Pseudo Random Postfix OFDM based channel estimation in a Doppler scenario

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***Abstract*— This contribution[1] proposes novel and low complexity channel estimation and tracking architectures in the context of the recently proposed Pseudo Random Postfix OFDM (PRP-OFDM). In a first time, the channel estimation is performed for a static environment exploiting order-one statistics of the received signal. Then, the results are extented to a Doppler scenerio. An MMSE estimator is proposed that avoids commonly used approximations of Jake's Doppler model (such as the order-one autoregressive approach). Replacing the standard Cyclic-Prefix OFDM (CP-OFDM) modulator in IEEE802.11a or BRAN HIPERLAN/2 by a PRP-OFDM modulator is shown to improve the mobility from 3m/s (pedestrian speed) to 36m/s for QPSK and to 72m/s for BPSK constellations.**

I. Introduction

Nowadays, Orthogonal Frequency Division Multiplexing (OFDM) seems the preferred modulation scheme for modern broadband communication systems. Indeed, the OFDM inherent robustness to multi-path propagation and its appealing low complexity equalization receiver makes it suitable either for high speed modems over twisted pair (digital subscriber lines xDSL), terrestrial digital broadcasting (Digital Audio and Video Broadcasting: DAB, DVB) and 5GHz Wireless Local Area Networks (WLAN: IEEE802.11a and ETSI BRAN HIPERLAN/2) [1]–[4].

All these systems are based on a traditional Cyclic Prefix OFDM (CP-OFDM) modulation scheme. The role of the cyclic prefix is to turn the linear convolution into a set of parallel attenuations in the discrete frequency domain. Recent contributions have proposed an alternative: replacing this time domain redundancy by null samples leads to the so called Zero Padded OFDM (ZP-OFDM) [5]–[8]. This solution, relying on a larger FFT demodulator, has the merit to guarantee symbol recovery irrespective of channel null locations in absence of noise when the channel is known (coherent modulations are assumed).

Channel coefficients estimation is usually performed using known training sequences periodically transmitted (e.g. at the start of each frame), implicitly assuming that the channel does not vary between two training sequences. Thus in order to enhance the mobility of wireless systems and cope with the Doppler effects, reference sequences have to be repeated more often, resulting in a significant loss of useful bitrate. An alternative solution is to track the channel variations by refining the channel coefficients blindly using the training sequences as initializations for the estimator.

Semi-blind equalization algorithms based on second order statistics have already been proposed for the CP-OFDM and ZP-OFDM modulators [7]–[9]. However, their inherent computational complexity is quite important.

These drawbacks motivated the recent proposal of the Pseudo Random Postfix OFDM (PRP-OFDM) modulation [10]–[12] that capitalizes on the advantages of ZP-OFDM. The null samples of ZP-OFDM inserted between all OFDM modulated blocks are replaced by a known vector weighted by a pseudo random scalar sequence. This way, unlike the former OFDM modulators, the receiver can exploit an additional information: the prior knowledge of a part of the transmitted block. This paper explains how to build on this knowledge and proposes a very low complexity order one semi-blind channel estimation and tracking algorithm, very efficient in static and Doppler contexts. The decoding procedure for PRP-OFDM symbols is not addressed here, but various approaches with different complexity/performance trade-offs are available in [10], [11].

This paper is organized as follows. Section II introduces the notations and presents the new PRP-OFDM modulator. Then a blind channel estimation method is presented in section III for the static context. Section IV proposes a new Doppler model and extends the CIR estimation results to the Doppler scenario. A corresponding optimum estimator in the MMSE sense is proposed. Finally, simulation results in the context of 5GHz IEEE802.11a and ETSI BRAN HIPERLAN/2 illustrate the behavior of the proposed scheme compared to the standardized CP-OFDM systems in section V.

II. Notations and PRP-OFDM modulator

Figure 1 depicts the baseband discrete-time block equivalent model of an N carrier PRP-OFDM system. The ith $N \times 1$ input digital vector [2] $\tilde{\mathbf{s}}_N(i)$ is first modulated by the IFFT matrix $\mathbf{F}_N^H = \frac{1}{\sqrt{N}}\left(W_N^{ij}\right)^H, 0 \le i < N, 0 \le j < N$ and $W_N = e^{-j\frac{2\pi}{N}}$. Then, a deterministic postfix vector $\mathbf{c}_D = (c_0, \ldots, c_{D-1})^T$ weighted by a pseudo random value $\alpha(i) \in \mathbb{C}$ is appended to the IFFT outputs $\mathbf{s}_N(i)$. A pseudo random $\alpha(i)$ prevents the postfix time domain signal from being deterministic and avoids thus

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[2]Lower (upper) boldface symbols will be used for column vectors (matrices) sometimes with subscripts N or P emphasizing their sizes (for square matrices only); tilde will denote frequency domain quantities; argument i will be used to index blocks of symbols; ^{H} (^{T}) will denote Hermitian (Transpose).

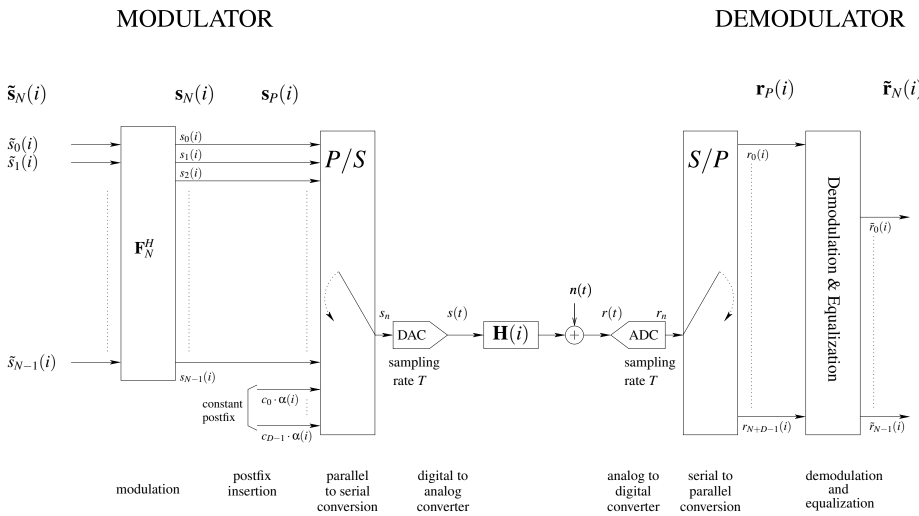

Fig. 1. Discrete model of the PRP-OFDM modulator.

spectral peaks [10]. With $P = N + D$, the corresponding $P \times 1$ transmitted vector is $\mathbf{s}_P(i) = \mathbf{F}_{\mathrm{ZP}}^H \tilde{\mathbf{s}}_N(i) + \alpha(i)\mathbf{c}_P$, where

$$\mathbf{F}_{\mathrm{ZP}}^H = \left[\frac{\mathbf{I}_N}{\mathbf{0}_{D,N}} \right]_{P\times N} \mathbf{F}_N^H \quad \text{and} \quad \mathbf{c}_P = \left(\mathbf{0}_{1,N}\, \mathbf{c}_D^T\right)^T$$

The samples of $\mathbf{s}_P(i)$ are then sent sequentially through the channel modeled here as a Lth-order FIR $H(z) = \sum_{n=0}^{L-1} h_n z^{-n}$ of impulse response $(h_0, \cdots, h_{L-1})$. The OFDM system is designed such that the postfix duration exceeds the channel memory $L \leq D$.

Let $\mathbf{H}_{\mathrm{ISI}}(P)$ and $\mathbf{H}_{\mathrm{IBI}}(P)$ be respectively the Toeplitz inferior and superior triangular matrices of first column: $[h_0, h_1, \cdots, h_{L-1}, 0, \rightarrow, 0]^T$ and first row $[0, \rightarrow, 0, h_{L-1}, \cdots, h_1]$. As already explained in [13], the channel convolution can be modeled by $\mathbf{r}_P(i) = \mathbf{H}_{\mathrm{ISI}}\mathbf{s}_P(i) + \mathbf{H}_{\mathrm{IBI}}\mathbf{s}_P(i-1) + \mathbf{n}_P(i)$. $\mathbf{H}_{\mathrm{ISI}}(P)$ and $\mathbf{H}_{\mathrm{IBI}}(P)$ represent respectively the intra and inter block interference. Since $\mathbf{s}_P(i) = \mathbf{F}_{\mathrm{ZP}}^H \tilde{\mathbf{s}}_N(i) + \alpha(i)\mathbf{c}_P$, we have as illustrated by figure 2:

$$\mathbf{r}_P(i) = (\mathbf{H}_{\mathrm{ISI}} + \beta_i \mathbf{H}_{\mathrm{IBI}})\mathbf{s}_P(i) + \mathbf{n}_P(i)$$

where $\beta_i = \frac{\alpha(i-1)}{\alpha(i)}$ and $\mathbf{n}_P(i)$ is the ith AWGN vector of element variance σ_n^2. Note that $\mathbf{H}_{\beta_i} = (\mathbf{H}_{\mathrm{ISI}} + \beta_i \mathbf{H}_{\mathrm{IBI}})$ is pseudo circulant, i.e. a circular matrix whose $(D-1)\times(D-1)$ strictly upper triangular part (without the main diagonal) is weighted by β_i.

The expression of the received block is thus:

$$\begin{aligned} \mathbf{r}_P(i) &= \mathbf{H}_{\beta_\mathrm{i}} \left(\mathbf{F}_{\mathrm{ZP}}^H \tilde{\mathbf{s}}_N(i) + \alpha(i)\mathbf{c}_P\right) + \mathbf{n}_P(i) \qquad (1) \\ &= \mathbf{H}_{\beta_\mathrm{i}} \begin{pmatrix} \mathbf{F}_N^H \tilde{\mathbf{s}}_N(i) \\ \alpha(i)\mathbf{c}_D \end{pmatrix} + \mathbf{n}_P(i) \end{aligned}$$

Please note that equation (1) is quite generic and captures also the CP and ZP modulation schemes. Indeed ZP-OFDM corresponds to $\alpha(i) = 0$ and CP-OFDM is achieved for $\alpha(i) = 0$, $\beta_i = 1 \forall i$ and $\mathbf{F}_{\mathrm{ZP}}^H$ is replaced by $\mathbf{F}_{\mathrm{CP}}^H$, where

$$\mathbf{F}_{\mathrm{CP}}^H = \left[\frac{\mathbf{0}_{D,N-D} \mid \mathbf{I}_D}{\mathbf{I}_N} \right]_{P\times N} \mathbf{F}_N^H.$$

III. An inherent order one semi-blind channel estimation

PRP-OFDM allows an order one and low-complexity channel estimation. For explanation sake assume that the Channel Impulse Response (CIR) is static.

Define $\mathbf{H}_{\mathrm{CIR}}(D) = \mathbf{H}_{\mathrm{ISI}}(D) + \mathbf{H}_{\mathrm{IBI}}(D)$ as the $D \times D$ circulant channel matrix of first row $\mathrm{r}ow_0(\mathbf{H}_D) = [h_0, 0, \rightarrow, 0, h_{L-1}, \cdots, h_1]$. Note that $\mathbf{H}_{\mathrm{ISI}}(D)$ and $\mathbf{H}_{\mathrm{IBI}}(D)$ contain respectively the lower and upper triangular parts of $\mathbf{H}_{\mathrm{CIR}}(D)$.

Denoting by $\mathbf{s}_N(i) = [s_0(i), \cdots, s_{N-1}(i)]^T$, extracting 2 parts from this vector: $\mathbf{s}_{N,0}(i) = [s_0(i), \cdots, s_{D-1}(i)]^T$, $\mathbf{s}_{N,1}(i) = [s_{N-D}(i), \cdots, s_{N-1}(i)]^T$, and performing the same operations for the noise vector: $\mathbf{n}_P(i) = [n_0(i), \cdots, n_{P-1}(i)]^T$, $\mathbf{n}_{D,0}(i) = [n_0(i), \cdots, n_{D-1}(i)]^T$, $\mathbf{n}_{D,1}(i) = [n_{P-D}(i), \cdots, n_{P-1}(i)]^T$, the received vector $\mathbf{r}_P(i)$ can be expressed as:

$$\begin{bmatrix} \mathbf{H}_{\mathrm{ISI}}(D)\mathbf{s}_{N,0}(i) + \alpha(i-1)\mathbf{H}_{\mathrm{IBI}}(D)\mathbf{c}_D + \mathbf{n}_{D,0}(i) \\ \vdots \\ \mathbf{H}_{\mathrm{IBI}}(D)\mathbf{s}_{N,1}(i) + \alpha(i)\mathbf{H}_{\mathrm{ISI}}(D)\mathbf{c}_D + \mathbf{n}_{D,1}(i) \end{bmatrix} \qquad (2)$$

As usual the transmitted time domain signal $\mathbf{s}_N(i)$ is zero-mean. Thus the first D samples $\mathbf{r}_{P,0}(i)$ of $\mathbf{r}_P(i)$ and its last D samples $\mathbf{r}_{P,1}(i)$ can be exploited very easily to find the channel matrices relying on the deterministic nature of the postfix as follows:

$$\mathbf{r}_{c,0} = \mathrm{E}\left[\frac{\mathbf{r}_{P,0}(i)}{\alpha(i-1)}\right] = \mathbf{H}_{\mathrm{IBI}}(D)\mathbf{c}_D, \qquad (3)$$

$$\mathbf{r}_{c,1} = \mathrm{E}\left[\frac{\mathbf{r}_{P,1}(i)}{\alpha(i)}\right] = \mathbf{H}_{\mathrm{ISI}}(D)\mathbf{c}_D. \qquad (4)$$

Since $\mathbf{H}_{\mathrm{ISI}}(D) + \mathbf{H}_{\mathrm{IBI}}(D) = \mathbf{H}_{\mathrm{CIR}}(D)$ is circular and diagonal in the frequency domain combining equations (3) and (4) and

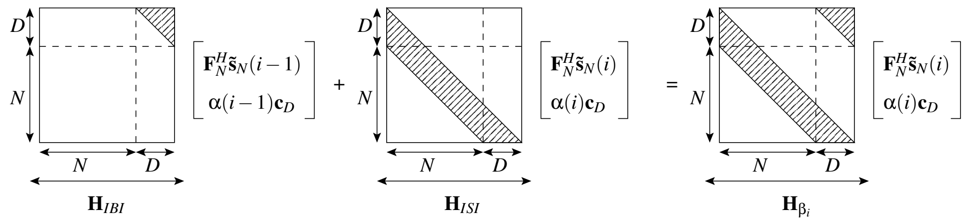

Fig. 2. Circularization for PRP-OFDM.

using the commutativity of the convolution yields:

$$\begin{aligned}\mathbf{r}_c = \mathbf{r}_{c,1} + \mathbf{r}_{c,0} &= \mathbf{H}_{\text{CIR}}(D)\mathbf{c}_D \\ &= \mathbf{C}_D\mathbf{h}_D = \mathbf{F}_D^H\tilde{\mathbf{C}}_D\mathbf{F}_D\mathbf{h}_D,\end{aligned}$$

where $\mathbf{C}_D$ is a $D \times D$ circulant matrix with first row $row_0(\mathbf{C}_D) = [c_0, c_{D-1}, c_{D-2}, \cdots, c_1]$ and $\tilde{\mathbf{C}}_D = \text{diag}\{\mathbf{F}_D\mathbf{c}_D\}$. Thus, an estimate of the CIR r_D can be retrieved as follows:

$$\hat{\mathbf{h}}_D = \mathbf{C}_D^{-1}\mathbf{r}_c = \mathbf{F}_D^H\tilde{\mathbf{C}}_D^{-1}\mathbf{F}_D\mathbf{r}_c.$$

Note that $\mathbf{c}_D$ is designed such that $\tilde{\mathbf{C}}_D$ is full rank. If the expectation operator E is approximated by mean value calculation over Z observations, an additional additive noise term $\mathbf{n}$ must be taken into account, as it will be illustrated below.

Sometimes design constraints, such as limited out-of-band radiation, impose a rank-deficient postfix matrix $\mathbf{C}_D$. In this case, it is more appropriate to keep (3) and (4) separately:

$$\begin{bmatrix} \mathbf{r}_{c,0} \\ \mathbf{r}_{c,1} \end{bmatrix} = \begin{bmatrix} \mathbf{H}_{\text{IBI}}(D) \\ \mathbf{H}_{\text{ISI}}(D) \end{bmatrix} \mathbf{c}_D + \mathbf{n} = \begin{bmatrix} \mathbf{C}_{\text{IBI}}(D) \\ \mathbf{C}_{\text{ISI}}(D) \end{bmatrix} \mathbf{h}_D + \mathbf{n}$$

$\mathbf{C}_{\text{IBI}}(D)$ and $\mathbf{C}_{\text{ISI}}(D)$ are constructed in the same way as $\mathbf{H}_{\text{IBI}}(D)$ and $\mathbf{H}_{\text{ISI}}(D)$, but based on $\mathbf{c}_D$. $\mathbf{h}_D$ is extracted by an MMSE approach (5) with $\mathbf{R}_{\mathbf{h},\mathbf{h}} = \text{E}\left[\mathbf{h}\mathbf{h}^H\right]$ and $\mathbf{R}_{\mathbf{n},\mathbf{n}} = \text{E}\left[\mathbf{n}\mathbf{n}^H\right]$. $\mathbf{n}$ is straightforwardly derived from (2) with k being the latest OFDM symbol number:

$$\mathbf{n} = \sum_{i=k-Z+1}^{k} \frac{1}{Z\alpha(i)} \begin{bmatrix} \mathbf{H}_{\text{ISI}}(D)\mathbf{s}_{N,0}(i+1) + \mathbf{n}_{D,0}(i+1) \\ \mathbf{H}_{\text{IBI}}(D)\mathbf{s}_{N,1}(i) + \mathbf{n}_{D,1}(i) \end{bmatrix}$$

We have detailed in this section a very simple method for blind estimation of the CIR only relying on first order statistics: the expectation of the received signal vector.

IV. Channel estimation in a Doppler context

In the context of a Doppler scenario, the above results must be adapted accordingly. Therefore, the choice of the Doppler model plays an essential role. Jake's commonly accepted Doppler model [14] shall be used throughout this paper (since it is applicable to many practical contexts) stating that $E\left[h_l(n)h_l^\star(n-1)\right] \approx J_0(2\pi f_D T_n)\,\text{E}\left[|h_l(n)|^2\right]$ with $h_l(n)$ being the l^{th} component of the CIR $\mathbf{h}_D(n)$ at instant $T_n = n\Delta T$, $J_0(\cdot)$ the 0^{th} order Bessel function of the first kind, f_D the Doppler frequency and ΔT being the OFDM-plus-postfix symbol duration. The channel is assumed to vary only insignificantly within one OFDM symbol including the postfix.

In the context of CIR estimation in a Doppler scenario, the tolerated *system latency* plays an important role. The following two estimation approaches can be considered: i) In order to estimate the CIR for the latest OFDM symbol, only consider previously received symbols. Thus, the *system latency* is not impacted. ii) Consider previous *and* future OFDM symbols for CIR estimation. Future symbols are available by increasing the *system latency*; received symbols are buffered over a given interval and are decoded after a corresponding delay. The latter approach leads to far better results in a high Doppler scenario, but it is not really compatible with automatic repeat request (ARQ) mechanisms as commonly found in WLANs. This is the reason why we focus on approach i) in this paper.

The following derivations provide a CIR estimation approach that is optimum in the MMSE sense. The CIR is estimated for each OFDM symbol separately based on the symbol itself plus the $Z-1$ preceding ones. The first D samples of the following OFDM symbol are taken into account as well, since it contains a contribution of the latest postfix after channel convolution. With k being the latest OFDM symbol number, the following observations are exploited for the estimation:

$$\begin{aligned}\mathbf{y}_{2DZ}(k) &= \left[\mathbf{y}_{2D}^T(k), \mathbf{y}_{2D}^T(k-1), \cdots, \mathbf{y}_{2D}^T(k-Z+1)\right]^T, \\ \mathbf{y}_{2D}(k) &= \begin{bmatrix} \mathbf{H}_{\text{IBI}}^D(k) \\ \mathbf{H}_{\text{ISI}}^D(k) \end{bmatrix} \mathbf{c}_D + \alpha^{-1}(k)\mathbf{n}(k), \\ \mathbf{n}(k) &= \begin{bmatrix} \mathbf{H}_{\text{ISI}}^D(k+1)\mathbf{s}_{N,0}(k+1) + \mathbf{n}_{D,0}(k+1) \\ \mathbf{H}_{\text{IBI}}^D(k)\mathbf{s}_{N,1}(k) + \mathbf{n}_{D,1}(k) \end{bmatrix}\end{aligned}$$

$\mathbf{H}_{\text{IBI}}^D(k)$ and $\mathbf{H}_{\text{ISI}}^D(k)$ are the $D \times D$ time-varying channel matrices based on $\mathbf{h}_D(k)$. With these observation, an estimation matrix $\mathbf{W}(k)$ of dimension $D \times 2DZ$ must be derived meeting the following MMSE criteria:

$$\mathbf{W}(k) = \underset{\mathbf{W}}{\text{argmin}}\, \|\mathbf{W}\mathbf{y}_{2DZ}(k) - \mathbf{h}_D(k)\|^2$$

$$\hat{\mathbf{h}}_D = \mathbf{R}_{\mathbf{h},\mathbf{h}} \begin{bmatrix} \mathbf{C}_{\text{IBI}}(D) \\ \mathbf{C}_{\text{ISI}}(D) \end{bmatrix}^H \left(\begin{bmatrix} \mathbf{C}_{\text{IBI}}(D) \\ \mathbf{C}_{\text{ISI}}(D) \end{bmatrix} \mathbf{R}_{\mathbf{h},\mathbf{h}} \begin{bmatrix} \mathbf{C}_{\text{IBI}}(D) \\ \mathbf{C}_{\text{ISI}}(D) \end{bmatrix}^H + \mathbf{R}_{\mathbf{n},\mathbf{n}} \right)^{-1} \begin{bmatrix} \mathbf{r}_{c,0} \\ \mathbf{r}_{c,1} \end{bmatrix} \tag{5}$$

$$
\begin{aligned}
\mathbf{W}(k) &= \left([1 J_1 \cdots J_{Z-1}] \otimes \left(\mathbf{R}_{\mathbf{h}_D(k),\mathbf{h}_D(k)} \left[{\mathbf{C}_{\mathrm{IBI}}}^H {\mathbf{C}_{\mathrm{ISI}}}^H\right]\right)\right) \left[\mathbf{T}(k) + \mathbf{R}_{\hat{\mathbf{n}},\hat{\mathbf{n}}}(k) + \sigma_s^2 \mathbf{I}_Z \otimes \Delta_{2D}(k)\right]^{-1}, \qquad (6)\\
\mathbf{T}(k) &= \begin{bmatrix} 1 & J_1 & J_2 & \cdots & J_{Z-1} \\ J_1 & 1 & J_1 & \cdots & J_{Z-2} \\ \vdots & \ddots & \ddots & \ddots & \vdots \\ J_{Z-1} & J_{Z-2} & J_{Z-3} & \cdots & 1 \end{bmatrix} \otimes \left(\begin{bmatrix} \mathbf{C}_{\mathrm{IBI}}(D) \\ \mathbf{C}_{\mathrm{ISI}}(D) \end{bmatrix} \mathbf{R}_{\mathbf{h}_D(k),\mathbf{h}_D(k)} \begin{bmatrix} \mathbf{C}_{\mathrm{IBI}}(D) \\ \mathbf{C}_{\mathrm{ISI}}(D) \end{bmatrix}^H \right),\\
\mathbf{R}_{\hat{\mathbf{n}},\hat{\mathbf{n}}}(k) &= \mathrm{E}\left[\left(\mathbf{n}_{D,0}^T(k+1),\ \mathbf{n}_{D,1}^T(k),\ \cdots,\ \mathbf{n}_{D,1}^T(k-Z+1)\right)^T \left(\mathbf{n}_{D,0}^H(k+1),\ \mathbf{n}_{D,1}^H(k),\ \cdots,\ \mathbf{n}_{D,1}^H(k-Z+1)\right)\right]\\
&= \sigma_n^2 \mathbf{I}_{2DZ},\\
\Delta_{2D}(k) &= \mathrm{diag}\left[\|h_0(k)\|^2, \sum_{p=0}^{1} \|h_p(k)\|^2, \cdots, \sum_{p=0}^{D-1} \|h_p(k)\|^2, \sum_{p=1}^{D-1} \|h_p(k)\|^2, \sum_{p=2}^{D-1} \|h_p(k)\|^2, \cdots, \|h_{D-1}(k)\|^2, 0\right]
\end{aligned}
$$

It is straightforward to show with standard mathematical tools that the resulting $\mathbf{W}(k)$ is as given by (6). Here, $\otimes$ is the Kronecker product, $\mathbf{R}_{\mathbf{h}_D(k),\mathbf{h}_D(k)} = \mathrm{E}\left[\mathbf{h}_D(k)\mathbf{h}_D^H(k)\right]$, $J_n = J_0(2\pi f_D n \Delta T)$ and $\|\cdot\|^2 = \mathrm{E}\left(|\cdot|^2\right)$. Note that the approach presented here does not require a model for the evolution of the channel over time, unlike Kalman filtering approaches (see [15] for a corresponding channel estimation process based on an order-one autoregressive model). The models in [11], [15] introduce approximations with respect to Jake's model $\left(\mathrm{E}\left[h_l(n)h_l^\star(n-1)\right] \approx J_0(2\pi f_D T_n)\,\mathrm{E}\left[|h_l(n)|^2\right]\right)$ which are not required here. Consequently, improved performances are obtained, in particular in a high Doppler scenario.
Obviously, equation (6) is of a certain complexity and does not seem to be compatible with low-complexity hardware implementation constraints. However, it only depends on the Doppler frequency f_D, channel statistics $\mathbf{R}_{\mathbf{h}_D(k),\mathbf{h}_D(k)}$ and the noise covariance $\mathbf{R}_{\hat{\mathbf{n}},\hat{\mathbf{n}}}(k)$. In practice, these quantities are difficult to estimate and usually only rough approximations are available. This is why it is recommended to precalculate $\mathbf{W}(k)$ for a limited number of such parameter sets. The corresponding estimation matrices are then stored in look-up tables in a hardware implementation and are available without requiring any computations. Then, each CIR estimation requires only $D \times 2DZ$ complex multiplications and a corresponding number of additions.

The performance of the CIR estimation in a Doppler environment is presented below in the framework of a 5GHz WLAN, such as IEEE802.11a or BRAN HIPERLAN/2.

V. Simulation results and conclusion

In order to illustrate the performances of our approach, simulations have been performed in the IEEE802.11a [1] or HIPERLAN/2 [2] WLAN context: a $N = 64$ carrier 20MHz bandwidth broadband wireless system operating in the 5.2GHz band using a 16 sample prefix or postfix. A rate $R = 1/2$, constraint length $K = 7$ Convolutional Code (CC) (o171/o133) is used before bit interleaving followed by BPSK/QPSK mapping.

Monte carlo simulations are run and averaged over 2500 realizations of a *normalized* BRAN-A [16] frequency selective channel without Doppler in order to obtain BER curves.

Figure 3 and Figure 4 present results where the CP-OFDM modulator has been replaced by a PRP-OFDM modulator for BPSK and QPSK constellations respectively. The postfix used for the simulations is given by Table I following the derivations in [17]. The curves compare the classical ZF CP-OFDM transceiver (standard IEEE802.11a) and PRP-OFDM with MMSE equalizers over the $P = N + D$ carriers. Each frame contains 2 known training symbols, followed by 100 OFDM data symbols. The training symbols are exploited for CP-OFDM decoding only.

For the PRP-OFDM, after initial acquisition, the channel estimate is then refined by the semi-blind procedure explained in the paper using an averaging window of 40 OFDM symbols. While standard IEEE802.11a/HIPERLAN/2 decoding schemes with initial preamble-based channel estimation perform poorly in high mobility contexts (error floor at a BER of 10^{-2} for BPSK at a mobility of 10m/s), the PRP-OFDM based approach leads to acceptable results up to a mobility of 72m/s for BPSK and 36m/s for QPSK at a 5GHz carrier frequency. At 72m/s, BPSK suffers no performance loss compared to the static CP-OFDM case at a target BER of 10^{-3}. The same performances are obtained for a mobility of 36m/s.

At high Carrier-to-Noise-plus-Interference (C/I) levels, classical preamble based channel estimation schemes outperform the PRP-OFDM based channel estimation approaches presented in this paper. In order to mitigate this problem, two possible solutions are applicable: i) increase of the mean-value calculation window for the estimation of the PRP-OFDM postfix convolved by the channel or ii) perform iterative interference cancellation, i.e. subtract the estimated OFDM data symbols after channel convolution from the received sequence prior to channel estimation. The latter scheme is presented in [18] and works efficiently at the cost of an increase in decoding complexity.

ZF equalization performs poorly due to the occasional amplification of noise on certain carriers that is then spread over all the carriers when changing the resolution of the frequency grid from $P = 80$ carriers back to $N = 64$.

VI. Conclusion

In this contribution a new OFDM modulation has been presented based on a pseudo random postfix: PRP-OFDM, using known samples instead of random data. This multicarrier scheme has the advantage to inherently provide a very simple blind channel estimation exploiting these deterministic values. The same overhead as CP-OFDM is kept while the mobility in a IEEE802.11a context can be increased from 3m/s to 72m/s (BPSK) or 36m/s (QPSK).

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#	Amplitude	#	Amplitude
1	1.5649 - 0.0356i	9	-0.4027 - 0.5203i
2	1.1404 - 0.2923i	10	-0.0363 - 0.0561i
3	-1.1347 + 0.3148i	11	0.2141 + 0.4081i
4	1.5316 + 0.1681i	12	0.3389 + 0.1818i
5	1.6562 + 0.2440i	13	0.0789 + 0.4082i
6	0.0843 + 0.4842i	14	-0.0430 - 0.2456i
7	0.0058 - 0.5014i	15	-0.0926 - 0.1566i
8	-0.9751 - 0.1925i	16	-0.0587 - 0.2248i

TABLE I

TIME DOMAIN SAMPLES OF A SUITABLE POSTFIX.

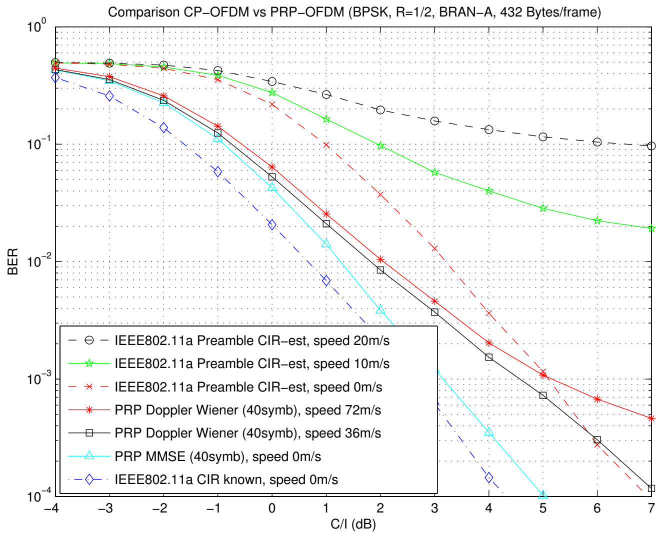

Fig. 3. Simulation results for BPSK, BRAN-A.

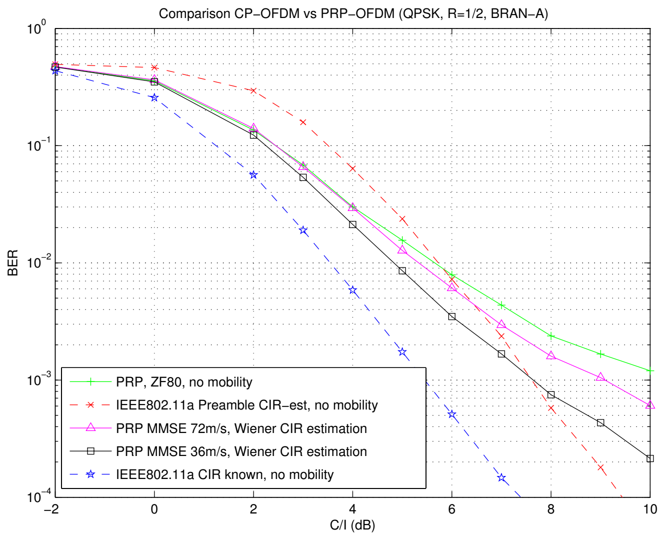

Fig. 4. Simulation results for QPSK, BRAN-A.