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Alain Y. Kibangou *1, Laurent Ros #2, Cyrille Siclet #3

* *Control Systems Department, GIPSA-Lab, University Joseph Fourier, CNRS*
BP 46, F-38402 Grenoble Cedex, France
[1] alain.kibangou@ujf-grenoble.fr

Image and Signal Department, GIPSA-Lab, University Joseph Fourier, Grenoble-INP
BP 46, F-38402 Grenoble Cedex, France
[2] laurent.ros@gipsa-lab.grenoble-inp.fr
[3] cyrille.siclet@gipsa-lab.grenoble-inp.fr

***Abstract*—A new scheme for Doppler estimation and data detection for OFDM underwater acoustic communications is derived in this paper. We design the OFDM symbol as a concatenation of two sub-blocks. The first one is shorter and carries out few data with few subcarriers whereas the second one contains both informative and pilot symbols. The first sub-block serves to estimate the Doppler scaling factor while the pilots in the second sub-block are used for estimating the equivalent channel. Both estimation approaches are based on high resolution methods for solving harmonic retrieval problems in time and frequency domains respectively. Each sub-block contains a cyclic suffix, which allows having more data to be processed. The effectiveness of the proposed scheme is evaluated by means of simulation results.**

I. Introduction

Underwater acoustic channels are typically wideband in nature due to the small ratio of carrier frequency to the signal bandwidth, which introduces frequency-dependent Doppler shifts [1]. They also exhibit several propagation paths, so that the received signal is equivalent to the sum of several signals with different amplitude and time delays. Multipath delay and Doppler effects constitute the main obstacles to robust underwater acoustic communication. One common approximation is to treat the channel as having a common Doppler scaling factor on all propagation paths.

Orthogonal Frequency Division Multiplexed (OFDM) signals are particularly attractive because they lead to relatively simple channel estimation and data detection algorithms. Indeed, since subcarrier only experiences flat fading, complex time-domain equalizers are not necessary [2]. However, OFDM signals are particularly vulnerable to Doppler effects that destroy the orthogonality between subcarriers and thus induce intercarrier interference (ICI). In order to adequately recover the transmitted information, algorithms at the receiver must include estimation and compensation of the Doppler scaling factor, channel estimation, and information symbols estimation.

For estimating the Doppler scaling factor, several approaches have been suggested in the literature (see [3] for example). In the OFDM case, they are based on the use of preamble and postamble of a packet consisting of multiple OFDM blocks [1] or by exploiting correlation induced by the cyclic prefix [4]. Then, the received signal is resampled by using a sampling period related to the estimated Doppler scaling factor. Recently, the authors introduced a new scheme for estimating both Doppler and channel parameters by solving harmonic retrieval problems [5]. In this paper, we propose an improved version of the previous algorithm with a better use of the bandwidth and without explicitly estimating the paths gain and delay. Indeed, the lack of robustness to noise of the channel parameter estimators was a major drawback of the approach in [5], [6]. Herein, instead of using a zero padding scheme, we introduce a cyclic suffix in the OFDM signal design. As in [5], the received data are processed block-by-block. The advantage of the proposed scheme is to avoid data resampling and residual CFO estimation and compensation.

II. System model

We consider an OFDM transmission system where each OFDM block is constituted with two sub-blocks of respective length T_u and T_d, with $T_d/T_u \in \mathbb{N}$, as depicted in Fig. 1. Each sub-block is followed by a cyclic suffix (CS) of duration T_g. Therefore, the overall OFDM block duration is $T = T_u + T_d + 2T_g$.

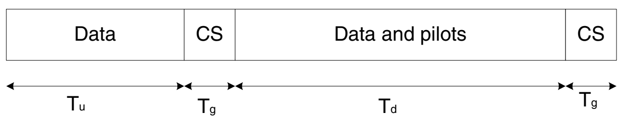

Fig. 1. Structure of the OFDM block

We denote by c_q the data symbols in the first sub-block. They are all informative and thus unknown to the receiver. In the second sub-block, all data symbols d_k are not informative; some of them are known to the receiver and will serve to estimate the equivalent channel parameters. In baseband, the signal $s(.)$ to be transmitted can be viewed as a mixture of two signals $s_u(.)$ and $s_d(.)$, each one corresponding respectively to the first and to the second sub-blocks:

$$s(t) = s_u(t) + s_d(t) \tag{1}$$

with

$$s_u(t) = \eta \sum_{q\in\mathcal{K}_u} c_q e^{j2\pi qQ\Delta f t}\Pi_{T_u+T_g}(t)$$
$$s_d(t) = \sum_{k\in\mathcal{K}_d} d_k e^{j2\pi k\Delta f t}\Pi_{T_d+T_g}(t-T_u-T_g)$$

where $\Delta f = 1/T_d$ denotes the minimal frequency spacing between consecutive subcarriers, Π_T allows taking the CS operation into account, i.e. $\Pi_T(t) = 1,\ t \in [0,T]$, and $\Pi_T(t) = 0$, otherwise, η allows controlling the power of the first sub-block so that both sub-blocks have the same power, Q is an integer that allows increasing the frequency spacing between the subcarriers in $s_u(.)$, $\mathcal{K}_u$ (resp. $\mathcal{K}_d$) the set of subcarriers in the first (resp. second) sub-block.

In the sequel, we consider that $\mathcal{K}_u \subset \mathcal{K}_d$, meaning that some of the subcarriers of the second sub-block are reused in the first sub-block. By denoting K_u (resp. K_d) the cardinality of $\mathcal{K}_u$ (resp. $\mathcal{K}_d$), i.e. the number of used subcarriers, we get $K_u < K_d$, $\mathcal{K}_d = \{-K_d/2,\cdots,K_d/2+1\}$, and $\mathcal{K}_u = \{q_0,\cdots,q_0+K_u-1\}$, for a given integer q_0.

In passband, the analytical representation of the signal to be transmitted is $x(t) = s_u(t)e^{j2\pi f_c t} + s_d(t)e^{j2\pi f_c t}$, $t \in [0,T]$, with f_c the central carrier frequency. This signal is transmitted through a multipath underwater channel whose impulse response is given by:

$$h(t,\tau) = \sum_{p=1}^{P} A_p(t)\delta(\tau-\tau_p(t)),$$

where $A_p(t)$ and $\tau_p(t)$ are respectively the gain and the delay associated with the pth path. The following assumptions are adopted:

- All paths are affected by a similar Doppler scaling factor a such that $\tau_p(t) = \tau_p - at$.
- The path delays τ_p, the gains A_p, and the Doppler scaling factor a are constant over the block duration T.
- The guard interval T_g is chosen so that $T_g > \max\{\tau_p\} = \tau_{max}$.

Therefore, the received signal results on a mixture of scaled, dilated or compressed, and delayed versions of the original OFDM block as shown in Fig. 2. Hence, we can define two

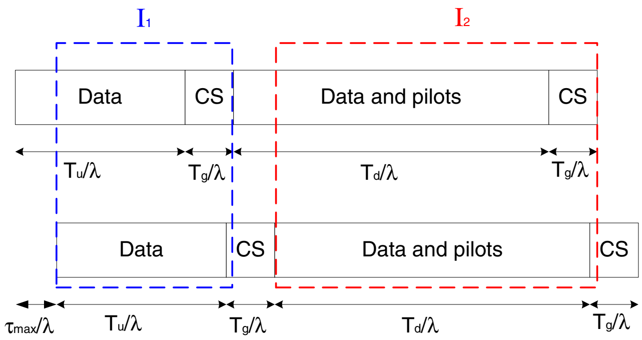

Fig. 2. Multipath and Doppler effect on the OFDM block

time windows I_1 and I_2 where the received signal strictly depends on the first and second sub-blocks respectively. These time windows are defined by: $I_1 = \left[\frac{\tau_{max}}{\lambda}, \frac{T_u+T_g+\tau_{min}}{\lambda}\right]$ and $I_2 = \left[\frac{T_u+T_g+\tau_{max}}{\lambda}, \frac{T+\tau_{min}}{\lambda}\right]$, where $\tau_{min} \le \tau_p \le \tau_{max}$, $p = 1,\cdots,P$, and $\lambda = 1+a$.

In I_1 and I_2, the analytical representation of the received signal is respectively given as follows:

$$y(t) = \eta \sum_{q=q_0}^{K_u+q_0-1} B_q c_q e^{j2\pi(1+a)\varphi_q t} + w(t), \quad t \in I_1 \quad (2)$$
$$y(t) = \sum_{k=-K_d/2}^{K_d/2-1} B_k d_k e^{j2\pi(1+a)f_k t} + w(t), \quad t \in I_2 \quad (3)$$

where $\varphi_q = f_c + qQ\Delta f$, $f_k = f_c + k\Delta f$, $w(t)$ denoting the additive noise, $B_q = \sum_{p=1}^{P} A_p e^{-j2\pi\varphi_q\tau_p}$, and

$$B_k = \sum_{p=1}^{P} A_p e^{-j2\pi f_k \tau_p}. \quad (4)$$

In both I_1 and I_2, the received signal $y(.)$ can be respectively viewed as a mixture of K_u and K_d harmonics with constant magnitudes. In the sequel, we intend to solve the problem of Doppler scale factor estimation as a Harmonic retrieval one. We make use of these two time windows for first estimating the Doppler scaling factor and then recovering the informative symbols contained in the OFDM block. Since the time windows involved in the estimation process depend on λ, we assume that λ is bounded, such that $\lambda_{min} \le \lambda \le \lambda_{max}$. One can note that λ_{max} and λ_{min} are related to the maximal and minimal velocities of the underwater vehicles, which can be *a priori* known. In addition, we assume that τ_{max} is known and $\tau_{min} = 0$. Therefore the involved time windows are modified as follows

$$I_1 = \left[\frac{\tau_{max}}{\lambda_{min}}, \frac{T_u+T_g}{\lambda_{max}}\right] \quad I_2 = \left[\frac{T_u+T_g+\tau_{max}}{\lambda_{min}}, \frac{T}{\lambda_{max}}\right]. \quad (5)$$

Owing to the CS, the time windows to be used are larger than those obtained when using a ZP scheme as in [6]. Obviously, with more observations, an improvement of the resulting estimation method is expected.

III. Doppler Estimation

The harmonic retrieval problem has been extensively studied in the literature (see [7], [8], [9], [10] for example). In general the magnitude are assumed to be constant as in the case for $y(.)$ in I_1 and I_2. For solving (2) as a harmonic retrieval problem, we make use of the high resolution method called HTLS (*Hankel Total Least Squares*) [11]. This algorithm can be viewed as a special case of the ESPRIT algorithm [9] and also as a Total Least Squares variant of Kung et al.'s algorithm [12]. Resorting to a high resolution method is mandatory since the Doppler is to be estimated from a very slight deviation of the actual subcarrier frequencies. A search on discretized space of frequencies could need a very short step of discretization, implying a huge complexity. In other hand, FFT based methods have a lower bound of resolution incompatible in such a situation.

In the noiseless case, we consider the N samples y_n, $n = n_0, \cdots, N + n_0 - 1$, of the first portion (2) of the received signal:

$$y_n = \eta \sum_{q=q_0}^{K_u+q_0-1} c_q B_q e^{j2\pi\varphi_q(1+a)nT_e} = \eta \sum_{q=q_0}^{K_u+q_0-1} c_q B_q \phi_q^n, \tag{6}$$

T_s being the sampling period. In a harmonic retrieval problem, we aim to estimate both the poles $\phi_q = e^{j2\pi\varphi_q(1+a)T_s}$ and the magnitudes $\eta c_q B_q$ of the K_u harmonics.

For this purpose, let us first build with the samples y_n the following $L \times M$ Hankel matrix

$$\mathbf{Y} = \begin{pmatrix} y_{n_0} & y_{n_0+1} & y_{n_0+2} & \cdots & y_{n_0+M-1} \\ y_{n_0+1} & y_{n_0+2} & \iddots & \cdots & y_{n_0+M} \\ y_{n_0+2} & \iddots & \iddots & \cdots & y_{n_0+M+1} \\ \vdots & \iddots & \iddots & \vdots & \vdots \\ y_{n_0+L-1} & y_{n_0+L} & \cdots & \cdots & y_{n_0+N-1} \end{pmatrix},$$

with $N = L + M - 1$, $L > K_u$, $M \geq K_u$. It admits the Vandermonde decomposition:

$$\mathbf{Y} = \mathbf{S}_1 diag(\boldsymbol{\alpha}) \mathbf{T}_1^T, \tag{7}$$

where $\mathbf{S}_1$ and $\mathbf{T}_1$ are two Vandermonde matrices with ϕ_q as generators:

$$\mathbf{S}_1 = \begin{pmatrix} 1 & \cdots & 1 \\ \phi_{q_0} & \cdots & \phi_{K_u+q_0-1} \\ \vdots & \ddots & \vdots \\ \phi_{q_0}^{L-1} & \cdots & \phi_{K_u+q_0-1}^{L-1} \end{pmatrix} \in \mathbb{C}^{L \times K_u}$$

$$\mathbf{T}_1 = \begin{pmatrix} 1 & \cdots & 1 \\ \phi_{q_0} & \cdots & \phi_{K_u+q_0-1} \\ \vdots & \ddots & \vdots \\ \phi_{q_0}^{M-1} & \cdots & \phi_{K_u+q_0-1}^{M-1} \end{pmatrix} \in \mathbb{C}^{M \times K_u},$$

while $\boldsymbol{\alpha}$ contains the complex magnitudes of the K_u harmonics:

$$\boldsymbol{\alpha} = \eta \begin{pmatrix} c_{q_0} B_{q_0} \phi_{q_0}^{n_0} & \cdots & c_{K_u+q_0-1} B_{K_u+q_0-1} \phi_{K_u+q_0-1}^{n_0} \end{pmatrix}^T$$

We intend to estimate the Doppler scaling factor from the angle of the generators ϕ_q. In order to avoid any ambiguity since angles are obtained up to 2π, the sampling period can be selected such that the angle of the generators belongs to $[-\pi, \pi]$. For this purpose, it is enough to select T_s such that:

$$T_s \leq \frac{1}{2\lambda_{max} \max\{\varphi_q\}}. \tag{8}$$

It should be noted that this condition corresponds precisely to the Nyquist-Shannon sampling theorem condition applied to the Doppler scaled spectrum of the received signal. We can also derive the following proposition:

Proposition 1: If the sampling period is such that $T_s \leq \frac{1}{2\lambda_{max} \max\{\varphi_q\}}$, the Vandermonde matrices $\mathbf{S}_1 \in \mathbb{C}^{L \times K_u}$ and $\mathbf{T}_1 \in \mathbb{C}^{M \times K_u}$, $L > K_u$, $M \geq K_u$, with $\phi_q = e^{j2\pi\varphi_q(1+a)T_s}$ as generators are full column rank.

Proof: The Vandermonde matrices loose their rank if at least two generators have the same angle modulo 2π. Owing to condition (8) on the sampling period we know that the angles of the generators belong to $[0, \pi]$. Therefore, the angle of the generator are all distinct modulo 2π. Thus, the resulting Vandermonde matrices are full column rank. ∎

Let us denote by $\hat{\phi}_q$ the poles estimated with the HTLS method (see [5] for details). If the sampling period T_s is chosen according to (8), then the angle of $\hat{\phi}_q$, denoted $\angle\hat{\phi}_q$, belongs to $[-\pi, \pi]$. We deduce the following estimator for the Doppler scaling factor:

$$\hat{a} = -1 + \frac{1}{K_u} \sum_{q=q_0}^{K_u+q_0-1} \frac{\angle\hat{\phi}_q}{2\pi\varphi_q T_e}. \tag{9}$$

One can note that such a Doppler scaling factor estimator is completely blind. However, since it is a SVD-based estimator, the computation load becomes significant when increasing K_u. So, we suggest to use few active subcarriers K_u. The choice of the optimal number is still an open question. Moreover, when dealing with subspace-based methods, the dimension of the initial matrix is very important. When applying HTLS, it has been shown that very good results are obtained for $L \times N - L + 1$ matrices where $L \in [N/3, 2N/3]$ [13].

IV. Data detection

Now, for recovering the transmitted information symbols, we assume that some pilot symbols have been inserted in the second OFDM sub-block. We first estimates channel parameters using pilot symbols before recovering the informative symbols. Recall that the symbols c_q are informative whereas only some of d_k are also informative. The samples corresponding to the time interval I_2 are given by:

$$y_n = y(t)|_{t=nT_e} = \sum_{k=-K_d/2}^{K_d/2-1} d_k B_k e^{j2\pi(1+a)f_k nT_e}.$$

In matrix form, we get:

$$\mathbf{y} = \mathbf{H}\mathbf{b}, \tag{10}$$

with $\mathbf{y} = \begin{pmatrix} y_{n_1} & \cdots & y_{n_1+N_1-1} \end{pmatrix}^T$, $\mathbf{b} = \begin{pmatrix} b_{-K_d/2} & \cdots & b_{K_d/2-1} \end{pmatrix}^T$, $b_k = d_k B_k$, $\hat{\phi}_k = e^{j2\pi(1+\hat{a})f_k T_e}$, and

$$\mathbf{H} = \begin{pmatrix} \hat{\phi}_{-K_d/2}^{n_1} & \cdots & \hat{\phi}_{K_d/2-1}^{n_1} \\ \vdots & \ddots & \vdots \\ \hat{\phi}_{-K_d/2}^{n_1+N_1-1} & \cdots & \hat{\phi}_{K_d/2-1}^{n_1+N_1-1} \end{pmatrix} \in \mathbb{C}^{N_1 \times K_d}.$$

It is straightforward to check that $\mathbf{H}$ is a full column rank matrix if $N_1 > K_d$ and $T_s \leq \frac{1}{2\lambda_{max} \max\{f_k\}}$. As a consequence, we can compute the least squares (LS) solution of (10) as:

$$\hat{\mathbf{b}} = \mathbf{H}^{\dagger}\mathbf{y} \tag{11}$$

where $\mathbf{H}^{\dagger}$ denotes the pseudo-inverse operator of the matrix $\mathbf{H}$. Since the entries of $\hat{\mathbf{b}}$ result on the product of a symbol d_k and the equivalent channel parameters B_k, without lack of generality, by assuming that the subcarriers numbered $k = 0, 1, \cdots, K_p - 1$, $K_p < K_d/2$, are devoted to pilot symbols, i.e. d_k,$k = 0, 1, \cdots, K_p - 1$, are known to the receiver, we get:

$$\hat{B}_k = \hat{b}_k / d_k. \tag{12}$$

Now, the question is: how constructing $\hat{B}_k$ associated with the information symbols from those estimated using the K_p pilots?

First, we can note that (4) can also be written as

$$B_k = \sum_{p=1}^{P} \gamma_p \zeta_p^k, \quad \gamma_p = A_p e^{-j2\pi f_c \tau_p}, \quad \zeta_p = e^{-j2\pi \Delta_f \tau_p}. \tag{13}$$

The expression above is particularly meaningful. Indeed, B_k can be viewed as a mixture of exponentials in the frequency domain. We can therefore obtain both γ_p and ζ_p by solving a harmonic retrieval problem.

Assuming that the number of significative paths P is known and provided $K_p \geq 2P$, we make use of the HTLS method for solving the problem (13) where B_k is replaced by its estimated value $\hat{B}_k$. For this purpose, we build an $L \times M$ Hankel matrix $\mathbf{B}$, with $K_p = L + M - 1$, $L > P$, $M \geq P$, which admits the following Vandermonde decomposition:

$$\mathbf{B} = \mathbf{S}_2 diag(\boldsymbol{\gamma}) \mathbf{T}_2^T. \tag{14}$$

with $\mathbf{S_2}$ and $\mathbf{T}_2$ some Vandermonde matrices having ζ_p as generators, and $\boldsymbol{\gamma}$ the vector with γ_p as entries.

Let us denote $\hat{\zeta}_p$ the estimated poles. By replacing the poles by their estimated values in $\mathbf{S}_2$ and $\mathbf{T}_2$, we can solve the following equation:

$$vec(\mathbf{B}) = (\mathbf{T}_2 \odot \mathbf{S}_2) \boldsymbol{\gamma}, \tag{15}$$

$vec(.)$ denoting the vectorization operator whereas $\odot$ stands for the Khatri-Rao product, i.e. a columnwise Kronecker product.

The entries of the least squares solution of (15), denoted $\hat{\gamma}$, can be used for computing $\hat{B}_k$ for any value of k as $\hat{B}_k = \sum_{p=1}^{P} \hat{\gamma}_p \hat{\zeta}_p^k$. Therefore, we can deduce the information symbols in the second sub-block as:

$$\hat{d}_k = \frac{\hat{b}_k}{\sum_{p=1}^{P} \hat{\gamma}_p \hat{\zeta}_p^k}. \tag{16}$$

while those of the first sub-block can be obtained as

$$\hat{c}_q = \frac{\hat{\alpha}_q}{\eta e^{j2\pi\varphi_q(1+\hat{a})n_0 T_s} \sum_{p=1}^{P} \hat{\gamma}_p \hat{\zeta}_p^q}. \tag{17}$$

$\hat{\alpha}_q$ being the corresponding entry of the least squares solution of the vectorized version of equation (7), i.e. $vec(\mathbf{Y}) = (\mathbf{T}_1 \odot \mathbf{S}_1) \boldsymbol{\alpha}$.

V. Simulation results

In these simulations, the range of frequency used by the underwater vehicles was $[16kHz - 27kHz]$. The Doppler scaling factor was set to $a = 8 \times 10^{-4}$. We assume that $|a_{max}| = 10^{-3}$, meaning that the maximal relative speed is 1.5 m/s. The carrier frequency was set equal to $f_c = 21$ kHz, whereas the guard interval was $T_g = 10$ ms. We used $K_d = 256$ subcarriers for the second OFDM sub-block. The duration of this sub-block is $T_d = K_d/B$, with $B = 11$ kHz, whereas that of the first sub-block is $T_u = T_d/10$. Hence, the duration of one OFDM block is 45.6 ms. The data was modulated using a QPSK constellation. All the results presented below are averaged values over 100 independent Monte Carlo runs. The channel impulse response used for the simulation is depicted in figure 3. The additive noise was a complex valued white Gaussian noise.

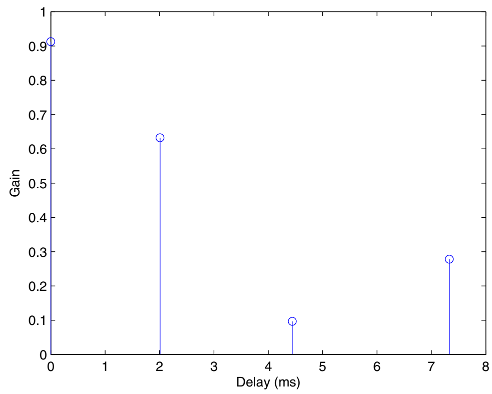

Fig. 3. Impulse response of the simulated channel

We evaluate the performance of the proposed scheme by computing the mean square error between the estimated and the actual values of the Doppler scaling factor. The data detection method is evaluated by means of the bit-error-rate (BER).

Fig. 4 depicts the mean value of the estimated Doppler scaling factor whereas the mean square error on the estimation of this factor is depicted in Fig. 5.

We can note the effectiveness of the Doppler estimation scheme. However, the performance is degraded when decreasing the SNR. The degradation is more accentuated when increasing the number of subcarriers in the first sub-block of the OFDM symbol. Indeed, by increasing K_u, the frequency spacing of the subcarriers is reduced. Therefore, the harmonic retrieval problem needs higher resolution and precision. Such a precision is difficult to obtain when the SNR decreases. Increasing K_u seems to be relevant only for high SNR values. This observation is directly related to the behaviour of HTLS in noisy cases.

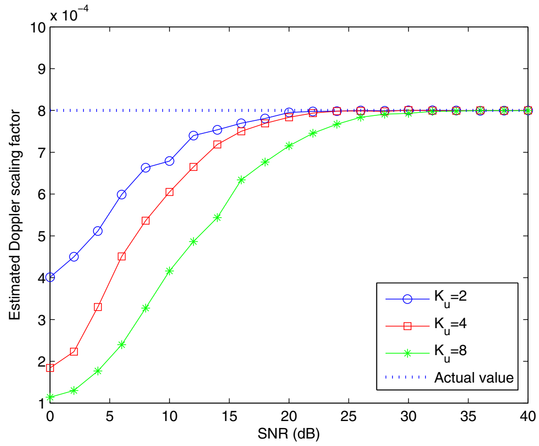

Fig. 4. Estimated value of the Doppler scaling factor.

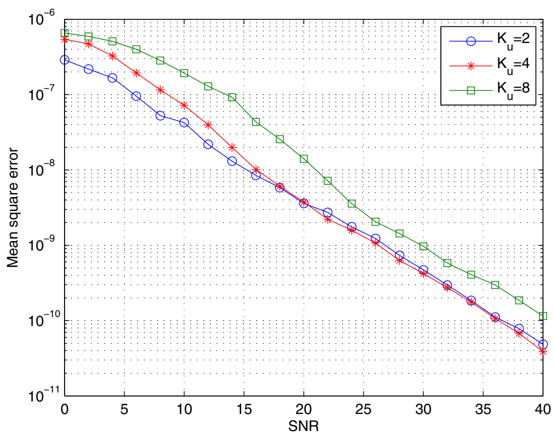

Fig. 5. Mean Square Error for the estimation of the Doppler scaling factor.

In Fig. 6, we depict the BER for three numbers K_p of pilots and two values for K_u. Increasing the number of pilot allows improving the BER.

From these simulations, we can conclude that increasing the number of pilots has a positive effect on the performance of the receiver while increasing the number of subcarriers K_u has a contrary effect.

VI. Conclusion

In this paper, we have presented a new scheme for both estimating the Doppler scaling factor and detecting the informative data. In the proposed scheme, each OFDM symbol is divided in two parts. The first one is shorter and carries out few data with few subcarriers whereas the second one contains both informative and pilot symbols. The first sub-block serves to estimate the Doppler scaling factor while the pilots in the second sub-block are used for estimating the equivalent channel. Both estimation methods are based on high resolution methods for solving harmonic retrieval problems in time and frequency domains respectively. The proposed scheme is particularly efficient for moderate to high SNR. Robustness to noise, for lower levels of SNR, and extension to the multi-user case will be investigated. In addition, effectiveness of the proposed approach using experimental data is to be considered in future works.

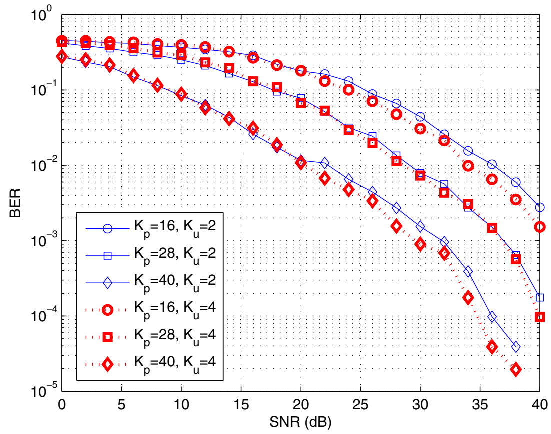

Fig. 6. BER for different number of pilots and two numbers of subcarriers in the first sub-block ($K_d = 256$)

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