Near LMMSE channel estimation performance with artificial channel at receiver for OFDM systems

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NEAR LMMSE CHANNEL ESTIMATION PERFORMANCE WITH ARTIFICIAL CHANNEL AT RECEIVER FOR OFDM SYSTEMS

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ABSTRACT

This paper proposes a LMMSE-based channel estimator which, unlike the classical LMMSE estimator, does not require the covariance matrix of the channel nor its estimation. Actually, we add at the receiver side a fully adjustable filter, which acts like an artificial channel and hides the physical channel. We then perform an LMMSE estimation of the sum of the physical and artificial channel using the filter covariance matrix and the channel estimation is obtained by subtracting the filter. Theoretical developments, shown in this paper, prove that the performance of the proposed solution can be driven by the parameters of this additional filter and can reach the one of the theoretical LMMSE estimator closely. Simulations, proposed in a DRM context, also display the validity of this technique for both preamble-based and scattered pilot distributions. It is shown that the proposed solution is only about 1 dB from the LMMSE technique in term of Minimum Mean Square Error (MMSE) and Bit Error Rate (BER).

1. INTRODUCTION

The Orthogonal Frequency Division Multiplexing (OFDM) modulation is nowadays known as a robust and powerful solution for wireless or wired transmissions and is now normalized in many standards of telecommunication like DVB-T, LTE, xDSL technologies, etc. This attractiveness mainly comes from the modulation robustness against the frequency selectivity due to the multipath channel. Indeed, by a simple addition of a guard interval (GI), we limit intersymbol interference in an efficient way. Similarly, the decomposition of the OFDM signal over several narrow frequency bands is an appropriate technique to fight against frequency selectivity and also allows an only one-tap per carrier equalization.

However, before equalization, an estimation of the channel coefficients is necessary. In many standards, some positions in the time-frequency lattice are reserved for this estimation and contain no information data. These positions are termed as pilot tones. The repartition and the number of these tones in the time-frequency lattice depends on the nature of the channel and its severity.

Among the wide range of possible estimation techniques (see references [1, 2, 3] for a short description), the LS (Least Square) solution offers an acceptable level of performance with a low complexity as one estimates the channel coefficient on a pilot tone by computing the ratio between the received data and the pilot data value. After that, an interpolation is performed to provide the entire channel at each instant and frequency. However, this technique is very sensitive to transmission noise and the interpolation that is considered.

Another well known estimation technique is the 2D Wiener filtering, described in [4], obtained thanks to the minimization of the MSE (Mean Square Error) criterion. In the case of separated pilots in an OFDM frame, reference [5] proves that we obtain the same level of performance as the 2D Wiener filtering by interpolating the channel coefficients, estimated with the LS criterion, with the LMMSE algorithm (Linear Minimum Mean Square Error). This performance is also very close to the perfect estimation bound. As the LMMSE estimation needs a channel covariance matrix to be performed [6], its main drawback lies in the necessity to know some channel parameters like path delays or gains. Unfortunately, these parameters (or their statistics) are a priori unknown and need to be estimated, as proposed in [7]. However, this covariance matrix estimation must be regularly updated to ensure an acceptable level of performance.

We here propose a LMMSE-based channel estimation technique which does not necessitate the channel covariance matrix nor its estimation. The basic idea of our technique is to inject pilots through a filter at the receiver side whose parameters can be fully set up by the user. Actually, this filter plays the role of an artificial channel. From the estimator block point of view, the resulting signal is the sum of pilots signal through this artificial channel and the OFDM signal through the physical channel. This hybrid channel (composed by the sum of the physical and artificial channels) can be estimated by a LMMSE estimator. Its performances can be driven by an appropriate choice of filter parameters. We get an estimation of the physical channel by subtracting the filter from the estimated hybrid channel. Using a suitable filter, our method is independent from the variations of the statistics of the channel. The covariance matrix is consequently computed only once.

In order to describe this solution in a better way, we first detail in Section 2 our system model. Then, in Section 3, the basic LS and LMMSE solutions are presented, leading to the description of our solution in Section 4. From the expression of the hybrid channel covariance matrix given in Section 5, we propose a set of parameters for the artificial channel allowing this estimation. Section 6 then proves the validity and shows the performance of our proposed solution in a DRM context, in preamble-based and scattered pilot distribution contexts.

2. BASEBAND SYSTEM MODEL

In this paper, we consider the transmission of an OFDM signal over a time-varying multipath fading channel according to the model describe in [8]. This channel, denoted $h(t)$, is composed of a sum of L different and independent paths:

$$h(t) = \sum_{l=0}^{L-1} h_l \delta(t - \tau_l), \tag{1}$$

where the variable h_l follows a zero-mean complex Gaussian process with a power-delay profile $\Theta(\tau_l)$ and the delay τ_l is distributed with a probability density function $\Pi(\tau_l)$ [5]. The OFDM system also includes a cyclic prefix (CP) which guarantees a transmission with a good robustness against the multipath effects of the channel. For discrete time OFDM systems we are considering in the rest of the paper, we note N the number of carriers and L_{CP} the length of the CP.

In the following, matrices are depicted in underlined and bold font ($\underline{\mathbf{M}}$), vectors in bold font ($\mathbf{V}$) and scalars in normal font. We also note $\mathbb{C}^{m\times n}$ the set of complex-valued matrices composed of m rows and n columns. Superscripts ^{T} and ^{H} respectively assign the matrix transpose and the Hermitian transpose. In the discrete time domain, the channel impulse response is represented by an element of $\mathbb{C}^{N\times 1}$, i.e. a complex vector of length N. Considering the n^{th} transmitted OFDM symbol, this impulse response $\mathbf{h}_n$ (resp. this frequential response $\mathbf{H}_n$) is formed by the coefficients $h_{l,n}$, $l \in \{0, \ldots, N-1\}$ in the time domain (resp. $H_{l,n}$ in the frequency domain). The coefficient $h_{l,n}$ is obtained by sampling $h(t)$ at the frequency sampling $f_e = N f_0$, with f_0 the intercarrier frequency of the OFDM system.

At the receiver side, after the removal of the CP, the n^{th} received symbol in the frequency domain $\mathbf{U}_n = (U_{0,n}, \ldots, U_{N-1,n})^T$ is such that:

$$\mathbf{U}_n = \underline{\mathbf{H}}_n \mathbf{C}_n + \mathbf{W}_n, \tag{2}$$

where $\underline{\mathbf{H}}_n$ and $\mathbf{W}_n = (W_{0,n}, \ldots, W_{N-1,n})^T$ are respectively the channel matrix and the noise in the frequency domain. In addition, the vector $\mathbf{C}_n = (C_{0,n}, \ldots, C_{N-1,n})^T$ is composed of P pilot tones and $N - P$ elements of a given constellation like BPSK or M-QAM. Reference [9] also ensures that $\underline{\mathbf{H}}_n$ is a $\mathbb{C}^{N\times N}$ diagonal matrix, formed with the channel coefficients $H_{m,n}$. As $\underline{\mathbf{H}}_n$ is diagonal matrix, we can rewrite (2) as:

$$\mathbf{U}_n = \underline{\mathbf{C}}_n \mathbf{H}_n + \mathbf{W}_n, \tag{3}$$

where $\mathbf{H}_n = (H_{0,n}, \ldots, H_{N-1,n})^T$ and $\underline{\mathbf{C}}_n$ is a $N \times N$ diagonal matrix whose diagonal elements are $C_{0,n}, \ldots C_{N-1,n}$.

3. CHANNEL ESTIMATION

In this section, we first recall the algorithm description and the performance of the LMMSE estimation techniques, which we consider as reference solutions. We also suppose that a frame of M OFDM symbols is transmitted. A preamble repartition is considered, denoting $n_0 \in \{0, \ldots, M-1\}$ its position in the frame.

The LMMSE estimation [5, 6] is a very efficient way to estimate a response channel all over the time-frequency lattice. After an estimation of the channel coefficients on the pilot tones with an LS estimation, we exploit the statistical properties of the channel to build this LMMSE estimator. Considering a preamble-based scheme, we may derive the expression of the LMMSE estimation [5, 6]:

$$\hat{\mathbf{H}}_{n_0}^{LMMSE} = \underline{\mathbf{R}}_H (\underline{\mathbf{R}}_H + \sigma^2 (\underline{\mathbf{C}}_{n_0} \underline{\mathbf{C}}_{n_0}^H)^{-1})^{-1} \hat{\mathbf{H}}_{n_0}^{LS}, \tag{4}$$

where σ^2 is the supposed known noise power and $\underline{\mathbf{R}}_H \in \mathbb{C}^{N\times N}$ is the frequency covariance matrix of the channel, defined by $\underline{\mathbf{R}}_H = E\{\mathbf{H}_{n_0} \mathbf{H}_{n_0}^H\}$. $\hat{\mathbf{H}}_{n_0}^{LS}$ is the Least Square estimation of the channel, i.e. $\hat{\mathbf{H}}_{n_0}^{LS} = \left(\frac{U_{0,n_0}}{C_{0,n_0}}, \frac{U_{1,n_0}}{C_{1,n_0}}, \ldots, \frac{U_{N-1,n_0}}{C_{N-1,n_0}}\right)$. From references [10, 11], we get the following expression of the minimum of MMSE:

$$MMSE_{LMMSE} = \frac{1}{\mathcal{P}/\sigma^2 + tr(\underline{\mathbf{R}}_H^{-1})}. \tag{5}$$

As $\underline{\mathbf{R}}_H$ is an Hermitian, definite and positive matrix [2], its eigenvalues are all strictly positive and it ensures that $tr(\underline{\mathbf{R}}_H^{-1}) > 0$. As for LS (see reference [10]) : $MMSE_{LS} = \frac{1}{\mathcal{P}/\sigma^2}$, we easily retrieve that, for all $\mathcal{P}/\sigma^2$ ratios, $MMSE_{LS}$ is greater than $MMSE_{LMMSE}$.

4. LMMSE ESTIMATION WITHOUT CHANNEL COVARIANCE MATRIX $\mathbf{R}_H$

4.1. How to estimate H without computing $\underline{\mathbf{R}}_H$?

The LMMSE estimator is known as an excellent estimator of the propagation channel but necessitates the calculation (or in practice the estimation) of the covariance matrix $\underline{\mathbf{R}}_H$. Our approach is different as here, we propose an estimation technique having an efficiency close to the LMMSE one, without computation of $\underline{\mathbf{R}}_H$. To do so, we add an artificial signal composed by pilots (placed at the same positions as at the emitter side) distorted by a filter G to the received signal. The illustration of this principle is given in Fig. 1. We assume that the pilot-signal is perfectly synchronized with the received signal. We also suppose that the time-varying coefficients of this filter G are randomly chosen according to a statistic that is perfectly known by the receiver. Due to the random nature of the filter coefficients, the filter G somehow acts like an artificial channel, but its statistical evolution is perfectly controlled through its statistics. Consequently, we use a channel terminology to describe G.

The basic idea of our proposed solution lies on the estimation of $K = G + H$ with LMMSE. In the following, K is called "hybrid channel" as it is the result of the addition of the physical channel H and the artificial channel G. As we do not also have any knowledge of the evolution of the propagation channel H, we propose a way to design G so that the covariance matrix of K can be fully computed thanks to the statistics of G only (or, so that $\underline{\mathbf{R}}_K \approx \underline{\mathbf{R}}_G$). We can easily retrieve an estimation of H by subtracting G from the estimation of K, without any a priori information on H.

As our proposed solution is obtained with the help of an artificial channel G, this channel estimation technique is referred in the rest of the paper as the artificial channel-aided LMMSE (ACA-LMMSE) estimation.

4.2. ACA-LMMSE estimation

In this section, we consider the same assumptions and notations as in Section 3. If we note $\mathbf{S}_{n_0}$ the set of data that are obtained after

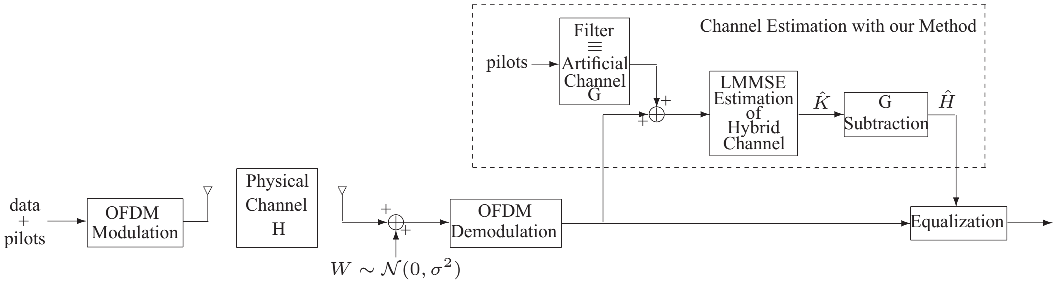

Fig. 1. Schema of the proposed solution

filtering by G and OFDM demodulation on the preamble position n_0, we get:

$$\begin{aligned} \mathbf{S}_{n_0} &= (\mathbf{H}_{n_0} + \mathbf{G}_{n_0})\, \underline{\mathbf{C}}_{n_0} + \mathbf{W}_{n_0} \\ &= \mathbf{K}_{n_0}\underline{\mathbf{C}}_{n_0} + \mathbf{W}_{n_0}. \end{aligned} \tag{6}$$

At this step in the transmission chain, a LMMSE estimation of the hybrid channel is performed. From (4), we obtain :

$$\hat{\mathbf{K}}_{n_0}^{LMMSE} = \underline{\mathbf{R}}_K(\underline{\mathbf{R}}_K + \sigma^2(\underline{\mathbf{C}}_{n_0}\underline{\mathbf{C}}_{n_0}^H)^{-1})^{-1}\hat{\mathbf{K}}_{n_0}^{LS}, \tag{7}$$

where $\underline{\mathbf{R}}_K$ is the $\mathbb{C}^{N\times N}$ covariance matrix of K. $\hat{\mathbf{K}}_{n_0}^{LS}$ contains the LS estimation of the hybrid channel coefficients on the preamble. This LS estimation can be made as it is performed on the preamble position n_0.

In order to get the LMMSE estimation of K in (7), the covariance matrix $\underline{\mathbf{R}}_K = \underline{\mathbf{R}}_{H+G}$ has to be computed. As the statistics of H (and consequently, those of K) are unknown, our solution aims at setting the statistics of G (recalling that these statistics are fixed by the user) so that $\underline{\mathbf{R}}_K \approx \underline{\mathbf{R}}_G$ whatever the variations of H. Furthermore, justifying $\underline{\mathbf{R}}_K \approx \underline{\mathbf{R}}_G$ implies that the matrix $\underline{\mathbf{R}}_G$ needs to be compute only once. By satisfying this condition, we ensure that the LMMSE of K can be performed. Section 5 explains how to correctly choose the statistics of G for a feasible LMMSE estimation of K. Finally, we retrieve an estimation of H by:

$$\hat{\mathbf{H}}_{n_0} = \hat{\mathbf{K}}_{n_0}^{LMMSE} - \mathbf{G}_{n_0}. \tag{8}$$

The coefficients of G can also be stored in a memory after their generation so that $\mathbf{G}_{n_0}$ can be fully accessible for the estimation step in Eq. (8).

4.3. MMSE of the ACA-LMMSE estimator

To compare the performance of the ACA-LMMSE estimator to the one of the LMMSE estimator, we derive an expression of its MMSE as follows:

$$\begin{aligned} J_{ACA} &= E\{||\mathbf{H}_{n_0} - \hat{\mathbf{H}}_{n_0}^{ACA}||_F^2\} \\ &= tr\left(E\left\{\left[\mathbf{H}_{n_0} - (\hat{\mathbf{K}}_{n_0}^{LMMSE} - \mathbf{G}_{n_0})\right]\right.\right. \\ &\quad \times \left.\left.\left[\mathbf{H}_{n_0} - (\hat{\mathbf{K}}_{n_0}^{LMMSE} - \mathbf{G}_{n_0})\right]^H\right\}\right) \\ &= tr\left(E\left\{\left[(\mathbf{H}_{n_0} + \mathbf{G}_{n_0}) - \hat{\mathbf{K}}_{n_0}^{LMMSE}\right]\right.\right. \\ &\quad \times \left.\left.\left[(\mathbf{H}_{n_0} + \mathbf{G}_{n_0}) - \hat{\mathbf{K}}_{n_0}^{LMMSE}\right]^H\right\}\right) \end{aligned} \tag{9}$$

As, from (6), $\mathbf{K}_{n_0} = \mathbf{H}_{n_0} + \mathbf{G}_{n_0}$, we get:

$$\begin{aligned} J_{ACA} &= tr\left(E\left\{\left[\mathbf{K}_{n_0} - \hat{\mathbf{K}}_{n_0}^{LMMSE}\right]\right.\right. \\ &\quad \times \left.\left.\left[\mathbf{K}_{n_0} - \hat{\mathbf{K}}_{n_0}^{LMMSE}\right]^H\right\}\right) \\ &= E\left\{\left\|\mathbf{K}_{n_0} - \hat{\mathbf{K}}_{n_0}^{LMMSE}\right\|_F^2\right\}. \end{aligned} \tag{10}$$

(10) then proves that the MSE of the proposed estimation technique is exactly the same as the MSE of the LMMSE estimator of the hybrid channel $K = H + G$. Consequently, from (5), we can derive the expression of the minimal MSE of this estimator:

$$MMSE_{ACA} = \frac{1}{\mathcal{P}/\sigma^2 + tr(\underline{\mathbf{R}}_K^{-1})}. \tag{11}$$

Recalling that we determine G so that $\underline{\mathbf{R}}_K \approx \underline{\mathbf{R}}_G$, (11) then proves that the performance of the ACA-LMMSE estimator can be controllable by an appropriate choice of G.

As LMMSE can be also used in a scattered pilot repartition [5], our technique is then also valid for that case. It then acts like an interpolation of the channel coefficients over the time and frequency dimensions but it is not possible to derive an analytical expression of the MMSE.

5. CHOICE OF THE ARTIFICIAL CHANNEL PARAMETERS

The goal of this section is, in a first time, to express the coefficients of $\underline{\mathbf{R}}_K$. From this, we put the light on the parameters that may impact the MSE value and that allows the condition: $\underline{\mathbf{R}}_K \approx \underline{\mathbf{R}}_G$.

5.1. Expression of the channel covariance matrix

In our mathematical developments, we suppose that the artificial channel also follows the model in (1). Then, by noting D the length of the impulse response of G, we get:

$$\begin{aligned} k(t) = (h+g)(t) &= \sum_{l=0}^{L-1} h_l\delta(t-\tau_l) + \sum_{d=0}^{D-1} g_d\delta(t-\tau_d) \\ &= \sum_{b=0}^{B-1} \gamma_b\delta(t-\tau_b), \end{aligned} \tag{12}$$

where $B \le L + D$ is the number of paths of $k(t)$ and $\gamma_b, b \in \{0, 1, ..., B-1\}$ the gain of each path in $k(t)$. Thus, for a given path delay τ_b, γ_b may be equal to h_b (respectively g_b) if $g(\tau_b) = 0$ (respectively $h(\tau_b) = 0$) or equal to $h_b + g_b$ if $g(t)$ and $h(t)$ have a common path at $t = \tau_b$. Recalling that τ_b is a random variable, we suppose that $\tau_b = \beta_b \tau_0$ with τ_0 being the sampling time and β_b a random real value which is distributed according to the multipath intensity profile $\Gamma(\beta_b)$ on the interval $[0, \beta_{max}]$. We then apply a N-points DFT (Discrete Fourier Transform) on (12), so for all $m \in \{0, \ldots, N-1\}$, we obtain the following expression of the frequency response of the hybrid channel at the frequencies $f_m = m/(N\tau_0)$:

$$K_m = K(f_m) = \sum_{b=0}^{B-1} \gamma_b \exp(-2j\pi \frac{m}{N} \beta_b). \tag{13}$$

By denoting $(\underline{\mathbf{R}}_K)_{u,v} = E\{K_u K_v^*\}$ the general term of the matrix $\underline{\mathbf{R}}_K$, we get from [5]:

$$(\underline{\mathbf{R}}_K)_{u,v} = \sum_{b=0}^{B-1} \int_0^{\beta_{max}} \Gamma(\beta_b) e^{-2j\pi \frac{(u-v)}{N} \beta_b} d\beta_b, \tag{14}$$

The delay $\tau_{max} = \beta_{max}\tau_0$ represents the maximal delay of the hybrid channel K. Thus, we respectively define τ_{maxG} and τ_{maxH} as the maximal delay of the artificial channel G and the physical channel H. Consequently, we necessarily get $\tau_{max} = \max(\tau_{maxG}, \tau_{maxH})$.

5.2. Filter G features

The analysis of the filter G leading to the approximation of the hybrid channel covariance matrix by the one of the artificial channel (i.e. $\underline{\mathbf{R}}_K \approx \underline{\mathbf{R}}_G$) being too long to be developed in this paper, we only give here the results. From Eq. (14), terms in the channel covariance matrix $\underline{\mathbf{R}}_K$ may vary according to three parameters:

- the number of paths in the hybrid channel B,
- the artificial channel maximal delay τ_{max},
- the multipath intensity profile $\Gamma(\beta)$.

Recalling that we want to obtain $\underline{\mathbf{R}}_K \approx \underline{\mathbf{R}}_G$, we notice that the first two parameters in the aforementioned list are easily driven by G. Indeed, as $B \leq L + D$, choosing D as large as possible is sufficient to guarantee that $B \approx D$. As $\tau_{max} = \max(\tau_{maxG}, \tau_{maxH})$, we ensure that $\tau_{max} = \tau_{maxG}$ by choosing $\tau_{maxG} = T_{GI}$, with T_{GI} the guard interval time duration. As a matter of fact, T_{GI} is chosen in standards using an OFDM modulation so that $T_{GI} > \tau_{maxH}$ to avoid intersymbol interferences. Finally, in (14), the multipath intensity profile is the one of K. From the previous analysis, as G has a very large number of paths and has a longer impulse response, here, we assume that the multipath intensity profile of G is somehow prevailing on the multipath intensity profile of H. Consequently, we make the assumption that, under theses conditions, the intensity profile of K is close to the one of G. Finally, by fixing an intensity profile for G, $D >> L$ and $\tau_{maxG} = T_{GI}$, we get: $\underline{\mathbf{R}}_K \approx \underline{\mathbf{R}}_G$.

6. SIMULATIONS

6.1. The DRM standard

Our simulations are based on the DRM standard [12], that proposes a set of digital audio broadcasting technologies designed to work over the bands currently used for AM broadcasting. The transmitted signal is modulated by an OFDM modulation, with different numbers of carriers and different constellation sizes according to the severity of the propagation channel. In the following, we consider the transmission mode C, i.e. the OFDM system counts 148 independent carriers, each carrier modulating a data issued from a 64-QAM constellation. A guard interval (GI) of time duration $T_{GI} = 5.33$ ms is added to each OFDM symbol. The sampling frequency of the considered system is 10 kHz. Finally, we generate DRM frames which is composed of 20 OFDM symbols (including the GI) whose time duration is then 400 ms. The DRM channel profiles are generated according to the description given in [12, Annex B]. Each channel is characterized by its number of paths and paths gains. For each path, the Doppler shift and Doppler spread is also provided. In all our simulations, we use the *US Consortium* channel with $L = 4$ and $\tau_{maxH} = 2.2$ ms. The delay profile follows the one given by the standard.

6.2. Analysis of the ACA-LMMSE performance

This paragraph aims at validating our proposed solution and the requirements given in subsection 5.2. We consider two different configurations for G. The first one (labelled as config. 1 in the following) does not satisfy our requirements: we set $D = 3$, $\tau_{maxG} = 1.18$ ms and the delay profile follows a decreasing and exponential profile. The second one (labelled as config. 2) has the following characteristics: $D = 15$, $\tau_{maxG} = T_{GI}$ and the delay profile also follows a decreasing and exponential profile. Config. 2 is then coherent regarding the requirements. Note that, unlike τ_{maxG} and the delay profile, we here set D knowing the number of paths of H given by the DRM standard. In other contexts where L is totally unknown, choosing D excessively large is then sufficient.

Fig. 2 compares the performance of the proposed method (ACA-LMMSE) with LS and theoretical LMMSE in term of MMSE, computed by $E\{\frac{1}{N}\sum_{m=0}^{N-1} |H_m - \hat{H}_m|^2\}$, according to two pilots repartitions : a preamble-based (PB), as our developments were obtained assuming this repartition scheme, and a scattered (SP) pilot repartition, according to the DRM standard . Concerning the SP case for LS, a polynomial interpolation over the time and frequency dimensions is made as it efficiently limits the degradation of the MSE.

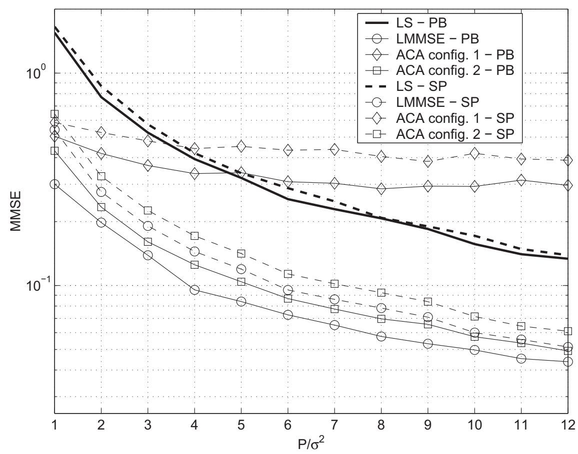

Fig. 2. Evolution of MMSE of ACA-LMMSE compared to LS and LMMSE as function of $\frac{\mathcal{P}}{\sigma^2}$ for two different pilot distributions (SP and PB).

Considering in the first time the PB case, we observe that an appropriate choice of the characteristics of G has a great impact on the MMSE performance. Besides, if we follow the requirements given in subsection 5.2 (config. 2), the performance of the estimation technique is clearly improved. This remark then validates the analysis about the parameters choice in subsection 5.2. Now considering ACA-LMMSE, it clearly outperforms LS as it offers an almost 10 dB MMSE gain. The ACA-LMMSE performance is slightly degraded compared to LMMSE, as our estimation uses the covariance matrix $\underline{\mathbf{R}}_K$ which is naturally different from $\underline{\mathbf{R}}_H$. However, this loss is very limited (only 1 dB from LMMSE bound). Our work is now based on the parameters characterization of G to obtain MMSE values that may reach the optimal performance.

For the SP repartition, we then notice that we exactly have the same MMSE evolutions, then proving that our technique is also efficient in this case. Indeed, as in this context, the MMSE depends on the nature of the interpolation technique, no theoretical expression of the MMSE can be found in the literature. Note then that the differences between the SP and PB cases are issued from the interpolation technique that inevitably degrades the estimation quality.

Following the DRM standard, the BER of ACA-LMMSE, LMMSE and LS estimations are computed in the SP case (Fig. 3), with the same parameters of parameters. Concerning the LS estimation, a polynomial interpolation in the frequency dimension is made (pointed out "poly." in the legend). Whereas the LS estimation reaches an error floor at $BER = 6.10^{-3}$, our method has performance close to the theoretical LMMSE estimation. We then observe only 1 dB loss when compared to the LMMSE estimation and just 2 dB loss when compared to the perfect estimation. These results confirm the ACA-LMMSE efficiency in channel estimation. In addition, it requires no direct computation of $\underline{\mathbf{R}}_H$ (as for LMMSE solution) and no estimation of this matrix.

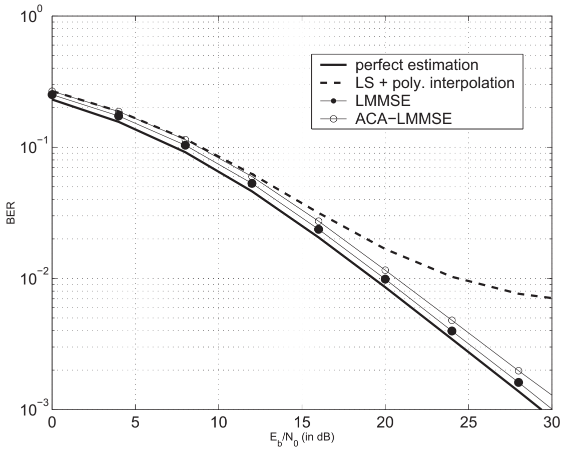

Fig. 3. BER of the proposed method comparing with LMMSE and LS.

7. CONCLUSION

This paper proposes a channel estimation technique, called ACA-LMMSE, which allows an efficient estimation of a physical channel without any knowledge of its covariance matrix nor the need to compute its estimation. This solution is based on the addition at the receiver side of a fully tunable filter that plays the role of an artificial channel and that permits an efficient estimation of the physical channel by an appropriate choice of parameters. We have also shown that it can be used in a preamble-based or a scattered pilot repartition scheme. As the performance of this estimator can be headed by the artificial channel parameters, further works are led concerning the optimization of the parameters set. It aims at justifying at best the approximation of the hybrid channel covariance matrix by the one of the only articial channel.

8. REFERENCES

[1] M. Morelli and U. Mengali, "A Comparisaon of Pilot-Aided Channel Estimation Methods for OFDM Systems," *IEEE Transactions on Signal Processing*, vol. 49, no. 12, December 2001.

[2] M. Biguesh and A. B. Gershman, "Training-Based MIMO Channel Estimation : A Study of Estimator Tradeoffs and Optimal Training Signals," *IEEE transactions on signal processing*, vol. 54, no. 3, March 2006.

[3] X. Dong, W.-S. Lu, and A.C.K. Soong, "Linear Interpolation in Pilot Symbol Assisted Channel Estimation for OFDM," *IEEE Trans. on wireless Communications*, vol. 6, no. 5, pp. 1910 – 1920, May 2007.

[4] P. Hoeher, S. Kaiser, and P. Robertson, "Two-dimensional pilot-symbol-aided channel estimation by Wiener filtering," in *ICASSP*, April 1997.

[5] O. Edfords, M. Sandell, J.-J. van de Beek, S. K. Wilson, and P. O. Brjesson, "OFDM Channel Estimation by Singular Value Decomposition," *IEEE Trans. on Communications*, vol. 46, no. 7, July 1998.

[6] J. J. van de Beek, O. Edfors, M. Sandell, S. K. Wilson, and P. O. Brjesson, "On Channel Estimation in OFDM Systems," in *IEEE conference on vehicular technology*, September 1995, vol. 2, pp. 815 – 819.

[7] W. Zhou and W. H. Lam, "A fast LMMSE channel estimation method for OFDM systems," *EURASIP Journal on Wireless Communications and Networking*, March 2009.

[8] M. Patzold, *"Mobile fading channels"*, John Wiley & Sons, 1999.

[9] R. M. Gray, "Toeplitz and circulant matrices : A review," Tech. Rep., Department of Electrical Engineering, Standford University, 2006.

[10] M. Biguesh and A. B. Gershman, "Downlink Channel Estimation in Cellular Systems with Antenna Arrays at Base Stations Using Channel Probing with Feedback," *EURASIP Journal on Applied Signal Processing*, pp. 1330–1339, September 2004.

[11] H. Lütkepohl, *Handbook of matrices*, John Wiley & Sons, 1996.

[12] ETSI, "Digital Radio Mondiale (DRM);System Specification," Tech. Rep. ETSI EN 300 744 V1.5.1, ETSI, August 2009.