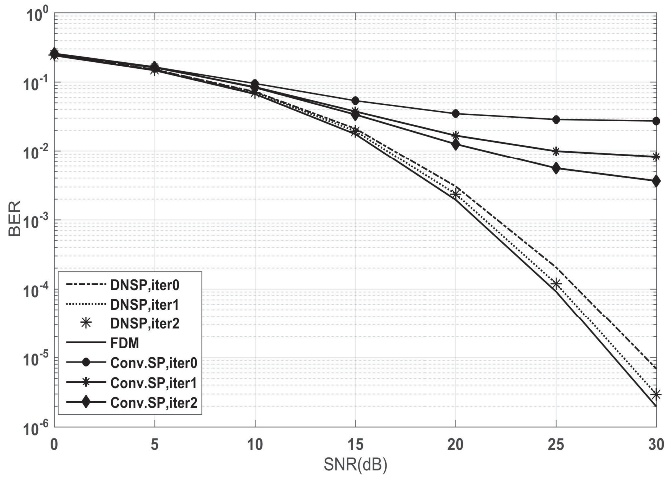

Fig. 3. BER of the DNSP scheme, Conventional superimposed pilot scheme and FDM scheme for $N_T = N_R = 4$ MIMO system

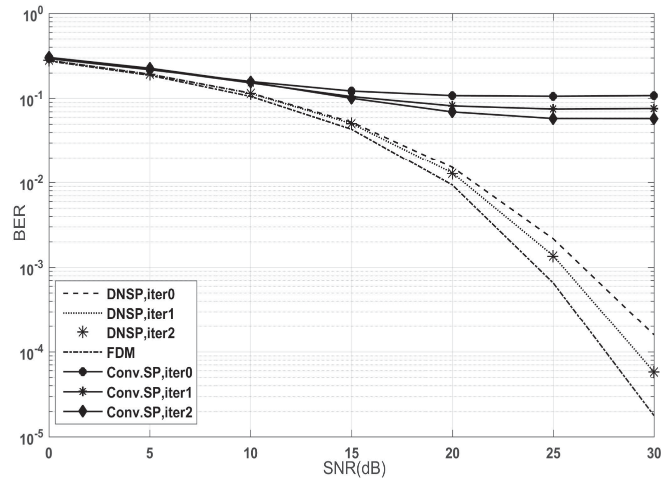

Fig. 4. BER of the DNSP scheme, Conventional superimposed pilot scheme and FDM scheme for $N_T = N_R = 8$ MIMO system

VII. Conclusion

A new channel estimation method is proposed for (MIMO-OFDM) system. This method depends on the usage of a special kind of superimposed pilots named data nulling superimposed pilots which cancels the mutual interference between the superimposed pilots and the data symbols by spreading data symbols on all subcarriers then introducing a distortion to the data symbols. So, this method can offer accurate channel estimation results as those of the methods that depend on FDM pilots but with higher spectral efficiency as there is no pilots symbol overhead but with some excess in the receiver complexity because an iterative detection algorithm is used to compensate the distortion occurred to data symbols. This method offers much better performance in terms of channel estimation accuracy, BER performance and computational complexity compared to the conventional superimposed pilots methods that allow the mutual interference between superimposed pilots and data symbols and uses an iterative joint channel estimation and data detection algorithms to mitigate interference between superimposed pilots and data symbols.

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