Classical Discrete-Time Fourier TransformBased Channel Estimation for MIMO-OFDM Systems in Vehicular Communications Technology

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ABSTRACT

In this document, we look at various time domain channel estimation methods with this constraint of null carriers at spectrumborders.We showin detail howto gauge the importance of the "border effect" depending on the number of null carriers, which may vary from one system to another. Thereby we assess the limit of the technique discussed when the number of null carriers is large. Finally the DFT with the truncated singular value decomposition (SVD) technique is proposed to completely eliminate the impact of the null subcarriers whatever their number. A technique for the determination of the truncation threshold for any MIMO-OFDM system is also proposed.

KEYWORD

Singular Value Decomposition,Discrete-Time Fourier Transform

1. INTRODUCTION

Today, there are many existing technologies designed to make vehicular road travel safer, easier and more enjoyable, using geographical positioning system, proximity sensors, multimedia communication, etc. The current data transmission requirements of these technologies, unfortunately, place great demand on both the algorithms and equipment, which often perform less than optimally, especially when having to interact with other vehicles. For example, GPS can trace a route to a specific location, but does so without taking into account some very important variables such as congestion caused by road conditions, high traffic volume and traffic accidents, which can entirely block one-lane traffic and affect two-lane traffic by almost 65% [1].

Presently, GPS permits users to obtain real-time location information. However, expanded communications among vehicles and with roadside infrastructure can substantially expand services drivers currently enjoy in the areas of traffic flow, safety, information (Internet), communications (VoIP) and comfort applications, among others [2].

According to [2] applications for vehicular communications include the following:

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- Proactive safety applications: geared primarily to improve driver reaction and decision making to avoid possible accidents (e.g. broadcast warnings from a vehicle that has ignored red stop light) or minimize the impacts of an imminent crash (automated braking systems).

- Traffic management applications: mainly implemented to improve traffic flow and reduce travel time, which is particularly useful for emergency vehicles.

- Traffic coordination and traffic assistance: principally concerned with improving the distribution and flow of vehicles by helping drivers pass, change lanes, merge and form columns of vehicles that maintain constant relative speeds and distances (platooning).

- Traveler Information Support: mainly focused on providing specific information about available resources and assistance persons require, making their traveling experience less stressful and more efficient.

The great increase in the demand for high-speed data services requires the rapid growthof mobile communications capacity. Orthogonal frequency division multiplexing (OFDM)provides high spectral efficiency, robustness to intersymbol interference (ISI), as well asfeasibility of low cost transceivers [1]. Multiple input multiple output(MIMO) systems offer the potential to obtain a diversity gain and to improve system capacity [2],[3], [4]. Hence the combination of MIMO andOFDM techniques (MIMO-OFDM) is logically widely considered in the new generation ofstandards for wireless transmission [5]. In these MIMO-OFDM systems,considering coherent reception, the channel state information (CSI) is required for recoveringtransmitted data and thus channel estimation becomes necessary.

Channel estimation methods can be classified into three distinct categories: blind channelestimation, semi-blind channel estimation and pilot-aided channel estimation. In thepilot-aided methods, pilot symbols known from the receiver are transmitted as a preambleat the beginning of the frame or scattered throughout each frame in a regular manner. On thecontrary, in blind methods, no pilot symbols are inserted and the CSI is obtained by relying onthe received signal statistics [6]. Semi-blind methods combine both the trainingand blind criteria [7] .In this paper, we focus our analysis on the time domain(TD) channel estimation technique using known reference signals. This technique is attractiveowing to its capacity to reduce the noise component on the estimated channel coefficients[8].

The vast majority of modern multicarrier systems contain null subcarriers at the spectrumextremities in order to ensure isolation from/to signals in neighboring frequency bands[9] as well as to respect the sampling theorem [10]. It wasshown that, in the presence of these null subcarriers, the TD channel estimation methodssuffer from the "border effect" phenomenon that leads to a degradation in their performance[9]. A TD approach based on pseudo inverse computation is proposedin [11] in order to mitigate this "border effect". Howeverthe degradation of the channel estimation accuracy persists when the number of the nullsubcarriers is large.

In this document, we look at various time domain channel estimation methods with thisconstraint of null carriers at spectrumborders.We showin detail howto gauge the importanceof the "border effect" depending on the number of null carriers, which may vary from onesystem to another. Finally the DFT with thetruncated singular value decomposition (SVD) technique is proposed to completely eliminatethe impact of the null subcarriers whatever their number. A technique for the determinationof the truncation threshold for any MIMO-OFDMsystem is also proposed.

The paper is organized as follows. Section 2 describes the studied Main goals of DFT based channel estimation,including the Noise reduction, Drawback of DFT based channel estimation in realistic system, MSE performance of DFT based channel estimation.in section 3.1 we have Pseudo inverse computation using SVD, Impact of pseudo inverse conditional number on channel estimation accuracy and in section 4 we describe DFT with truncated SVD channel estimation and finally we have a simulation result.

2.Main goals of DFT based channel estimation

2.1. Noise reduction

DFT-based channel estimation methods allow a reduction of the noise component owing tooperations in the transform domain, and thus achieve higher estimation accuracy [12],[8], [13]. In fact, after removing the unused subcarriers, theLS estimates are first converted into the time domain by the IDFT (inverse Discrete FourierTransform) algorithm. A smoothing filter is then applied in the time domain assuming thatthe maximum multi-path delay is kept within the cyclic prefix (CP) of the OFDM symbol. Asa consequence, the noise power[21] is reduced in the time domain. The DFT is finally applied toreturn to the frequency domain.

Fig.1.shows an example of an OFDM channel with a lot of sub-channels in a domain with time and frequency : each sub-channel amplitude $E_{t,f}$ l changes severely and deeply in the domain.

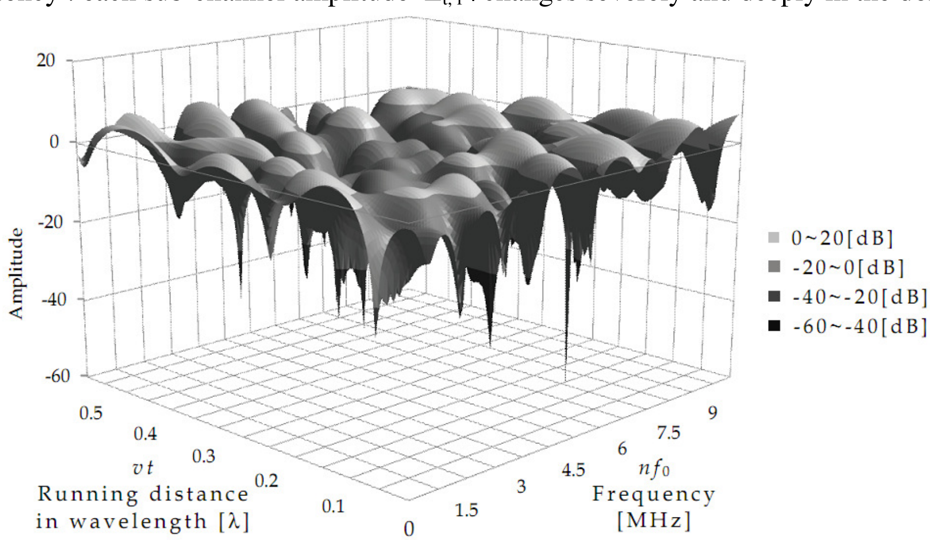

Fig.1. Example of OFDM channel in multipath condition

2.2.Drawback of DFT based channel estimation in realistic system

In a realistic context, only a subset of M subcarriers is modulated among the N due to the insertion of null subcarriers at the spectrum's extremities for RF mask requirements. The application of the smoothing filter in the time domain will lead to a loss of channel power when these non-modulated subcarriers are present at the border of the spectrum. That can be demonstrated by calculating the time domain channel response.

The time domain channel response of the LS estimated channel is given by (8). From (6) we can divide $h_{n,LS}^{IDFT}$ into two parts:

$$\mathrm{h}_{\mathrm{ij,n,LS}}^{\mathrm{IDFT}} = \sqrt{\frac{1}{\mathrm{N}}}\sum_{\mathrm{k}=\frac{\mathrm{N-M}}{2}}^{\frac{\mathrm{N+M}}{2}} \mathrm{H}_{\mathrm{ij,k,LS}} e^{\frac{2\pi j n k}{M}} = \mathrm{h}_{\mathrm{ij,n}}^{\mathrm{IDFT}} + \varphi_{\mathrm{ij,n}}^{\mathrm{IDFT}}$$

Where $\varphi_{\mathrm{ij,n}}^{\mathrm{IDFT}}$ is the noise component in the time domain and $h_{ij,n}^{IDFT}$ is the IDFT of the LS estimated channel without noise. This last component can be further developed as follows:

$$\mathrm{h}_{\mathrm{ij,n}}^{\mathrm{IDFT}} = \sqrt{\frac{1}{\mathrm{N}}}\sum_{\mathrm{k=N_b}}^{\mathrm{N_e}} \left(\sum_{\mathrm{l}=0}^{\mathrm{L_{ij}}-1} \mathrm{h}_{\mathrm{ij,l}} e^{-j\frac{2k\pi\tau}{N}} \right) e^{-j\frac{2k\pi n}{N}} = \sqrt{\frac{1}{\mathrm{N}}}\sum_{\mathrm{l}=0}^{\mathrm{L_{ij}}-1} \mathrm{h}_{\mathrm{ij,l}} \sum_{\mathrm{k=N_b}}^{\mathrm{N_e}} e^{-j\frac{2k\pi(\tau-n)}{N}}$$

whereN_p = $\frac{(\mathrm{N-M})}{2}$ and N_e=$\frac{(\mathrm{N-M})}{2}$– 1.

that if all the subcarriers are modulated M = N, the last term ofwill verify:

$$\sum_{\mathrm{k}=\frac{\mathrm{N-M}}{2}}^{\frac{\mathrm{N+M}}{2}-1} e^{\frac{-2\pi j k(\tau-n)}{M}} = \begin{cases} \mathrm{N_p} & \mathrm{n}=\tau \\ 0 & \text{otherwise} \end{cases}$$

Where$\tau = 0,1,\ldots,\mathrm{L_{ij}}-1$ and $\mathrm{n}=0,\ldots,\mathrm{M}-1$,we can safely conclude that $\mathrm{h}_{\mathrm{ij,n}}^{\mathrm{IDFT}}$ =0.

Assuming CP >L_{ij},we do not lose part of the channel power in the time domain by applying the smoothing filter of length CP. Nevertheless, when some subcarriers are not modulated at the spectrum borders, i.e. M < N,it can be expressed as:

$$\sum_{\mathrm{k}=\frac{\mathrm{N-M}}{2}}^{\frac{\mathrm{N+M}}{2}-1} e^{\frac{-2\pi j k(\tau-n)}{M}} = \begin{cases} M & n=\tau \\ \dfrac{1-e^{-j2\pi\frac{M}{N}(\tau-n)}}{1-e^{-j\frac{2\pi}{N}(\tau-n)}} & n\neq\tau \end{cases}$$

Where$\tau = 0,1,\ldots,L_{\mathrm{ij}}-1$ and n = 0, ...,M– 1.

The channel impulse response $\mathrm{h}_{\mathrm{ij,n}}^{\mathrm{IDFT}}$ can therefore be rewritten in the following form:

$$\mathrm{h}_{\mathrm{ij,n}}^{\mathrm{IDFT}} = \frac{1}{\sqrt{N}}\begin{cases} M.h_{ij,l=n} + \displaystyle\sum_{l=0,1\neq n}^{L_{ij}-1} h_{ij,l}\frac{1-e^{-j2\pi\frac{M}{N}(\tau-n)}}{1-e^{-j\frac{2\pi}{N}(\tau-n)}} & n<L_{ij} \\ \displaystyle\sum_{l=0,1\neq n}^{L_{ij}-1} h_{ij,l}\frac{1-e^{-j2\pi\frac{M}{N}(\tau-n)}}{1-e^{-j\frac{2\pi}{N}(\tau-n)}} & L_{ij}-1<n<M \end{cases}$$

We can observe that $\mathrm{h}_{\mathrm{ij,n}}^{\mathrm{IDFT}}$ is not null for all the values of n due to the phenomenon called here "Inter-Taps Interference (ITI)". Consequently, by using the smoothing filter of length CP in the time domain, the part of the channel power contained in samples n = CP, ...,M−1islost.This loss of power leads to an important degradation on the estimation of the channel response. In OFDM systems, [9] shows that when null carriers are inserted at the spectrum extremities,the performance of the DFT based channel estimation is degraded especially at the borders of the modulated subcarriers [9],[18]. This phenomenon is called the “border effect”. This “border effect” phenomenon is also observed in MIMO context.

In order to evaluate the DFT based channel estimation, the mean square error (MSE) performance for the different modulated subcarriers is considered in the following subsection.

2.3. MSE performance of DFT based channel estimation

In MIMO-OFDM context with $N_t \leq \frac{M}{CP}$ transmit antennas and N_r receive antennas, the (MSE) on the k -th subcarrier is equal to:

$$MSE(k) = \frac{\sum_{i=0}^{N_t} \quad \sum_{i=0}^{N_r} E\,[||\,\widehat{H(k)} - H(k)||^2]}{N_t N_r}$$

Where$\widehat{H(k)}$ and H(k) represent the estimated frequency channel response and the ideal one respectively.

MSE performance are provided here over frequency and time selective MIMO SCME (spatial channel model extension) channel model typical of macro urban propagation [16]. The DFT based channel estimation is applied to a 2 ×2 MIMO system with the number of FFT points set equal to 1024. The orthogonally between pilots is obtained using null symbol insertion described in 2 and interpolation is performed to obtain channel estimates[20],[14] for all modulated subcarriers.First, when all the subcarriers are modulated (N = M = 1024), there is no “border effect”and the MSE is almost the same for all the subcarriers. This is due to the fact that all the channel power is retrieved in the first CP samples of the impulse channel response.

However when null subcarriers are inserted on the edge of the spectrum (N = M), the MSE performance is degraded by the loss of a part of the channel power[22] in the time domain and then the “border effect” occurs. It is already noticeable that the impact of the “border effect” phenomenon increases greatly with the number of null subcarriers.

3. DFT with pseudo inverse channel estimation

Classical DFT based channel estimation described in the previous section can be also expressed in amatrix form. The unitary DFT matrix F of size N × N is defined with the following expression:

$$F = \begin{bmatrix} 1 & \cdots & 1 \\ \vdots & \ddots & \vdots \\ 1 & \cdots & W_N^{(N-1)(N-1)} \end{bmatrix}$$

where $W_N^i = e^{-j\frac{2\pi i}{N}}$.

To accommodate the non-modulated subcarriers, it is necessary to remove the rows of the matrix F corresponding to the position of those null subcarriers. Furthermore, in order to reduce the noise component in the time domain by applying smoothing filtering, only the first CP columns of F are used (Fig.2). Hence the transfer matrix becomes:

$$\acute{F} \in C^{\acute{M} \times CP} : \acute{F} = F\left(\frac{N-M}{2} : \frac{N-M}{2} - 1, 1{:}\,CP\right)$$

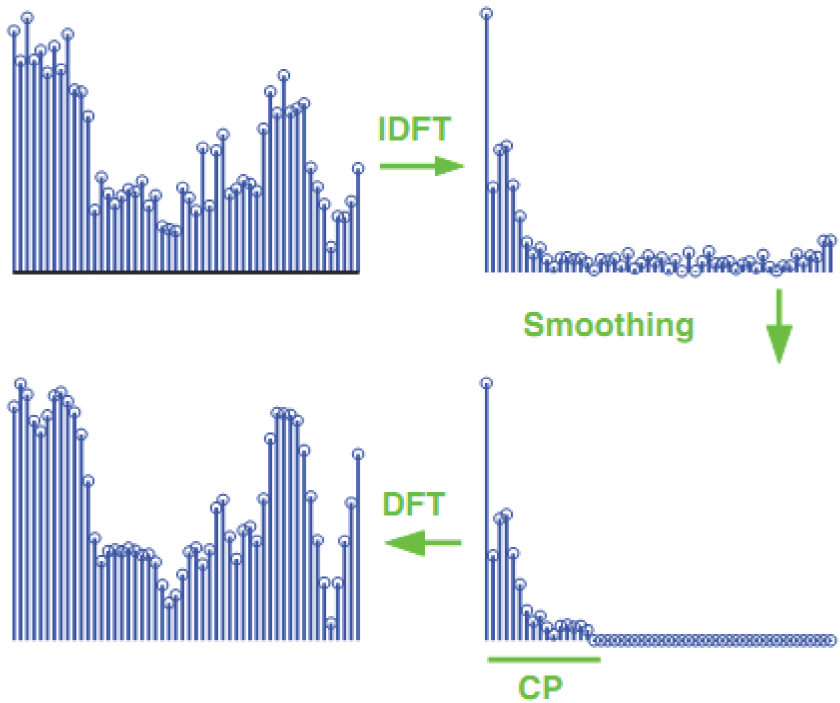

Fig.2. Smoothing using DFT.

We can then express the impulse channel response, after the smoothing filter, in a matrix form:

$$h_{ij,LS}^{IDFT} = \acute{F}^{H} H_{ij,Ls} \text{ where } h_{ij,LS}^{IDFT} \in C^{CP \times 1}, H_{ij,LS} \in C^{M\times 1}$$

To reduce the “border effect” [11] propose the use of the following minimization problem.
The use of the pseudoinverse allows the minimization of the power loss in the time domain which was at the originof the “border effect”. The pseudo inverse of the matrix $\widehat{F}$which is noted $\widehat{F} \in C^{CP \times M}$. It can be used to transform the LS estimates of$\acute{F}^{H}$ as previously proposed.

This technique is often used for the resolution of linearequations system due to its capacity to minimize the Euclidean norm andthentotendtowardsthe exact solution.The pseudo inverse$\acute{F}^{\top}$of $\acute{F}$ is defined as unique matrix satisfying all fourfollowing criteria.

$$\begin{cases} 1: \acute{F}\,\acute{F}^{\top}\acute{F} = \acute{F} \\ 2: \acute{F}^{\top}\acute{F}\,\acute{F}^{\top} = \acute{F}^{\top} \\ \left(\acute{F}^{\top}\acute{F}\right)^{H} = \acute{F}^{\top}\acute{F} \end{cases}$$

3.1. Pseudo inverse computation using SVD

The pseudo inverse can be computed simply and accurately by using the singular valuedecomposition [17]. Applying SVD to the matrix$\acute{F}$ consists in decomposing $\acute{F}$ inthe following form:

$\acute{F} = USV^H$ where $U \in C^{M\times M}$ *and* $V \in C^{CP\times CP}$ are unitary matrices and $S \in C^{M\times CP}$is a diagonal matrixwith non-negative real numbers on the diagonal, called singular values.

The pseudo inverse of the matrix $\acute{F}$ with singular value decomposition is:

$$\acute{F}^{\dagger} = VS^{\dagger}U^{H}$$

3.2. Impact of pseudo inverse conditional number on channel estimation accuracy

The accuracy of the estimated channel response depends on the calculation of the pseudoinverse $\acute{F}^{\dagger}$The conditional number (CN) can give an indication of the accuracy of thisoperation [19]. The higher the CN is, the more the estimated channel responseis degraded.

3.2.1. Definition of the conditional number

It is defined as the ratio between the greatest and the smallest singular values of the transfer matrix $\acute{F}$ By noting s$\in C^{CP\times 1}$ the vector which contains the elements (the singular values)on the diagonal of the matrix S,CN is expressed as follows:

$$N = \frac{\max(s)}{\min(s)}$$

where max(s) and min(s) give the greatest and the smallest singular values respectively.

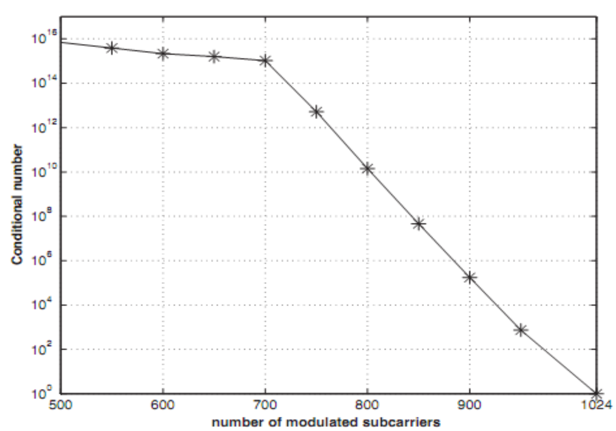

Fig.3. Conditional number of $\acute{F}$ versus the number of modulated subcarriers (M)whereCP = 72 and N = 1024.

3.2.2. Behavior of the conditional number

It only evolves according to the number of modulated subcarriers. Fig.3 shows this behaviorfor different values of M when N = 1024 and CP = 72. When all the subcarriers aremodulated, the

CN is equal to 1. However when null carriers areinserted at the edge of the spectrum (M < N), the CN increases according to the numberof non-modulated subcarriers (N – N) and can become very high.We can note that i f M = 600 as in 3GPP standard (where N = 1024, CP = 72 and M = 600),the CN is equal to 2.17 1015.

4. DFT with truncated SVD channel estimation

The DFT with pseudo inverse technique previously described allows the reduction of the"border effect". But it remains insufficient when the number of null subcarriers is large. Tofurther reduce this "border effect", it is necessary to attain a small CN in this operation. Theaim of the proposed approach is to reduce both the "border effect" and the noise componentby considering only the most significant singular values of matrix S.

4.1. Principle of DFT with truncated SVD channel estimation

To reduce the CN, the lowest singular values have to be eliminated.Hence, any singular valuesmaller than an optimized threshold is replaced by zero. The principle isdepicted in Fig.4. The SVD calculation of matrix F provides the matrices U, S and V (20). The matrix S becomes S_{Th} where$T_h$is the number of considered singular values.

The channel impulse response after the smoothing process in the time domain can thus beexpressed as follows:

$$h_{ij,LS}^{TSVD} = \acute{F}_{T_h}^{T} H_{ij,LS} = V\, S_{T_h}^{T} U^{H} H_{ij,LS}$$

$$h_{ij,LS}^{TSVD} = \begin{cases} \acute{F}_{T_h}^{T} H_{ij,LS} = V\, S_{T_h}^{T} U^{H} H_{ij,LS} & n<CP \\ 0 & \text{otherwise} \end{cases}$$

where n = 0, ...,M− 1.

Finally $\acute{F}$of size (CP × M) is used to return to the frequency domain.

It is important to note the two following points:

- On the one hand, the "border effect" is obviously further reduced due to the reduction ofthe CN.
- On the other hand, the suppression of the lowest singular values allows the noisecomponent in the estimated channel response to be reduced.
- The rank of the matrix $\acute{F}$is CP, which means that the useful power of the channel is distributed into CP virtualpaths with singular values as weightings.
- The paths corresponding to the weakest singularvalues are predominated by noise and their elimination benefits the noise componentreduction.

4.2. Threshold determination for DFT with truncated SVD

4.2.1. Discussion

The choice T_h (∈ 1, 2, ...,CP) can be viewed as a compromise between the accuracy ofthe pseudo-inverse calculation and the CN magnitude. Its value will depend only on thesystemparameters:

- The number and position of the modulated subcarriers (M)
- The smoothing window size (CP)
- The number of FFT points (N). All these parameters are predefined andare known prior to any

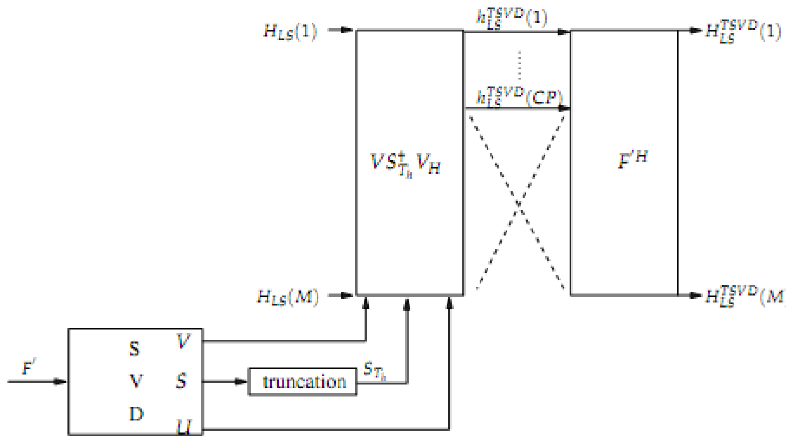

Fig. 4. Block diagram of DFT with Truncated SVD.

channel estimation implementation. It is thus feasible to optimize the T_h value prior to any MIMO-OFDM system implementation.

5. Simulation results

Perfect time and frequency synchronizations are assumed. Monte Carlo simulation results interms of bit error rate (BER) versus $\frac{\mathrm{E_b}}{\mathrm{N_0}}$are presented here for the different DFT based channelestimation methods: classical DFT, DFT with pseudo inverse and DFT with truncated SVD.

The $\frac{\mathrm{E_b}}{\mathrm{N_0}}$ value can be inferred from the signal to noise ratio (SNR):

$$\frac{\mathrm{E_b}}{\mathrm{N_0}} = \frac{N_t\sigma_s^2}{mR_cR_M\sigma_s^2} = \frac{N_t}{mR_cR_M}.SNR$$

where σ_{noise}^2 and σ_x^2represent the noise and signal variances respectively. R_c, R_m and mrepresent the coding rate, the MIMO scheme rate and the modulation order respectively.

Fig.5 show the performance results in terms of BER versus $\frac{\mathrm{E_b}}{\mathrm{N_0}}$ or perfect, leastsquare (LS), classical DFT, DFT with pseudo inverse (DFT $- T_h$ = CP)and DFT withtruncated SVD for channel estimation methods in 3GPP/LTE and 802.11n systemenvironments respectively.

In the context of 3GPP/LTE, the classical DFT based method presents poorer results due to thelarge number of null carriers at the border of the spectrum (424 among 1024). The conditionalnumber is as a consequence very high (CN = 2.17 1015) and the impact of the "bordereffect" is very important. For this reason, the DFT with pseudo inverse (DFT $- T_h$ = CP = 72) cannot greatly improve the accuracy of the estimated channel response. The classicalDFT and the DFT with pseudo inverse estimated channel responses are thus considerablydegraded compared to the LS one. The DFT with a truncated SVD technique and optimizedT_h (T_h=46,45,44) greatly enhances the accuracy of the estimated channel response by bothreducing the noise component and eliminating the impact of the "border effect" (up to 2dBgain compared to LS). This last method presents an error floor when T_h = 55duetothefactthat the "border effect" is still present and very bad results are obtained when T_h is small(T_h = 43) due to the large loss of energy.

Comparatively, in the context of 802.11n, the number of null carriers is less important and theclassical DFT estimated channel response is not degraded even if it does not bring about anyimprovement compared to the LS. The pseudo inverse technique completely eliminates the"border effect" and thus its estimation (DFT − T_h = CP = 16) is already very reliable.

DFTwith a truncated SVD channel estimation method does not provide any further performanceenhancement as the "border effect" is quite limited in this system configuration.

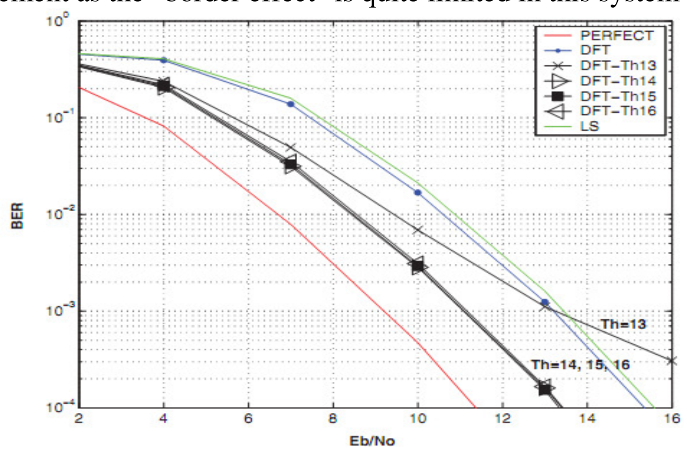

Fig. 5. BER versus $\frac{E\_b}{N_0}$ for classical DFT, DFT pseudo inverse (T_h = CP = 16) and DFT withtruncated SVD (T_h = 15, T_h= 14 and T_h = 13) based channel estimation methods in 802.11ncontext. N_t = 2, N_r = 2, N = 1024, CP = 72 and M = 600

6. Conclusion

Several channel estimation methods have been investigated in this paper regarding theMIMO-OFDM system environment. All these techniques are based on DFT and are soprocessed through the time transform domain. The key system parameter, taken into accounthere, is the number of null carriers at the spectrum extremities which are used on the vastmajority of multicarrier systems. Conditional number magnitude of the transform matrix hasbeen shown as a relevant metric to gauge the degradation on the estimation of the channelresponse. The limit of the classical DFT and the DFT with pseudo inverse techniques has beendemonstrated by increasing the number of null subcarriers which directly generates a highconditional number. The DFT with a truncated SVD technique has been finally proposed tocompletely eliminate the impact of the null subcarriers whatever their number. A techniquewhich allows the determination of the truncation threshold for any MIMO-OFDM system isalso proposed.

The truncated SVD calculation is constant and depends only on the systemparameters: the number and position of the modulated subcarriers, the cyclic prefix size andthe number of FFT points. All these parameters are predefined and are known at the receiverside and it is thus possible to calculate the truncated SVD matrix in advance.

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