PSEUDO RANDOM POSTFIX OFDM BASED MODULATOR FOR NEXT GENERATION 60GHZ WLANS

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ABSTRACT

The IST-2001-32686 project BROADWAY proposes a hybrid dual frequency system operating at 5GHz based on ETSI BRAN HIPERLAN/2 and at 60GHz through an innovative fully ad-hoc extension named HIPERSPOT. In order to allow Single Frequency Network (SFN) deployments at 60GHz while granting compatibility between the two concepts, similar OFDM based physical layers are used with some key improvements for the HIPERSPOT extension. This contribution details a new OFDM modulator based on the use of a Pseudo Random Postfix (PRP-OFDM) and low complexity equalization architectures. The main advantage of this new modulation scheme is the ability to estimate and track the channel variations blindly using order one statistics of the received signal. This scheme is thus very well suited for 60GHz where channel tracking is essential due to the presence of large doppler spreads. Moreover, the proposal of various equalization structures derived from the zero padded transmission schemes, allows implementations ranging from low-complexity/medium performance to increased-complexity/high performance.

1. INTRODUCTION

Latest Wireless Local Area Network (WLAN) solutions provide throughputs on top of the physical layer (PHY) up to 54Mbps at 5GHz (ETSI BRAN HIPERLAN/2, IEEE802.11a). These data rates are foreseen to be insufficient for very dense urban deployment, such as for hot spot coverage. This motivates the proposal of a solution coping with these requirements.

The IST BROADWAY concept proposes an extension of existing WLAN technology in terms of capacity and privacy in an evolutionary way while providing additionally ad-hoc network functionalities [1]. BROADWAY consists in the combination of ETSI BRAN HIPERLAN/2 operating in the 5GHz range and a fully ad-hoc extension in the 60GHz band. The main idea is to offload the 5GHz spectrum into the 60GHz range when in range of communication. Therefore the 5GHz network is used for signalization of both systems and 60GHz for very high data rate peer to peer communications. It is thus planned to provide user data rates up to 500Mbps by offering new modes in the unlicensed 59-65GHz bands.

A first low-complexity mode requires to simply up-convert the 5GHz HIPERLAN/2 20MHz signal to 60GHz. The HIPERSPOT mode, however, introduces some key modifications of the physical layer in order to increase the throughput and reliability. In order to allow large coverage areas through a Single Frequency Network (SFN) deployment and for obvious compatibility reasons, HIPERSPOT is also based on an OFDM modulation similar to the one defined by the HIPERLAN/2 physical layer. However, some modifications are introduced in order to adapt it to the 60GHz constraints.

OFDM has proven to be the preferred modulation for practically all traditional wireless broadband communication systems (Digital Audio and Video Broadcasting: DAB, DVB, and broadband wireless local area networks: WLANs such as IEEE802.11a and ETSI BRAN HIPERLAN/2) thanks to its inherent robustness against multi-path propagation and its inherent low-complexity equalization schemes [2–4]. Two main challenges exist for designing a 60GHz OFDM system: i) Obtain an RF architecture that yields to an acceptable phase noise mask compatible with the OFDM modulation and carrier spacing. The use of fixed oscillators at 60GHz meets this requirement. ii) The ability to fight against large Doppler spreads due to the high frequency of operation which occur even at low mobility and shorten the channel coherence time considerably. This motivates the proposal of enhanced channel estimation and tracking strategies. This paper will focus on the latter issue and proposes a new Pseudo Random Postfix OFDM (PRP-OFDM) modulator. Actually, the virgin 60GHz band allows the proposal of new concepts since yet no standard exists. More robust multicarrier schemes have already been proposed such as the zero padded (ZP) OFDM scheme [5]: they rely on a larger FFT demodulator granting in absence of noise a perfect symbol recovery regardless of the channel null locations. The idea here is to extend this modulator replacing the zeroes by a known pseudo random sequence that would allow an order one semi-blind channel estimation/tracking procedure. Thus, in section 2 the PRP-OFDM transmitter is presented. Several equalization schemes are derived ranging from low-complexity/medium performance to increased-complexity/high performance solutions. Note that altering the postfix doesn't prevent the use of an equalizer of a complexity comparable to the one of the classical Cyclic Prefix based OFDM (CP-OFDM) scheme. The use of a denser frequency grid for the demodulation and equalization introduces a coloration of the noise that needs to be accounted for the Viterbi metric calculation for Maximum Likelihood (ML) decoding in presence of bit interleaved convolutional coded bitstreams (COFDM). Section 3 proposes suboptimal low complexity metrics allowing near to optimal performance decoding. Section 4 details the inherent semi-blind channel estimation and tracking method relying on the exploitation of the pseudo random sequence embedded in the postfix of PRR-OFDM. Finally, section 6 illustrates through simulation results the potential of the new modulation.

2. PRP-OFDM MODULATION AND DEMODULATION

This section presents the PRP-OFDM modulator, a discrete baseband channel model and possible receiver architectures reaching various complexity/performance trade-offs. Please note that the model proposed is not 60GHz specific.

Figure 1: *Discrete model of the PRP-OFDM modulator* depicts the baseband discrete-time block equivalent model of a complete PRP-OFDM transceiver. The ith $N \times 1$ input digital vector [1] $\tilde{\mathbf{s}}_N(i)$ is first modulated by the IFFT matrix $\mathbf{F}_N^H = \frac{1}{\sqrt{N}}\left(W_N^{ij}\right)^H, 0 \le i < N, 0 \le j < N$ where $W_N = e^{-j\frac{2\pi}{N}}$. Then, a deterministic postfix vector $\mathbf{c}_D = (c_0, \dots, c_{D-1})^T$ weighted by a pseudo random phase $\alpha(i) \in \mathbb{C}$ is appended to the IFFT output $\mathbf{s}_N(i)$. The pseudo random $\alpha(i)$ aims at avoiding the introduction of a deterministic component in the transmitted signal resulting in the presence of spectral rays. With $P = N + D$, the corresponding $P \times 1$ transmitted vector is $\mathbf{s}_{\text{ZP}}(i) = \mathbf{F}_{\text{ZP}}^H \tilde{\mathbf{s}}_N(i) + \alpha(i)\mathbf{c}_P$, where

$$\mathbf{F}_{\text{ZP}}^H = \mathbf{M}_{\text{ZP}}\mathbf{F}_N^H = \left[\frac{\mathbf{I}_N}{\mathbf{0}_{D,N}}\right]_{P\times N}\mathbf{F}_N^H \tag{1}$$

and $\mathbf{c}_P = \left(\mathbf{0}_{1,N}, \mathbf{c}_D^T\right)^T$. The subscript ZP makes a reference to zero padded OFDM (ZP-OFDM) [5] where *trailing zeros* are used for the postfix. $\mathbf{s}_{\text{ZP}}(i)$ is serialized and transmitted through the Lth-order FIR channel with impulse response $h_l, l = 0, 1, \dots, L-1$, $L \le D$; then, a zero-mean Gaussian noise is added to its outputs.

Figure 2: *Circularization for ZP-OFDM* illustrates the convolution of $\mathbf{F}_{\text{ZP}}^H \tilde{\mathbf{s}}_N(i)$ by the channel; the corresponding time domain channel matrix has a first column $col_1(\mathbf{H}_o) = \begin{pmatrix} h_0 & h_1 & \dots & h_{L-1} & 0 & \to & 0 \end{pmatrix}$, a second one $col_2(\mathbf{H}_o) = \begin{pmatrix} 0 & h_0 & h_1 & \dots & h_{L-1} & 0 & \to & 0 \end{pmatrix}$, etc. However, such a convolution matrix $\mathbf{H}_o$ is arithmetically costly to handle for equalization. Usually, circulant channel convolution matrices are preferred, since they can be diagonalized on a Fourier basis and thus equalization becomes simple. Herefore, Figure 2: *Circularization for ZP-OFDM* further illustrates how the resulting expression can be re-written without changing the received vector. In particular, the convolution matrix is made circulant by adding some contribution which will be multiplied by zero values of the transmitted vector only; in the end, a circulant convolution matrix occurs and standard frequency-domain equalization schemes known from CP-OFDM can be applied. A second way leading to a circulant channel matrix is illustrated by Figure 3: *Circularization for ZP-OFDM via overlap-add* using the well-known 'overlap-add' technique, the last D samples of the received vector are added to the first ones. Afterwards, the last D samples are truncated and simple equalization schemes are possible. Please note that the first method introduces noise correlation, since the equalization by a diagonal matrix is done in the $P \times P$ Fourier domain and the transformation back to the original $N \times N$ domain mixes up the noise. This problem does not exist in the case of the 'overlap-add' technique where the equalization is directly performed in the $N \times N$ Fourier domain. Of course, the convolution of the postfix has to be considered separately.

Figure 4: *Circularization for PRP-OFDM* illustrates the impact if the pseudo-randomly weighted postfix is considered within the channel convolution matrix. If the weighting factor $\alpha(i)$ would be $\alpha(i) = const$ for all cases, the inter-block-interference (IBI) would be sufficient to make the convolution matrix circulant, similar to the CP-OFDM case. A similar technique has been applied by Deneire et al. [6] to single carrier systems. With any $\alpha(i)$, however, the upper triangular part of the channel convolution matrix is weighted by a factor $\beta_i = \frac{\alpha(i-1)}{\alpha(i)}$. Such a matrix $\mathbf{H}_{\beta_i}$ is usually called *pseudo circulant*. With $\mathbf{b}_P$ being a Gaussian noise vector, the received $P \times 1$ vector is then:

$$\begin{aligned} \mathbf{r}_P(i) &= \mathbf{H}\mathbf{F}_{\text{ZP}}^H\tilde{\mathbf{s}}_N(i) + \mathbf{H}_{\beta_i}(\alpha(i)\cdot\mathbf{c}_P) + \mathbf{b}_P & (2)\\ &= \mathbf{H}_o\mathbf{F}_N^H\tilde{\mathbf{s}}_N(i) + \mathbf{H}_{\beta_i}(\alpha(i)\cdot\mathbf{c}_P) + \mathbf{b}_P & (3)\\ &= \mathbf{H}_{\beta_i}\begin{pmatrix} \mathbf{F}_N^H\tilde{\mathbf{s}}_N(i) \\ \alpha(i)\cdot\mathbf{c}_D \end{pmatrix} + \mathbf{b}_P. & (4) \end{aligned}$$

Corresponding to the first N columns of $\mathbf{H}$, the $P \times N$ matrix $\mathbf{H}_o = \mathbf{H}_{0\le i<P, 0\le j<N}$ is Toeplitz and is always guaranteed to be invertible in the Moore-Penrose sense, which enables symbol recovery regardless of the channel zero locations. Please note that (2) generalizes other OFDM modulation schemes [5]. In the case of zero padded OFDM (ZP-OFDM), the deterministic postfix vector $\mathbf{c}_P$ in (2) must be set to zero and the equation holds. In the case of cyclic prefix OFDM, $\mathbf{c}_P$ still is set to zero and the modulation matrix $\mathbf{F}_{\text{ZP}}^H$ is replaced by $\mathbf{F}_{\text{CP}}^H$, being expressed as

$$\begin{aligned} \mathbf{F}_{\text{CP}}^H &= \mathbf{M}_{\text{CP}}\mathbf{F}_N^H & (5)\\ &= \left[\frac{\mathbf{0}_{D,N-D} \mid \mathbf{I}_D}{\mathbf{I}_N}\right]_{P\times N}\mathbf{F}_N^H. & (6) \end{aligned}$$

The equalization of the received vector $\mathbf{r}_P(i)$ may be based on two fundamentally different approaches.

Firstly, (2) may be reduced to the ZP-OFDM case by subtracting the known postfix convolved by the pseudo-circulant channel matrix: $\mathbf{r}_P^{\text{ZP}}(i) = \mathbf{r}_P(i) - \hat{\mathbf{H}}_{\beta_i}\cdot\alpha(i)\cdot\mathbf{c}_P$, with $\hat{\mathbf{H}}_{\beta_i}$ being an estimation of $\mathbf{H}_{\beta_i}$. The corresponding ZF and MMSE equalizers are given by [5, 9]: $\mathbf{G}_{\text{ZF}} = \mathbf{F}_N\mathbf{H}_o^\dagger$ and $\mathbf{G}_{mmse} = \mathbf{F}_N\mathbf{H}_o^H(\sigma_n^2\mathbf{I} + \mathbf{H}_o\mathbf{H}_o^H)^{-1}$, where σ_n^2 denotes the AWN variance and the symbols are assumed without loss of generality to have variance $\sigma_s^2 = 1$. [5] presents several alternative equalization approaches for ZP-OFDM symbols.

Secondly, the equalization of (2) may be based on the diagonalization of the pseudo circulant matrix $\mathbf{H}_{\beta_i}$, exploiting the relation

$$\mathbf{H}_{\beta_i} = \mathbf{V}_{\mathbf{P}}^{-1}(i)\,diag\left\{H(\beta_i^{-\frac{1}{P}}), \dots, H(\beta_i^{-\frac{1}{P}}\cdot e^{j2\pi\cdot\frac{P-1}{P}})\right\}\mathbf{V}_{\mathbf{P}}(i), \tag{7}$$

where

$$\begin{aligned} \mathbf{V}_{\mathbf{P}}(i) &= \left(\frac{1}{P}\sum_{n=0}^{P-1}|\beta_i|^{\frac{2n}{P}}\right)^{-\frac{1}{2}}\mathbf{F}_N\,diag\left\{1, \beta_i^{\frac{1}{P}}, \dots, \beta_i^{\frac{P-1}{P}}\right\} & (8)\\ H(z) &= \sum_{n=0}^{P-1} z^{-n}\cdot h_n. & (9) \end{aligned}$$

It is recommended to choose β_i to be a phase $\beta_i = e^{j\frac{m_i}{M}}$, $m_i \in (0, 1, \dots, M-1)$. Otherwise, important performance drawbacks are observed.

The corresponding equalization matrices for the ZF and

[1] 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).

MMSE case are

$$\begin{aligned} \mathbf{G}_{\text{ZF}}^{PRP}(i) &= \mathbf{F}_N \left[\mathbf{I}_N \mathbf{0}_{N,D}\right] \mathbf{H}_{\beta_\mathbf{i}}^{-1} \\ &= \mathbf{F}_N \left[\mathbf{I}_N \mathbf{0}_{N,D}\right] \mathbf{V}_\mathbf{P}^H(i)\mathbf{D}_i^{-1}\mathbf{V}_\mathbf{P}(i), \qquad (10) \\ \mathbf{G}_{mmse}^{PRP}(i) &= \mathbf{F}_N \left[\mathbf{I}_N \mathbf{0}_{N,D}\right] \mathbf{H}_{\beta_\mathbf{i}}^H \left(\sigma_n^2 \mathbf{I} + \mathbf{H}_{\beta_\mathbf{i}} \mathbf{H}_{\beta_\mathbf{i}}^H\right)^{-1} \\ &= \mathbf{F}_N \left[\mathbf{I}_N \mathbf{0}_{N,D}\right] \mathbf{V}_\mathbf{P}^H(i)\mathbf{D}_i^H \left(\sigma_n^2 \mathbf{I} + \mathbf{D}_i \mathbf{D}_i^H\right)^{-1} \mathbf{V}_\mathbf{P}(i), \qquad (11) \end{aligned}$$

where σ_n^2 denotes the AWN variance and the symbols are assumed without loss of generality to have variance $\sigma_s^2 = 1$. Due to the problem of noise correlation mentioned before, the MMSE equalization is usually the preferred solution.

Please note that the expression $\left[\mathbf{I}_N \mathbf{0}_{N,D}\right]$ in (11) and (10) erases the equalized constant postfix. Since the latter one is known a-priori, it might be used in order to estimate the signal-to-noise (SNR) ratio, if necessary.

As a conclusion to this section, it shall be pointed out that PRP-OFDM leads to a very simple modulation scheme on the transmitter side. In the receiver, a variety of demodulation and equalization approaches are possible, each characterized by different complexity/performance trade-offs.

3. DECODING PRP-OFDM SYMBOLS

The upper section presents several equalization approaches. Now, the equalized symbols are considered from a decoding point of view, assuming a channel coder in the transmitter. Usually, a maximum-likelihood decoding algorithm is applied in the receiver. Herefore, vector (2) is expressed after equalization as follows, using any $N \times P$ equalization matrix $\mathbf{G}$ of the ones previously presented:

$$\hat{\tilde{\mathbf{s}}} = \mathbf{G}\mathbf{r}_P(i) = \mathbf{G}_d \tilde{\mathbf{s}}_N(i) + \hat{\mathbf{b}}_N, \qquad (12)$$

where $\mathbf{G}_d$ is a diagonal weighting matrix and $\hat{\mathbf{b}}_N$ the resulting noise vector which is assumed to be Gaussian and zero-mean. For decoding, $\mathbf{G}_d$ and the covariance matrix of $\hat{\mathbf{b}}_N$ are required. These are presented in the following, based on

$$\begin{aligned} \hat{\tilde{\mathbf{s}}} &= \mathbf{G}\mathbf{r}_P(i) \qquad (13) \\ &= \mathbf{G}\left(\mathbf{H}_{\beta_\mathbf{i}} \begin{pmatrix} \mathbf{F}_N^H \tilde{\mathbf{s}}_N(i) \\ \alpha(i)\cdot \mathbf{c}_D \end{pmatrix} + \mathbf{b}_P\right) \qquad (14) \\ &= \mathbf{G}_d \tilde{\mathbf{s}}_N(i) + \mathbf{G}_f \tilde{\mathbf{s}}_N(i) + \mathbf{G}_p \alpha(i)\cdot \mathbf{c}_D + \mathbf{G}\mathbf{b}_P, \qquad (15) \end{aligned}$$

where $\mathbf{G}_d$ is a $N \times N$ diagonal matrix and $\mathbf{G}_f$ a $N \times N$ full matrix with the main diagonal being zero such that

$$\begin{aligned} (\mathbf{G}_d + \mathbf{G}_f)\tilde{\mathbf{s}}_N(i) &= \mathbf{G}\mathbf{H}_{\beta_\mathbf{i}} \begin{pmatrix} \mathbf{F}_N^H \tilde{\mathbf{s}}_N(i) \\ \mathbf{0}_{D,1} \end{pmatrix} \qquad (16) \\ &= \mathbf{G}\mathbf{H}_{\beta_\mathbf{i}} \begin{bmatrix} \mathbf{I}_N \\ \mathbf{0}_{D,N} \end{bmatrix} \mathbf{F}_N^H \tilde{\mathbf{s}}_N(i). \qquad (17) \end{aligned}$$

$\mathbf{G}_p$ is a $N \times D$ matrix containing the last D columns of the matrix $\mathbf{G}\mathbf{H}_{\beta_\mathbf{i}}$. Thus, $\mathbf{G}_f \tilde{\mathbf{s}}_N(i)$ is the inter-symbol interference which shall be approximated to be Gaussian noise in the following. Thus, the total noise vector is $\hat{\mathbf{b}}_N = \mathbf{G}_f \tilde{\mathbf{s}}_N(i) + \mathbf{G}\mathbf{b}_P + \mathbf{G}_p \alpha(i)\cdot \mathbf{c}_D$ and the corresponding covariance is

$$R_{\hat{\mathbf{b}}_N,\hat{\mathbf{b}}_N} = \sigma_s^2 \mathbf{G}_f \mathbf{G}_f^H + \sigma_b^2 \mathbf{G}\mathbf{G}^H + \mathbf{G}_p \mathbf{c}_D \mathbf{c}_D^H \mathbf{G}_p^H \qquad (18)$$

for the equalization based on pseudo-circulant matrices. If the received vector is transformed to ZP-OFDM, as explained before, $\mathbf{c}_D$ is set to zero. Hereby, σ_s^2 is the signal sample variance and σ_b^2 the noise sample variance.

For maximum-likelihood decoding, usually a log-likelihood approach is chosen based on a multivariate Gaussian law leading to an expression of the type

$$\hat{\mathbf{d}}_n = \max_{\hat{\mathbf{d}}_n} \left\{ -\sum_{i=0}^{S-1} \left(\mathbf{G}_d m_N\left(\hat{\tilde{\mathbf{d}}}(n,i)\right) - \hat{\tilde{\mathbf{s}}}(i)\right)^H \cdot R_{\hat{\mathbf{b}}_N,\hat{\mathbf{b}}_N}^{-1} \cdot \left(\mathbf{G}_d m_N\left(\hat{\tilde{\mathbf{d}}}(n,i)\right) - \hat{\tilde{\mathbf{s}}}(i)\right) \right\}. \qquad (19)$$

In (19), $\hat{\mathbf{d}}_n$ is a vector containing an estimation of the original information bits encoded in the transmitter, $\hat{\tilde{\mathbf{d}}}(n,i)$ contains the corresponding bits after encoding, puncturing, etc. where n is an index over all encoded information bits and i is the corresponding OFDM symbol number. S is the number of OFDM symbols in the sequence to be decoded, $m_N(\cdot)$ is an operator representing the mapping of encoded information bits onto the N carriers of one OFDM symbol. One way to achieve a reasonable decoding complexity is to approximate (18) by a matrix containing its main diagonal only. Then, standard VITERBI decoding is applicable. However, the calculation of (18) is costly.

The question is whether an exact calculation of the covariance (18) and its inverse is really necessary or whether neglecting the inter-symbol-interference and the interference introduced by the postfix is possible while keeping a reasonable decoding performance in practice. This problem has been considered for a pseudo circulant MMSE decoding approach (11). Hereby, the equalization is performed in the optimum manner. For the metric calculation, however, the the noise covariance matrix is approximated by

$$\begin{aligned} R_{\hat{\mathbf{b}}_N,\hat{\mathbf{b}}_N} &= \sigma_b^2 \mathbf{G}\mathbf{G}^H \qquad (20) \\ &\approx \sigma_b^2 \cdot Diag\left(\frac{|H_0|^2}{\left(|H_0|^2+\sigma_b^2\right)^2}, \ldots, \frac{|H_{N-1}|^2}{\left(|H_{N-1}|^2+\sigma_b^2\right)^2}\right). \qquad (21) \end{aligned}$$

In other words, for the metric weighting, the equalization has been approximated by the standard term known from CP-OFDM, where $H_n, n = 0, \ldots, N-1$ are the frequency domain channel coefficients. Moreover, the matrix $\mathbf{G}_d$ has been approximated in a corresponding way by

$$\begin{aligned} \mathbf{G}_d &= Diag\left(\mathbf{G}\mathbf{H}_{\beta_\mathbf{i}} \begin{bmatrix} \mathbf{I}_N \\ \mathbf{0}_{D,N} \end{bmatrix} \mathbf{F}_N^H\right) \qquad (22) \\ &\approx Diag\left(\frac{|H_0|^2}{|H_0|^2+\sigma_b^2}, \ldots, \frac{|H_{N-1}|^2}{|H_{N-1}|^2+\sigma_b^2}\right). \qquad (23) \end{aligned}$$

As illustrated by Figure 5: *BER for the HiperLAN/2 channel model A, BPSK* for a HIPERLAN/2 like scenario with BPSK constellations, the above presented approximations degrade the bit-error-rate (BER) performance by approx. 0.2dB only, which is very acceptable. This result is very motivating, since it proves that even in the context of coded OFDM (COFDM), PRP-OFDM offers the possibility to find low-cost equalization architectures.

4. CHANNEL ESTIMATION

As mentioned in the introduction, PRP-OFDM allows simple and low-complexity channel estimation. The general approach is presented in the following. In a first time, the channel impulse response is assumed to be static.

Let $\mathbf{H}_D$ be a $D \times D$ circulant channel matrix with the first row being $row_1(\mathbf{H}_D) = (\ h_0 \quad 0 \quad \rightarrow \quad 0 \quad h_{L-1} \quad \ldots \quad h_1\ )$. Moreover, $\mathbf{H}_{D,0}$ shall contain the lower triangular part of $\mathbf{H}_D$ including the main diagonal; $\mathbf{H}_{D,1}$ shall contain the upper triangular part of $\mathbf{H}_D$, such that $\mathbf{H}_D = \mathbf{H}_{D,0} + \mathbf{H}_{D,1}$. With $\mathbf{s}_N(i) = (s_0(i), \ldots, s_{N-1}(i))^T$, $\mathbf{s}_{N,0}(i) = (s_0(i), \ldots, s_{D-1}(i))^T$ and $\mathbf{s}_{N,1}(i) = (s_{N-D}(i), \ldots, s_{N-1}(i))^T$ as well as $\mathbf{b}_P(i) = (b_0(i), \ldots, b_{P-1}(i))^T$, $\mathbf{b}_{D,0}(i) = (b_0(i), \ldots, b_{D-1}(i))^T$ and $\mathbf{b}_{D,1}(i) = (b_{P-D}(i), \ldots, b_{P-1}(i))^T$, the received vector $\mathbf{r}_P(i)$ can be expressed as

$$\mathbf{r}_P(i) = \begin{pmatrix} \mathbf{H}_{D,0}\mathbf{s}_{N,0}(i) + \mathbf{H}_{D,1}(\alpha(i-1)\mathbf{c}_D) + \mathbf{b}_{D,0} \\ \vdots \\ \mathbf{H}_{D,1}\mathbf{s}_{N,1}(i) + \mathbf{H}_{D,0}(\alpha(i)\mathbf{c}_D) + \mathbf{b}_{D,1} \end{pmatrix} \tag{24}$$

Assuming $\mathbf{s}_N(i)$ to be zero-mean, 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 as follows:

$$\hat{\mathbf{h}}_{c,1} = E\left[\frac{1}{\alpha(i-1)}\mathbf{r}_{P,0}(i)\right] \tag{25}$$

$$= \underbrace{E\left[\frac{\mathbf{H}_{D,0}\mathbf{s}_{N,0}(i)}{\alpha(i-1)}\right]}_{=0} + E\left[\mathbf{H}_{D,1}\mathbf{c}_D\right] \tag{26}$$

$$= \mathbf{H}_{D,1}\mathbf{c}_D, \tag{27}$$

$$\hat{\mathbf{h}}_{c,0} = E\left[\frac{1}{\alpha(i)}\mathbf{r}_{P,1}(i)\right] \tag{28}$$

$$= \underbrace{E\left[\frac{\mathbf{H}_{D,1}\mathbf{s}_{N,1}(i)}{\alpha(i)}\right]}_{=0} + E\left[\mathbf{H}_{D,0}\mathbf{c}_D\right] \tag{29}$$

$$= \mathbf{H}_{D,0}\mathbf{c}_D. \tag{30}$$

Based on these results, two different ways of estimating the channel impulse response are presented in the following.

Firstly, (27) and (30) may be combined as

$$\hat{\mathbf{h}}_c = \hat{\mathbf{h}}_{c,1} + \hat{\mathbf{h}}_{c,0} \tag{31}$$

$$= \mathbf{H}_D\mathbf{c}_D \tag{32}$$

$$= \mathbf{C}_D\mathbf{h}_D \tag{33}$$

$$= \mathbf{F}_D^H\tilde{\mathbf{C}}_D\mathbf{F}_D\mathbf{h}_D, \tag{34}$$

where $\mathbf{C}_D$ is a $D \times D$ circulant matrix with the first row being $row_1(\mathbf{C}_D) = (\ c_0 \quad c_{D-1} \quad c_{D-2} \quad \ldots \quad c_1\ )$ and $\tilde{\mathbf{C}}_D = diag\{\mathbf{F}_D\mathbf{c}_D\}$. Hereby, the commutativity of the convolution has been exploited. Thus, the channel impulse response is obtained by

$$\hat{\mathbf{h}}_D = \mathbf{C}_D^{-1}\hat{\mathbf{h}}_c \tag{35}$$

$$= \mathbf{F}_D^H\tilde{\mathbf{C}}_D^{-1}\mathbf{F}_D\hat{\mathbf{h}}_c. \tag{36}$$

Please note that $\tilde{\mathbf{C}}_D^{-1}$ is a diagonal matrix that is a-priori known to both the transmitter and receiver and can thus be precalculated. Then, $\hat{\mathbf{h}}_D$ is usually transformed to the $P \times P$ frequency domain by $\hat{\tilde{\mathbf{h}}}_P = \mathbf{F}_P \begin{bmatrix} \mathbf{I}_D \\ \mathbf{0}_{N,D} \end{bmatrix} \hat{\mathbf{h}}_D$. Alternatively, a MMSE approach can be used in order to estimate the channel impulse response. All results presented above are based on the assumption that the channel is time-invariant. In the case of a Doppler scenario, the optimum channel impulse response estimate in the mean-square error sense is obtained not by calculating $\hat{\mathbf{h}}_c$ via a mean value approach, but rather by Kalman filtering. A corresponding study will be presented in a separate paper.

This section has thus demonstrated that PRP-OFDM allows a simple channel estimation by replacing the cyclic prefix of CP-OFDM by a deterministic vector, known to both the transmitter and the receiver. Contrarily to CP-OFDM, no additional training symbols nor pilot tones are required for channel estimation. But, a combination with the latter ones, the channel estimation can be further refined.

5. CHOICE OF THE POSTFIX

The deterministic postfix of length D is proposed to be designed meeting the following criteria: (i) low peak-to-average-power ratio (PAPR) in time domain, (ii) low out-of-band radiation, i.e. concentrate signal power on useful carriers and (iii) spectral flatness over useful carriers, i.e. SNR of channel estimates shall be as constant as possible over all useful carriers. One way of obtaining a suitable postfix is to express the upper criteria by cost functions and to perform a multi-dimensional optimization in order to find the postfix minimizing the given cost function. A corresponding study will be presented in a separate paper. If the PAPR criterium is not an issue, the solution is given by a corresponding Kaiser-window.

6. SIMULATION RESULTS

Figure 5: *BER for the HiperLAN/2 channel model A, BPSK* presents simulation results that have been performed in the framework of IEEE802.11a and BRAN HIPERLAN/2. The CP-OFDM modulator has been replaced by a PRP-OFDM modulator inserting a suitable postfix. All simulations have been performed for BPSK symbols, a code rate of $R = 1/2$ and the BRAN-A channel model [7] without Doppler for a frame length of four LCHs containing 54 Bytes of uncoded data each. The channel impulse response has been normalized for each realization to $\sum_{n=0}^{L-1} |h_n|^2 = 1$:

- PRP-OFDM based on the MMSE equalization (11), optimum case, i.e. the receiver knows the exact channel impulse response.
- as before, but channel estimation is performed by mean value calculation over 40 received symbols tolerating the resulting latency ; i.e. 40 OFDM symbols are received before the decoding is begun and thus a latency of 40 OFDM symbols lengths is introduced; optimum metrics (18) are used for the VITERBI decoding.
- as before, but simplified metrics (21), (23) are used for the VITERBI decoding.
- PRP-OFDM based on the MMSE equalization (11), channel impulse response is estimated based on an optimum combination in the mean square error sense of the HIPERLAN/2 learning symbols (C-Field) and up to 40 received PRP-OFDM symbols; no latency is introduced.

- Standard HIPERLAN/2 with ZF equalization.

If the channel impulse response is estimated over a large number K of PRP-OFDM symbols, a total gain of approx. 3dB is possible. With the very realistic number of 40 OFDM symbols, a gain of approx. 2dB is possible. With approx. 200 PRP-OFDM symbols the 3dB limit is reached. The simplified metrics lead to an additional loss of approx. 0.2dB, but the decoding complexity is largely decreased. In order to decrease the system latency, the channel impulse response estimation obtained by a learning symbol and the PRP-OFDM symbols can be combined in the mean square error sense; here, a slight loss of approx. 0.2dB is the price for the speedier availability of the received data.

7. CONCLUSION AND ACKNOWLEDGEMENT

A new OFDM modulation scheme has been presented: pseudo random postfix OFDM (PRP-OFDM). Several equalization approaches have been discussed, allowing to choose in the receiver different complexity/performance trade-offs. In all cases, the transmitter is independent of the equalization scheme chosen in the receiver. Then, assuming channel coding in the transmitter, the way to perform maximum-likelihood decoding in the receiver is indicated by considering inter-symbol-interference as Gaussian noise. It has been illustrated how the channel impulse response is estimated based on the deterministic postfixes only. The design criteria of the deterministic postfix have been briefly indicated. The simulation results show a performance gain of up to 3dB compared to IEEE802.11a and BRAN HIPERLAN/2 for BPSK constellations in a BRAN-A channel environment [7] without Doppler.

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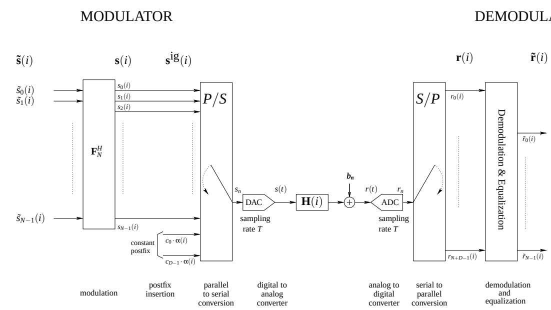

Figure 1: Discrete model of the PRP-OFDM modulator.

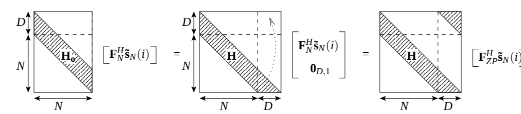

Figure 2: Circularization for ZP-OFDM.

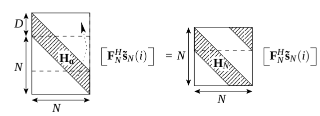

Figure 3: Circularization for ZP-OFDM via overlap-add.

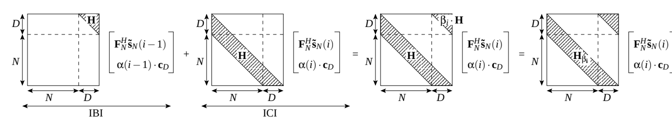

Figure 4: Circularization for PRP-OFDM.

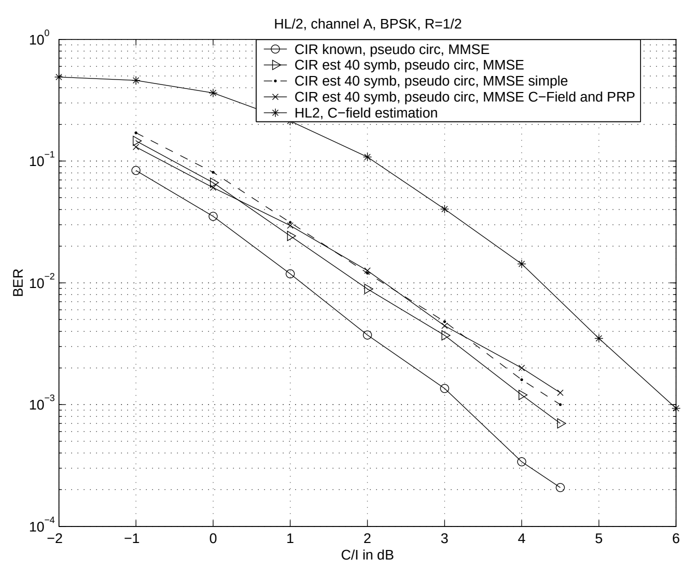

Figure 5: BER for the HiperLAN/2 channel model A, BPSK.