Noise Interpolation for Unique Word OFDM

Alexander Onic and Mario Huemer

***Abstract*—Unique word-orthogonal frequency division multiplexing (UW-OFDM) is known to feature an excellent bit error ratio performance when compared to conventional OFDM using cyclic prefixes (CP). In recent papers classical and Bayesian linear data estimators as well as the non-linear sphere decoding (SD) approach have been investigated in the UW-OFDM context. In general, the remaining error samples after a linear data estimation in UW-OFDM are correlated. In this work, noise interpolation (NI) is proposed, which uses preliminary data decisions together with Wiener interpolation to exploit these correlations. It turns out that the selection of samples to be used for NI is crucial. With the sample selection policy suggested in this work, NI clearly outperforms LMMSE data estimation, and due to its lower complexity compared to SD, it can be regarded as an attractive compromise for UW-OFDM reception.**

***Index Terms*—LMMSE estimation, noise interpolation, noise prediction, OFDM, sphere decoding, unique word.**

I. Introduction

THE Unique Word (UW) OFDM signaling scheme introduced in [1] uses a deterministic sequence in the guard interval instead of the usual cyclic prefix. The introduction of the unique word within the DFT (discrete Fourier transform) interval is conducted by introducing redundancy in the frequency domain. This redundancy can be utilized by sophisticated receivers in order to improve the data estimation performance compared to straight forward receivers like e.g. a simple linear zero forcing estimator. Most prominently, linear minimum mean square error (LMMSE) data estimation [2] and sphere decoding (SD) [3], [4], especially designed for UW-OFDM, are able to exploit the redundancy advantageously, and demonstrate an impressive performance in terms of the bit error ratio (BER). While the LMMSE estimator is a quite simple receiver with good performance, the SD is the optimum receiver for an uncoded UW-OFDM system. Besides the higher complexity of non-linear estimation approaches, the SD also has the drawback that reliability information to be used in soft channel decoders is very difficult to compute. Usually, only an approximation can be derived reasonably.

Manuscript received February 19, 2014; revised April 08, 2014; accepted April 09, 2014. Date of publication April 15, 2014; date of current version April 24, 2014. This work was supported by the Austrian Science Fund (FWF) under Grant I683-N13. The associate editor coordinating the review of this manuscript and approving it for publication was Prof. Richard K. Martin.

The authors are with the Institute of Signal Processing, Johannes Kepler University Linz, 4040 Linz, Austria (e-mail: alexander.onic@jku.at; mario.huemer@jku.at).

Color versions of one or more of the figures in this paper are available online at http://ieeexplore.ieee.org.

Digital Object Identifier 10.1109/LSP.2014.2317512

The second order statistics of the remaining error of the data estimates (remaining "noise") after an LMMSE estimation reveal a correlation of the noise samples. This correlation together with preliminary data decisions can be exploited in order to obtain more accurate data estimates. For single-carrier transmission this technique has been studied extensively, but for causality reasons, it is usually referred to as noise *prediction*, see e.g. [5]–[7]. Due to the block processing in UW-OFDM, all values in one OFDM symbol can be used for the aspired *interpolation* of noise samples. However, the sample selection turns out to be an essential issue. Parallels that can be drawn to decision feedback equalization are addressed in [8].

The paper is organized as follows: The UW-OFDM system description will be presented in Section II, and LMMSE data estimation and sphere decoding are briefly presented in Section III. Next, the idea of decision directed noise interpolation and a possible implementation of the sample selection process are introduced in Section IV. In order to quantify the detection performance of the proposed concept, we will show simulation results of a unique word OFDM system in a multipath as well as in the AWGN (additive white Gaussian noise) environment in Section V, and we compare the results with the ones after LMMSE estimation and sphere decoding, respectively. In Section VI we conclude this work.

II. Unique Word OFDM System Model

We briefly review the approach of introducing unique words in OFDM, for further details see [1]. Let $\mathbf{x}_u \in \mathbb{C}^{N_u \times 1}$ be a predefined sequence, which we call unique word. This unique word shall form the tail of the OFDM time domain symbol vector. Hence, the time domain symbol vector, as the result of the length N IDFT (inverse DFT), consists of two parts and is of the form $\mathbf{x}' = [\mathbf{x}_p^\top \quad \mathbf{x}_u^\top]^\top \in \mathbb{C}^{N \times 1}$, whereas only $\mathbf{x}_p \in \mathbb{C}^{(N-N_u)\times 1}$ is payload that is affected by the data, and thus random. In the concept suggested in [9], we generate an OFDM symbol $\mathbf{x} = [\mathbf{x}_p^\top \quad \mathbf{0}^\top]^\top$ with a zero UW in a first step and determine the final UW-OFDM symbol $\mathbf{x}' = \mathbf{x} + [\mathbf{0}^\top \quad \mathbf{x}_u^\top]^\top$ by adding the desired UW in time domain in a second step. As in conventional OFDM, the QAM data symbols denoted by the vector $\mathbf{d} \in \mathbb{C}^{N_d \times 1}$ and a number of N_z zero subcarriers, usually at the band edges and at DC, are specified in frequency domain as part of the vector $\tilde{\mathbf{x}}$. UW-OFDM has the additional zero word constraint in time domain as part of the vector $\mathbf{x} = \mathbf{F}_N^{-1}\tilde{\mathbf{x}}$, where $\mathbf{F}_N$ is the N point DFT matrix with the elements $[\mathbf{F}_N]_{kl} = \mathrm{e}^{-\mathrm{j}\frac{2\pi}{N}kl}$ for $k, l = 0, 1, \ldots, N-1$. The final UW-OFDM frequency domain symbol is assembled by

$$\tilde{\mathbf{x}} = \mathbf{B}\mathbf{G}\mathbf{d}, \tag{1}$$

with the help of an appropriate complex valued generator matrix $\mathbf{G}^{(N-N_z)\times N_d}$ and the matrix $\mathbf{B} \in \{0,1\}^{N \times (N-N_z)}$ which introduces the zero subcarriers and consists of zero-rows at the positions of the zero subcarriers, and of appropriate unit row vectors at all other positions. Let $\mathbf{W}$ be the matrix built by the $N_r = N - N_z - N_d$ lowermost rows of $\mathbf{F}_N^{-1}\mathbf{B}$. Then, in order to generate the zero UW in time domain, a valid UW-OFDM generator matrix has to fulfill

$$\mathbf{WG} = \mathbf{0}, \tag{2}$$

which says that the columns of a valid $\mathbf{G}$ have to lie in the null space of $\mathbf{W}$. This generator matrix has to be designed carefully, two variants will be used in the following.

As derived in detail in [1], [2], [9], after subtraction of a UW depending portion, the receive symbol $\mathbf{y} \in \mathbb{C}^{(N-N_z)\times 1}$ can be modeled as $\mathbf{y} = \tilde{\mathbf{H}}\mathbf{G}\mathbf{d} + \mathbf{n}$, where the diagonal matrix $\tilde{\mathbf{H}} \in \mathbb{C}^{(N-N_z)\times(N-N_z)}$ contains the sampled channel frequency response on its main diagonal, and $\mathbf{n}$ represents a zero-mean Gaussian noise vector in frequency domain with the covariance matrix $\mathbf{C}_{nn} = \sigma_n^2\mathbf{I}$. The channel propagation and generator matrix are treated together as a channel matrix $\mathbf{H} = \tilde{\mathbf{H}}\mathbf{G}$, which yields the linear system model

$$\mathbf{y} = \mathbf{Hd} + \mathbf{n}. \tag{3}$$

A. Systematic UW-OFDM Symbol Generation

In our original UW-OFDM concept presented in [1], [2], [9], we chose

$$\mathbf{G} = \mathbf{P}\begin{bmatrix}\mathbf{I}\\ \mathbf{T}\end{bmatrix}, \tag{4}$$

where $\mathbf{P} \in \{0,1\}^{(N-N_z)\times(N-N_z)}$ is a carefully selected permutation matrix, cf. [10]–[14]. Let $\mathbf{M} = \mathbf{WP} = [\mathbf{M}_1 \quad \mathbf{M}_2]$ with $\mathbf{M}_1 \in \mathbb{C}^{N_r\times N_d}$ and $\mathbf{M}_2 \in \mathbb{C}^{N_r\times N_r}$, then the constraint (2) is fulfilled by choosing $\mathbf{T} = -\mathbf{M}_2^{-1}\mathbf{M}_1 \in \mathbb{C}^{N_r\times N_d}$. We call $\mathbf{r} = \mathbf{Td} \in \mathbb{C}^{N_r\times 1}$ the vector of redundant symbols. Then, according to (1), the frequency domain version of the UW-OFDM symbol is given by $\tilde{\mathbf{x}} = \mathbf{BP}[\mathbf{d}^\top \quad \mathbf{r}^\top]^\top$. The individual data symbols from $\mathbf{d}$ can still be observed in $\tilde{\mathbf{x}}$, which is why this approach is called systematic symbol generation.

B. Non-Systematic UW-OFDM Symbol Generation

In [15] we introduced the concept of non-systematically generated UW-OFDM, where we propose code generator matrices $\mathbf{G}$ that distribute the redundancy over all subcarriers. The approach for the non-systematic symbol generation is given by

$$\mathbf{G} = \mathbf{A}\begin{bmatrix}\mathbf{I}\\ \mathbf{T}\end{bmatrix}, \tag{5}$$

with a real non-singular matrix $\mathbf{A} \in \mathbb{R}^{(N-N_z)\times(N-N_z)}$ that replaces the permutation matrix and spreads the redundancy over the whole bandwidth. In [15], two generator matrices are constructed as a result of numerical optimizations with different initializations. In this work, the generator matrix obtained with the initialization $\mathbf{A}^{(0)} = \mathbf{P}$ is used.

III. Reference Receivers for UW-OFDM

The UW-OFDM concept allows for a number of different linear and non-linear data estimation approaches. As reference receivers in this work we use the LMMSE estimator as the best performing linear estimator, as well as sphere decoding, which constitutes the optimum BER reference for uncoded transmission.

Following Bayesian estimation theory, the linear system model (3) and the linear approach $\hat{\mathbf{d}} = \mathbf{Ey}$ yield an LMMSE data estimation matrix [2]

$$\mathbf{E}_{\mathrm{LMMSE}} = \left(\mathbf{H}^{\mathsf{H}}\mathbf{H} + \frac{\sigma_n^2}{\sigma_d^2}\mathbf{I}\right)^{-1}\mathbf{H}^{\mathsf{H}}. \tag{6}$$

The covariance matrix of the remaining error $\mathbf{e} = \hat{\mathbf{d}} - \mathbf{d}$ can be used to determine reliability information for a soft decision channel decoder and is given by

$$\mathbf{C}_{ee} = \sigma_n^2\left(\mathbf{H}^{\mathsf{H}}\mathbf{H} + \frac{\sigma_n^2}{\sigma_d^2}\mathbf{I}\right)^{-1}. \tag{7}$$

Sphere decoding is able to achieve optimum detection results for uncoded transmission [3], as it constitutes a maximum likelihood sequence estimator. The SD method solves the minimization problem

$$\hat{\mathbf{d}} = \underset{\mathbf{d}\in\mathcal{A}^{N_d}}{\arg\min}\, \|\mathbf{Hd} - \mathbf{y}\|_2^2, \tag{8}$$

over all possible data vectors $\mathbf{d} \in \mathcal{A}^{N_d}$, if $\mathcal{A}$ is the set of possible QAM values, in a tractable amount of time. Reliability information can be approximated [4], which is ignored in this work due to complexity reasons.

IV. Noise Interpolation

After a linear data estimation using the LMMSE in (6), data estimates $\hat{\mathbf{d}} = \mathbf{E}_{\mathrm{LMMSE}}\mathbf{y}$ are obtained along with statistics of the remaining error $\mathbf{e} = \hat{\mathbf{d}} - \mathbf{d}$, which will be treated as noise from here on. In general, the noise covariance matrix $\mathbf{C}_{ee}$ in (7) has several significant entries aside from the main diagonal, suggesting that the noise samples are correlated. In this section, we apply Wiener filtering by exploiting the correlations inherent in (the unknown) $\mathbf{e}$ in order to obtain a good estimate $\hat{\mathbf{e}}$ of the noise vector.

A. Full Range Noise Interpolation

We start with the assumption that only the k-th noise value e_k is unknown. Then $\hat{e}_k$ could be determined by a Wiener interpolation using all values of $\mathbf{e}$, but the k-th. According to Wiener theory (see e.g. [16]), an estimate of the error value is obtained by

$$\hat{e}_k = \mathbf{w}_k'^{\mathsf{H}}\bar{\mathbf{e}}_k, \tag{9}$$

$$\mathbf{w}_k'^{\mathsf{H}} = \bar{\mathbf{c}}_k^{\mathsf{H}}\bar{\mathbf{C}}_k^{-1}, \tag{10}$$

where $\bar{\mathbf{e}}_k$ is the noise vector excluding the k-th value, and the k-th estimator $\mathbf{w}_k'^{\mathsf{H}}$ is determined by excerpts of the $\mathbf{C}_{ee}$ matrix: The vector $\bar{\mathbf{c}}_k^{\mathsf{H}}$ is the k-th row of $\mathbf{C}_{ee}$ excluding value number k.

Respectively, $\bar{\mathbf{C}}_k$ excludes row and column number k of $\mathbf{C}_{ee}$, diminishing it to size $(N_d - 1) \times (N_d - 1)$.

Clearly, in our system the assumption of knowing all e_l for $l \neq k$ does not hold. However, the noise estimates

$$\tilde{e}_l = \hat{d}_l - \lfloor \hat{d}_l \rceil, \tag{11}$$

where $\lfloor \hat{d}_l \rceil$ refers to the slicing operation to the nearest valid data symbol, can be used instead of the true values e_l.

After determination of the estimators $\mathbf{w}_k'^{\mathsf{H}}$ for all $k = 0, 1, \ldots, N_d - 1$ and producing full N_d-size estimator vectors $\mathbf{w}_k^{\mathsf{H}}$ by inserting a zero at the k-th position $\mathbf{w}_k^{\mathsf{H}} = [w'_{k,0} \ldots, w'_{k,k-1}, 0, w'_{k,k}, \ldots, w'_{k,N_d-2}]$, we can assemble the estimator matrix $\mathbf{W} = [\,\mathbf{w}_0 \quad \cdots \quad \mathbf{w}_{N_d-1}\,]^{\mathsf{H}}$ that allows to write the noise interpolation of all samples as

$$\hat{\mathbf{e}} = \mathbf{W}\tilde{\mathbf{e}}, \tag{12}$$

where $\tilde{\mathbf{e}} = [\,\tilde{e}_0, \tilde{e}_1, \cdots, \tilde{e}_{N_d-1}\,]^{\top}$, $\hat{\mathbf{e}} = [\,\hat{e}_0, \hat{e}_1, \cdots, \hat{e}_{N_d-1}\,]^{\top}$. In a last step these noise estimates are subtracted from the linear data estimates in order to obtain improved ones:

$$\hat{\mathbf{d}}^{\mathrm{NI}} = \hat{\mathbf{d}} - \hat{\mathbf{e}}. \tag{13}$$

Furthermore, the noise covariance matrix after noise interpolation providing reliability information for soft decision channel decoding can be approximated as (cf. [16])

$$\mathbf{C}_{ee}^{\mathrm{NI}} = (\mathbf{I} - \mathbf{W})\mathbf{C}_{ee}. \tag{14}$$

Let us discuss the concept described so far: An element $\tilde{e}_l$ of vector $\tilde{\mathbf{e}}$ equals the true value e_l, if the corresponding slicing operation yields the correct data symbol. If $\tilde{e}_l = e_l$ for all $l = 0, 1, \ldots, N_d - 1$, noise interpolation would not be necessary at all, as the LMMSE data estimates would all lie in the correct decision region. On the other hand, a slicing mistake to the wrong data symbol in (11) for a single index l produces a wrong noise estimate $\tilde{e}_l \neq e_l$, which has an impact on every $\hat{e}_k$ for $k \neq l$, constituting an error propagation. Let's assume the slicing operation is correct for all $k \in \mathcal{S}$, and wrong for all $k \in \mathcal{U}$. The perfect approach would be to set $\hat{e}_k = \tilde{e}_k \,(= e_k)$ for $k \in \mathcal{S}$, to derive all $\hat{e}_k$ for $k \in \mathcal{U}$ by the Wiener interpolation process, and to only use $\tilde{e}_k$ with $k \in \mathcal{S}$ as input for the interpolation process. Then we would avoid error propagation, and all $\hat{e}_k$ for $k \in \mathcal{S}$ would definitely be correct. However, in practice we do not know $\mathcal{S}$ and $\mathcal{U}$. In the following subsection we try to come close to this optimum approach by at least performing a careful selection of the samples to be used as input for the interpolation process.

B. Sample Selection

As a preparation step we define $\mathcal{S}_k$ to be the set containing the indexes corresponding to the noise estimates $\tilde{e}_l$ to be used for the interpolation of the k-th noise sample. The noise estimate $\hat{e}_k$ is then found by

$$\hat{e}_k = \mathbf{w}_k'\tilde{\mathbf{e}}_{\mathcal{S}_k}, \tag{15}$$

$$\mathbf{w}_k'^{\mathsf{H}} = \mathbf{c}_{\mathcal{S}_k}^{\mathsf{H}}\mathbf{C}_{\mathcal{S}_k}^{-1}. \tag{16}$$

Analog to the previous approach, excerpts of the LMMSE error covariance matrix are used to determine the estimator: $\mathbf{c}_{\mathcal{S}_k}^{\mathsf{H}}$ is the k-th row of $\mathbf{C}_{ee}$ with only the columns indicated by the indexes in $\mathcal{S}_k$, and $\mathbf{C}_{\mathcal{S}_k}$ having only the columns and rows of $\mathbf{C}_{ee}$ with indexes from $\mathcal{S}_k$. Also a diminished noise vector $\tilde{\mathbf{e}}_{\mathcal{S}_k}$ with only the values indicated in $\mathcal{S}_k$ is needed. Consequently, the estimator $\mathbf{w}_k'$ has a length corresponding to the cardinality of $\mathcal{S}_k$. Introducing zeros at the positions corresponding to noise samples that are not used for the interpolation process yields the length N_d estimation vectors $\mathbf{w}_k$. With these, $\mathbf{W}$ is built to be used as in (12) and (14).

Different strategies are possible to define reasonable sets $\mathcal{S}_k$ (for $k = 0, 1, \cdots, N_d - 1$). A good way would be the use of a-posteriori probabilities of wrong slicing decisions for all elements of $\hat{\mathbf{d}}$. This method, however, is very costly, as it incorporates the actual linear estimates $\hat{d}_k$ in the probability calculation. In this work we therefore suggest the a-priori probabilities as a reasonable compromise. The a-priori symbol error probability (SEP) for data symbol number k in 4-QAM transmission is given by [5]

$$\chi_k = \Pr\left(\lfloor \hat{d}_k \rceil \neq d_k\right) = 2\mathrm{Q}\left(\sqrt{\frac{\sigma_d^2}{\sigma_{e_k}^2}}\right), \tag{17}$$

where σ_d^2 is the variance of the data symbols, $\sigma_{e_k}^2$ is the variance of the (complex) remaining error, apparent at the k-th position of the main diagonal of $\mathbf{C}_{ee}$ and $\mathrm{Q}(x) = \frac{1}{2\pi}\int_x^{\infty} \mathrm{e}^{-t^2/2}\mathrm{d}t$.

We introduce a threshold θ for the a-priori symbol error probability and only select noise samples for NI that do not exceed this threshold, such that the set of samples to interpolate $\hat{e}_k$ from is given by

$$\mathcal{S}_k = \{l | \chi_l < \theta, l = \{0, \ldots, N_d - 1\} \backslash k\}. \tag{18}$$

The actual threshold has been determined rather empirically for the UW-OFDM setup described in Section V. To choose θ appropriately, extensive BER simulations have been performed in the multipath environment also detailed in Section V. In Fig. 1, the BERs are displayed as they are achieved with a certain SEP threshold at a given E_b/N_0. The BER is plotted in relation to the minimum BER achieved for the current E_b/N_0, such that the blue valley in Fig. 1 determines the optimum which can be reached by setting θ appropriately in dependence of E_b/N_0. Fitting a straight line of the form $\log_{10}(\theta) = m \cdot E_b/N_0 + b$ yields the rather empirically determined threshold

$$\theta = 10^{-0.0037 E_b/N_0 - 1.3}, \tag{19}$$

which reaches the minimum BER in good approximation. This threshold, desgined for the UW-OFDM setup at hand, is visible in red in the plot.

V. Simulation Results

In order to quantify its performance, the NI concept is compared with the conventional LMMSE estimator with and without channel coding. For encoding we use the industry standard convolutional code with generator polynomials (133, 171), punctured for the coding rate $R = 3/4$ according to

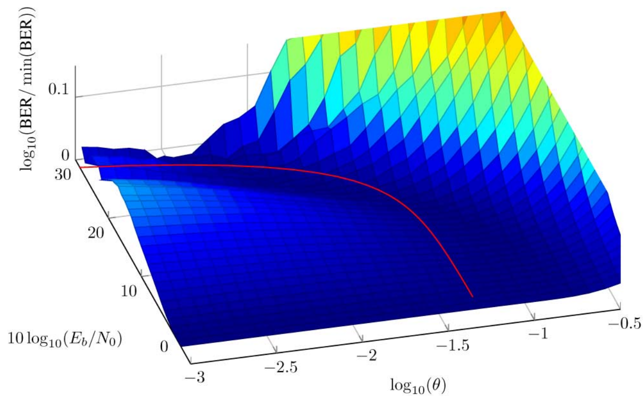

Fig. 1. BER relative to the minimum BER achieved depending on the E_b/N_0-ratio with a given threshold (all measures in logarithmic scale).

TABLE I
PARAMETERS OF THE INVESTIGATED UW-OFDM SYSTEM

Modulation scheme		4-QAM
Coding rates	R	uncoded, 3/4
DFT length	N	64
No. of data (redundant) subcarriers	N_d (N_r)	48 (16)
Indexes of red. subcarriers		$\{1, 5, \ldots, 61\}$
Unique Word	$\mathbf{x}_u$	$0^{(16\times 1)}$
Sampling time	T_s	50 ns

[17]. For decoding a soft decision Viterbi decoder is applied. As a reference for the uncoded transmission we also include results of the SD. For the chosen UW-OFDM system parameters, the determination of reliability information for the SD is computationally too complex to deliver results in feasible time. Hence, no coded results are shown for this estimator. The most important system parameters are specified in Table I. Since we focus on data estimation in this work rather than approaches to make use of the UW, we chose the zero UW for the BER simulations below. The use of a non-zero UW would cause a right shift of the presented curves, resulting from the increased transmit energy compared to the zero UW case.

The multipath channel is modeled as a tapped delay line with an exponentially decaying power profile, as described in [18]. We stored 10 000 realizations of channel impulse responses, featuring (on average) a delay spread of 100 ns and a total length not exceeding the guard interval. Furthermore, the channel impulse responses have been normalized such that the receive power is independent of the actual channel. Perfect channel knowledge is assumed at the receiver.

In Fig. 2, the simulation results of the uncoded transmission are displayed for the AWGN channel and for an indoor multipath environment, both for systematic and non-systematic UW-OFDM. In all performance comparisons the dB-measurements are taken at a BER of 10^{-5}. As discovered in [14] and apparent in Fig. 2 all receivers perform equally for non-systematically generated UW-OFDM in the AWGN channel. For systematic UW-OFDM in the AWGN channel, the proposed NI improves the LMMSE estimator by 1 dB, while the optimum but complex SD is only 0.7 dB in front of the NI. In the multipath environment the performance differences are significant. For systematic UW-OFDM the NI is 1.5 dB ahead of the LMMSE estimator, and 3.3 dB behind the SD performance. For the non-systematic approach the NI performs even better: It supersedes the LMMSE estimator by impressive 4.5 dB, leaving 2.6 dB to the SD.

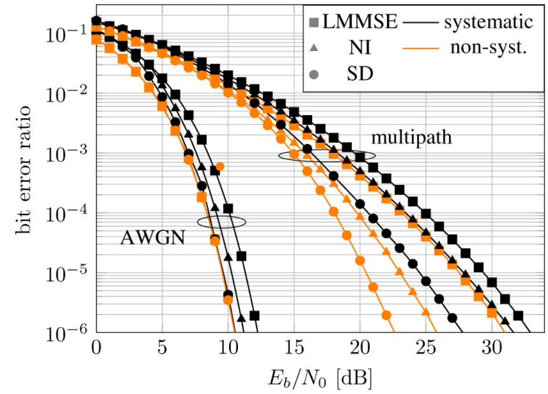

Fig. 2. Simulated BER for uncoded transmission: LMMSE estimator, SD and NI receiver in AWGN and multipath environment.

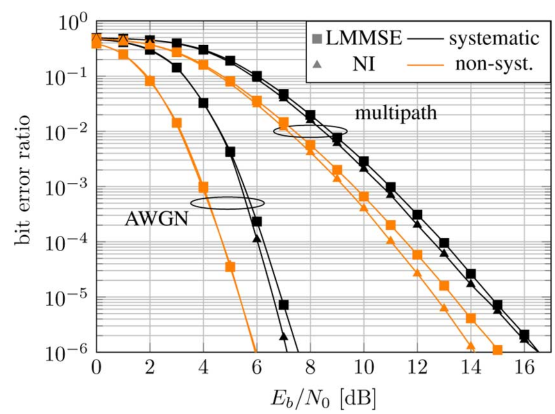

Fig. 3. Simulated BER for coded transmission, $R = 3/4$: LMMSE estimator and NI receiver in AWGN and multipath environment.

The results for coded transmission are shown in Fig. 3. In the AWGN channel, the gain of the NI receiver over the LMMSE estimator is limited. However, in the frequency selective environment, a considerable gain of 0.7 dB is achieved when applying the favorable non-systematic approach. Further investigations for varying coding rates revealed a shrinking gain of the NI receiver over the LMMSE estimator with decreasing R.

VI. CONCLUSION

The rather complex sphere decoder is known to outperform the simple LMMSE estimator in UW-OFDM communications. In this work, an alternative but less complex non-linear data estimator is introduced for UW-OFDM that shows analogies to the well known noise prediction and noise whitening approaches in single carrier systems. After an initial LMMSE estimation, the data estimation performance is improved by exploiting the correlations inherent in the remaining error. This is achieved by applying Wiener noise interpolation. In this approach the selection of the samples involved in the interpolation process is crucial. A simple sample selection criterion has been introduced in an empirical way. From our simulation results we conclude that the NI concept shows its strengths in frequency selective channels, and it is especially beneficial in combination with high coding rates.

REFERENCES

[1] M. Huemer, C. Hofbauer, and J. B. Huber, "The potential of unique words in OFDM," in *Proc. 15th Int. OFDM Workshop*, Hamburg, Germany, Sep. 2010, pp. 140–144.

[2] M. Huemer, A. Onic, and C. Hofbauer, "Classical and bayesian linear data estimators for unique word OFDM," *IEEE Trans. Signal Process.*, vol. 59, no. 12, pp. 6073–6085, Dec. 2011.

[3] A. Onic and M. Huemer, "Sphere decoding for unique word OFDM," in *Proc. IEEE Global Telecommunications Conf. (GLOBECOM)*, Houston, TX, USA, Dec. 2011.

[4] A. Onic, A. Schenk, M. Huemer, and J. B. Huber, "Soft-output sphere detection for coded unique word OFDM," in *Conf. Rec. 46th Asilomar Conf. Signals, Systems and Computers*, Asilomar, CA, USA, Nov. 2012, pp. 138–142.

[5] J. Proakis, *Digital Communications*, 4th ed. ed. New York, NY, USA: McGraw-Hill, Aug. 2000.

[6] Y. Zhu and K. B. Letaief, "Single carrier frequency domain equalization with time domain noise prediction for wideband wireless communications," *IEEE Trans. Wireless Commun.*, vol. 5, no. 12, pp. 3548–3557, Dec. 2006.

[7] F. Pancaldi, G. M. Vitetta, R. Kalbasi, N. Al-Dhahir, M. Uysal, and H. Mheidat, "Single-carrier frequency domain equalization," *IEEE Signal Process. Mag.*, vol. 25, no. 5, pp. 37–56, Sep. 2008.

[8] A. Onic, "Receiver Concepts for Unique Word OFDM," Ph.D. dissertation, Inst. Networked and Embedded Systems, Alpen-Adria-Univ. Klagenfurt, Klagenfurt, Germany, Nov. 2013.

[9] A. Onic and M. Huemer, "Direct vs. two-step approach for unique word generation in UW-OFDM," in *Proc. 15th Int. OFDM Workshop*, Hamburg, Germany, Sep. 2010, pp. 145–149.

[10] H. Steendam, "On the redundant carrier distribution for UW-OFDM," in *Proc. IEEE Wireless Communications and Networking Conf. (WCNC)*, Paris, France, Apr. 2012, pp. 32–36.

[11] H. Steendam, "Analysis of the redundant energy in UW-OFDM," *IEEE Trans. Commun.*, vol. 60, no. 6, pp. 1692–1701, Jun. 2012.

[12] H. Steendam, "The quasi-uniform redundant carrier placement for UW-OFDM," in *IEEE Vehicular Technology Conf. (VTC Fall)*, Sep. 2012.

[13] H. Steendam, "On the selection of the redundant carrier positions in UW-OFDM," *IEEE Trans. Signal Process.*, vol. 61, no. 5, pp. 1112–1120, Mar. 2013.

[14] M. Huemer, C. Hofbauer, A. Onic, and J. B. Huber, "On the exploitation of the redundant energy in UW-OFDM: LMMSE versus sphere detection," *IEEE Signal Process. Lett.*, vol. 19, no. 6, pp. 340–343, Jun. 2012.

[15] M. Huemer, C. Hofbauer, and J. B. Huber, "Non-systematic complex number RS coded OFDM by unique word prefix," *IEEE Trans. Signal Process.*, vol. 60, no. 1, pp. 285–299, Jan. 2012.

[16] S. M. Kay, *Fundamentals of Statistical Signal Processing, Volume I: Estimation Theory*, 1st ed. ed. Upper Saddle River, NJ, USA: Prentice-Hall, Apr. 1993.

[17] "IEEE Std 802.11a-1999, Part 11: Wireless LAN Medium Access Control (MAC) and Physical Layer (PHY) specifications: High-Speed Physical Layer in the 5 GHz Band," IEEE, 1999.

[18] J. Fakatselis, Criteria for 2.4 GHz PHY Comparison of Modulation IEEE Doc. p802.11-97/157r1, 1997.