Intercarrier Interference Suppression for OFDM Transmission at Very High Velocities

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***Abstract*—In transmission scenarios where the transmitter and/or receiver move at very high velocities, the performance of an OFDM-based transmission system can severely suffer from the effects of Doppler. In the presented paper, we therefore propose the application of a low-complexity sorted QR decomposition of the channel interference matrix in order to suppress the intercarrier interference resulting from high Doppler spreads. The hereby achievable system performance improvements compared to existing MMSE and MMSE-DFE approaches are shown in our simulation results.**

***Index Terms*— OFDM transmission, intercarrier interference suppression, Doppler spread, QR decomposition**

I. Introduction

The orthogonal frequency division multiplexing (OFDM) is a very popular transmission technique for frequency selective multipath channels due to the significantly simplified equalization process, compared to an equivalent single carrier system in the same environment. Provided that the delay spread of the channel does not exceed the length of the cyclic prefix (CP) the effects of intersymbol interference (ISI) are completely suppressed. However, in scenarios where the transmitter and/or receiver move at a very high velocity, e.g. in case of high-speed trains or airplanes, OFDM shows its sensitivity to the Doppler effect: Due to the shifting and widening of the spectrum of each subcarrier by the so called Doppler spread, the orthogonality of the subchannels is violated. This leads to intercarrier interference (ICI) which can strongly degrade the overall system performance [1], [2].

The shifting of the subcarrier spectra can be corrected by one of the numerous carrier frequency offset (CFO) estimation and correction algorithms found in the literature, e.g. [3], [4], and therefore is not covered in this paper. Instead, we focus on the suppression of ICI resulting from the widening of the spectrum. For the following considerations we assume that the channel coefficients are perfectly estimated at each time. On that condition a straightforward solution would be a minimum mean square error (MMSE) equalization of all subcarriers at the same time. The drawback of this method is the large computational effort due to the inversion of the channel interference matrix whose size increases quadratically with the number of subcarriers. In [2] an approach with lower complexity is proposed, where the channel interference matrix is decomposed into submatrices which are inverted recursively.

In order to improve the BER while still keeping the complexity low, an MMSE decision-feedback equalizer (DFE) approach is presented in [2], which is also based on a recursive submatrix inversion. The performance of this method could be strongly improved if the decision order was sorted on the basis of the SINR for each subcarrier. In [5], a similar problem, the layer detection of a BLAST architecture, was efficiently solved by performing a sorted QR decomposition (SQRD) of the multiple input multiple output (MIMO) channel matrix with respect to the MMSE criterion. In the presented paper, we therefore study the application of this successive detection structure for the ICI cancellation in an OFDM system and demonstrate its performance gain over the low-complexity MMSE and MMSE-DFE techniques presented in [2]. Additionally, we derive a novel approach performing a blockwise MMSE-SQRD of the channel interference matrix in order to reduce the computational complexity significantly.

The paper is organized as follows: Section II deals with the description of the OFDM system and section III with ICI suppression by recursive MMSE and MMSE-DFE solutions. In section IV we present our novel approach of ICI cancellation based on the MMSE-SQRD of the channel interference matrix. Section V comprises the simulation results for the comparison of the algorithms, followed by a conclusion of the paper in section VI.

II. System Description including Doppler Effects

We consider a conventional OFDM system with N subcarriers, an OFDM core symbol duration T_s, a subcarrier spacing of $\Delta f = 1/T_s$, and a sampling frequency of $f_s = N\Delta f$. The guard interval (GI) of length N_g/f_s is dimensioned larger than the maximum channel delay τ_{max} in order to avoid ISI. Applying the $(N \times N)$ discrete Fourier transform (DFT) matrix $\mathbf{F}$ containing the elements[1] $\mathbf{F}(\mu,\nu) := 1/\sqrt{N} \cdot \exp(-j2\pi\mu\nu/N)$, and defining a vector comprising the source symbols in frequency domain $\mathbf{d} := [d_0, ..., d_{N-1}]^T$, the time discrete signal at the transmitter output $\mathbf{s} := [s(N-N_g), ..., s(N-1), s(0), ..., s(N-1)]^T$ can be written as

$$\mathbf{s} = \mathbf{T}_g \mathbf{F}^H \mathbf{d}\,. \tag{1}$$

The matrix $\mathbf{T}_g = [\mathbf{I}_g^T, \mathbf{I}_N]^T$, with $\mathbf{I}_g$ as the last N_g rows of the $(N \times N)$ identity matrix $\mathbf{I}_N$, accomplishes the insertion of the GI.

The channel is assumed to have a wide-sense-stationary uncorrelated-scattering (WSSUS) characteristic. The time over which the channel can be supposed as time-invariant is called coherence time $t_c = 1/f_{D,max}$. It directly depends on the maximum Doppler frequency $f_{D,max} = v_0 f_c / c_0$, where v_0 denotes the relative velocity between transmitter and receiver, f_c the carrier frequency, and c_0 the speed of light.

In order to obtain the frequency response at the nth subchannel for the time k/f_s, the discrete channel coefficients $h(k,\kappa)$, with time index $k = tf_s$ and delay index $\kappa = \tau f_s$, are transformed into frequency domain:

$$\tilde{h}_n(k) = \sqrt{N}\mathbf{F}(n, 0\!:\!L)\mathbf{h}(k) \tag{2}$$

with $\mathbf{h}(k) := [h(k,0), ..., h(k,L)]^T$, and $L = \lfloor \tau_{max} f_s \rfloor + 1$ as the order of the channel impulse response. Defining $\tilde{\mathbf{h}}_n := [\tilde{h}_n(0), ..., \tilde{h}_n(N-1)]^T$ we can formulate the impulse response of the nth subchannel with $\mathrm{diag}[\mathbf{F}(n,:)^H]\tilde{\mathbf{h}}_n$ as well as the receive signal after removing the GI:

$$\mathbf{r} = \sum_{n=0}^{N-1} \mathrm{diag}\left[\mathbf{F}(n,:)^H\right] \tilde{\mathbf{h}}_n d_n + \mathbf{n}\,, \tag{3}$$

where $\mathbf{r} = [r(0), ..., r(N-1)]^T$, while $\mathbf{n} = [n(0), ..., n(N-1)]^T$ denotes additive white Gaussian noise (AWGN) with a variance of σ_n^2. After performing the DFT we obtain

$$\mathbf{x} = \mathbf{F}\mathbf{r} = \mathbf{H}\mathbf{d} + \tilde{\mathbf{n}}\,, \tag{4}$$

with $\mathbf{x} := [x_0, ..., x_{N-1}]^T$. The channel interference matrix $\mathbf{H}$ comprises the elements

$$\mathbf{H}(m,n) := \mathbf{F}(:,m)^T \mathrm{diag}\left[\mathbf{F}(n,:)^H\right] \tilde{\mathbf{h}}_n \tag{5}$$

and can be interpreted as follows: The elements in row m represent the signal components of each subcarrier received on the mth subchannel. Consequently, column n defines how the signal components of the nth subcarrier are spread over all the subchannels. Thus, if $\mathbf{H}$ is not a strict diagonal matrix the receive signal contains ICI, which is the case if the channel behaves time-variant within the period of one OFDM symbol, i.e. $t_c < T_s + T_g$.

III. ICI Suppression based on MMSE and Decision Feedback Approaches

The MMSE solution for the equalization of the receive vector in (4) can be written as

$$\mathbf{y} = (\sigma_s^2 \mathbf{H}^H \mathbf{H} + \sigma_n^2 \mathbf{I}_N)^{-1} \mathbf{H}^H \mathbf{x}\,, \tag{6}$$

with $\sigma_s^2 = \mathrm{E}\{|d_n|^2\}$. In case of a quadrature amplitude modulation (QAM) scheme an additional bias correction

$$\hat{\mathbf{y}} = \mathrm{diag}\left[(\mathbf{H}^H\mathbf{H} + \sigma_n^2 \mathbf{I}_N)(\mathbf{H}^H\mathbf{H})^{-1}\right] \mathbf{y} \tag{7}$$

becomes necessary in order to obtain the proper symbol magnitudes. Depending on the number of subcarriers the matrix inversion in (6) can become very complex leading to a high computational effort. Exploiting the fact that in case of Doppler spread the major symbol energy of a subcarrier is distributed over the actual as well as a few adjacent subchannels only, i.e. the channel matrix $\mathbf{H}$ shows a strongly diagonal characteristic, offers several ways of reducing the equalization effort.

[1] Applied notation: $\mathbf{X}(a\!:\!b, c\!:\!d)$ means a submatrix containing the rows a to b and the columns c to d of the matrix $\mathbf{X}$.

In [2] two such approaches are presented. The first one equalizes each subcarrier separately, based on the evaluation of the next Q subcarriers in each frequency direction, i.e. the total number of considered subcarriers in each step is $K = 2Q+1$. The starting point is the creation of a $(K \times 1)$ index vector $\boldsymbol{\rho}_n$ containing the elements $\boldsymbol{\rho}_n(\mu) := (n - Q + \mu) \bmod N,\ \mu = 0, ..., K-1$. With $\mathbf{x}_n := \mathbf{x}(\boldsymbol{\rho}_n)$, $\mathbf{H}_n := \mathbf{H}(\boldsymbol{\rho}_n, :)$, and $\tilde{\mathbf{n}}_n := \tilde{\mathbf{n}}(\boldsymbol{\rho}_n)$ we can rewrite (4) as

$$\mathbf{x}_n = \mathbf{H}_n \mathbf{d} + \tilde{\mathbf{n}}_n \,. \tag{8}$$

On the basis of (8) the MMSE solution for the equalization of the nth subcarrier is

$$y_n = \left((\sigma_s^2 \mathbf{H}_n \mathbf{H}_n^H + \sigma_n^2 \mathbf{I}_K)^{-1} \mathbf{H}_n(:, n)\right)^H \mathbf{x}_n \,. \tag{9}$$

As the equalization has to be performed for each subcarrier separately, the inverse of the $(K \times K)$ covariance matrix $\mathbf{R}_n = (\sigma_s^2 \mathbf{H}_n \mathbf{H}_n^H + \sigma_n^2 \mathbf{I}_K)$ from within the right hand side expression of (9) has to be calculated N times. The complexity of this operation depends on the number of involved subchannels, thus the value of K represents a tradeoff between accuracy and computational effort. Due to the fact that the matrices $\mathbf{R}_n$ and $\mathbf{R}_{n+1}$ only differ by removing the first row and appending a new one at the end, the inverse of $\mathbf{R}_{n+1}$ can be recursively calculated from the inverse of $\mathbf{R}_n$, reducing the total computational effort for detecting one OFDM symbol from $\mathcal{O}(N^3)$, as required for (6), to $\mathcal{O}(N^2 K)$ operations [2].

An alternative approach to the linear MMSE detection is the application of a decision-feedback equalizer exploiting the finite symbol alphabet in order to improve the equalization accuracy. After selecting a subcarrier index ξ to start with, e.g. by detecting the column of $\mathbf{H}$ offering the largest norm, the corresponding subcarrier x_ξ can be equalized according to the MMSE solution in (9). Then, the remaining subcarriers are equalized successively in the order given by the index vector $\boldsymbol{\varrho}$ containing the elements $\boldsymbol{\varrho}(\mu) := [(\xi + \mu) \bmod N];\ \mu = 0, ..., N-1$. In order to detect the currently processed subcarrier, the symbols of the previously equalized subcarriers are decided and their reconstructed signal components subtracted from the receive vector $\mathbf{x}_n$ defined in (8):

$$\hat{\mathbf{x}}_n := \mathbf{x}_n - \sum_{\mu=0}^{\nu_n} \mathbf{H}_n(:, \boldsymbol{\varrho}(\mu)) \hat{d}_{\boldsymbol{\varrho}(\mu)} \,, \tag{10}$$

where the symbol $\hat{d}_n$ represents the decision of y_n, $\nu_n = (n - \xi - 1) \bmod N$, and $n = \boldsymbol{\varrho}(1), ..., \boldsymbol{\varrho}(N-1)$. Provided that all of the previous decisions are error-free, another expression for $\hat{\mathbf{x}}_n$ can be formulated based on equation (8):

$$\hat{\mathbf{x}}_n = \tilde{\mathbf{H}}_n \tilde{\mathbf{d}} + \tilde{\mathbf{n}}_n \,, \tag{11}$$

with $\tilde{\mathbf{H}}_n = \mathbf{H}_n(:, \boldsymbol{\varrho}(\nu_n + 1 : N - 1))$, and $\tilde{\mathbf{d}} := \mathbf{d}(\boldsymbol{\varrho}(\nu_n + 1 : N - 1))$. This equation can be solved separately for each subcarrier in an analogous manner to the MMSE solution in (9). Thereby, the inverse of the covariance matrix again can be recursively calculated in order to reduce the computational effort. The complexity of this MMSE-DFE approach is comparable with the before presented recursive MMSE equalization [2].

IV. ICI Cancellation by Sorted QR Decomposition

The major drawback of the MMSE-DFE approach can be seen in the processing order of the subcarriers, because that does not take into account the reliability of a symbol decision, which directly depends on the signal-to-interference-and-noise ratio (SINR) of the corresponding subcarrier. In order to avoid wrong decisions as well as resultant error propagation, it is important to process the most reliable subcarriers first. This problem is very similar to the layer detection in multiple antenna systems based on the well-known V-BLAST architecture [6]. The generally applied successive interference cancellation technique searches for the strongest layer, detects it, and subtracts its interference from the receive signal before it searches for the strongest of the remaining layers which is processed next. For this purpose, a very efficient algorithm called MMSE-SQRD was presented in [5]. In the following, we will adapt this technique to the ICI cancellation in high Doppler environments and furthermore derive a low-complexity version exploiting the strong diagonal characteristic of the channel matrix $\mathbf{H}$.

Defining a $(2N \times N)$ extended channel matrix $\underline{\mathbf{H}} := \left[\mathbf{H}^T\ \sigma_n \mathbf{I}_N\right]^T$ as well as an extended receive vector $\underline{\mathbf{x}} := \left[\mathbf{x}^T\ \mathbf{0}_{1,N}\right]^T$ the MMSE solution in (6) can be rewritten as

$$\mathbf{y} = (\sigma_s^2 \underline{\mathbf{H}}^H \underline{\mathbf{H}})^{-1} \underline{\mathbf{H}}^H \underline{\mathbf{x}} \,. \tag{12}$$

Performing a QR decomposition of $\underline{\mathbf{H}}$ we obtain

$$\underline{\mathbf{H}} = \begin{bmatrix} \mathbf{H} \\ \sigma_n \mathbf{I}_N \end{bmatrix} = \underline{\mathbf{Q}}\underline{\mathbf{R}} = \begin{bmatrix} \mathbf{Q}_1 \\ \mathbf{Q}_2 \end{bmatrix} \underline{\mathbf{R}} \,, \tag{13}$$

where the unitary matrix $\underline{\mathbf{Q}}$ is partitioned into the two $(N \times N)$ matrices $\mathbf{Q}_1$ and $\mathbf{Q}_2$, while $\underline{\mathbf{R}}$ denotes an $(N \times N)$ upper triangular matrix. From (13) directly follows $\underline{\mathbf{Q}}^H \underline{\mathbf{H}} = \mathbf{Q}_1^H \mathbf{H} + \sigma_n \mathbf{Q}_2^H = \underline{\mathbf{R}}$ and $\sigma_n \mathbf{I}_N = \mathbf{Q}_2 \underline{\mathbf{R}}$, i.e. $\underline{\mathbf{R}}^{-1} = \mathbf{Q}_2/\sigma_n$. Together with (4), this leads to the expression for the filtered receive vector

$$\tilde{\mathbf{x}} := \underline{\mathbf{Q}}^H \underline{\mathbf{x}} = \mathbf{Q}_1^H \mathbf{x} = \underline{\mathbf{R}}\mathbf{d} - \sigma_n \mathbf{Q}_2^H \mathbf{d} + \mathbf{Q}_1^H \tilde{\mathbf{n}} \quad (14)$$

with $\tilde{\mathbf{x}} = [\tilde{x}_0, ..., \tilde{x}_{N-1}]^T$. Neglecting the remaining interference $-\sigma_n \mathbf{Q}_2^H \mathbf{d}$ as well as the noise term $\mathbf{Q}_1^H \tilde{\mathbf{n}}$, the nth element of $\tilde{\mathbf{x}}$ can be written as

$$\tilde{x}_n = \underline{\mathbf{R}}(n,n) d_n + \sum_{\mu=n+1}^{N} \underline{\mathbf{R}}(n,\mu) d_\mu \,, \quad (15)$$

because of the triangular structure of $\underline{\mathbf{R}}$. Consequently, starting from the last row, the subcarriers can be equalized successively by subtracting the interference from the previously decided ones:

$$y_n = 1/\underline{\mathbf{R}}(n,n) \left(\tilde{x}_n - \sum_{\mu=n+1}^{N} \underline{\mathbf{R}}(n,\mu) \hat{d}_\mu \right) , \quad (16)$$

where $\hat{d}_n$ denotes the symbol decision based on y_n.

As already mentioned, the processing order of the subcarriers is crucial for a successful interference cancellation. Therefore, prior to each orthogonalization step, i.e. calculating the next column of $\underline{\mathbf{Q}}$ and next row of $\underline{\mathbf{R}}$ in (13), the columns of $\underline{\mathbf{H}}$ are permutated in such a way, that the magnitude of the diagonal elements $\underline{\mathbf{R}}(n', n')$ increases with n' [5]. This procedure aims at maximizing the SINR of the equalized subcarriers belonging to the bottom rows of $\underline{\mathbf{R}}$, which are processed first in (16). However, the MMSE-SQRD approach does not always lead to the optimum detection order. This can only be assured by applying a subsequent post-sorting algorithm, also presented in [5], but at the cost of a slightly higher computational effort. The complexity of the MMSE-SQRD with $\mathcal{O}(N^3)$ per OFDM symbol is comparable to the conventional MMSE solution in (6).

As in the last section, the strong diagonal characteristic of the channel matrix $\mathbf{H}$ motivates a reduction of the computational complexity. In the following, we present a novel approach where the matrix $\mathbf{H}$ is divided into J submatrices of equal size, so that the ICI can be cancelled blockwise by a modified MMSE-SQRD. Let $P := N/J$ be the number of subcarriers to be equalized in each block. This value obviously limits the number of possible subchannel permutations and, on the other hand, the size of the submatrices, i.e the computational effort, for each SQRD cycle. Thus, the parameter J represents a tradeoff between the accuracy and the complexity of the equalization.

The block to be processed in the first SQRD cycle can be determined by searching a set of P consecutive subchannels with the greatest average column norm in $\mathbf{H}$. With α_0 denoting the index of the first subcarrier of this block, the index vector $\boldsymbol{\lambda}_0 := [(\alpha_0 - Q) \mathrm{mod} N, ..., (\alpha_0 + P + Q - 1) \mathrm{mod} N]^T$ defines the corresponding $((P+2Q) \times (P+2Q))$ submatrix $\mathbf{H}_0 := \mathbf{H}(\boldsymbol{\lambda}_0, \boldsymbol{\lambda}_0)$. As in the last section, the parameter Q denotes the number of considered subcarriers to both sides of the processed block, in order to cancel their ICI especially at the subchannels in the border area of this block.

Analogous to (13), a sorted QR decomposition

$$\underline{\mathbf{H}_0} = \begin{bmatrix} \mathbf{H}_0 \\ \sigma_n \mathbf{I}_{P+2Q} \end{bmatrix} = \underline{\mathbf{Q}}_0 \underline{\mathbf{R}}_0 = \begin{bmatrix} \mathbf{Q}_{0,1} \\ \mathbf{Q}_{0,2} \end{bmatrix} \underline{\mathbf{R}}_0 \quad (17)$$

is performed, where the columns of $\mathbf{H}_0$ are permutated in such a way, that the diagonal elements $\underline{\mathbf{R}}_0(n,n)$ belonging to the first Q as well as the last Q subcarriers of the submatrix are arranged to the top of $\underline{\mathbf{R}}_0$ without a special ordering. The remaining P columns are sorted according to the full MMSE-SQRD approach and make up the P lower rows of $\underline{\mathbf{R}}_0$. This assures that by orthogonalization the corresponding subcarriers with indices $\boldsymbol{\lambda}'_0 := [\boldsymbol{\lambda}_0(Q+1), ..., \boldsymbol{\lambda}_0(Q+P)]^T$ are freed from ICI of the other ones. In the subsequent equalization process only those subcarriers are processed analogous to (16), i.e. the elements $y_{\boldsymbol{\lambda}'_0(1)}, ..., y_{\boldsymbol{\lambda}'_0(P)}$ detected. Finally, these are decided and their reconstructed signal components subtracted from the receive vector:

$$\hat{\mathbf{x}}_0 = \mathbf{x} - \hat{d}_{\boldsymbol{\lambda}'_0} \mathbf{H}(:, \boldsymbol{\lambda}'_0) \quad (18)$$

The second cycle is intended for equalizing the next block of P subcarriers starting with index $\alpha_1 = \alpha_0 + P$. Consequently, the vector $\boldsymbol{\lambda}_1 := [(\alpha_1 - Q) \mathrm{mod} N, ..., (\alpha_1 + P + Q - 1) \mathrm{mod} N]^T$ contains the indices of the considered rows of the channel matrix $\mathbf{H}$. Because the ICI from the preceding P subcarriers has already been cancelled from the receive vector, the corresponding columns of $\mathbf{H}$ can be neglected. Supposing $P \leq Q$, we obtain the $((P+2Q) \times (P+Q))$ submatrix $\mathbf{H}_1 := \mathbf{H}(\boldsymbol{\lambda}_1, \boldsymbol{\lambda}_1(P+1 : 2P+Q))$. While performing the sorted QR decomposition, the columns of $\mathbf{H}_1$ are again rearranged in such a way, that the rows of $\underline{\mathbf{R}}_1$ representing the last Q considered subcarriers can be found at the top of $\underline{\mathbf{R}}_1$, while the P subcarriers which are intended to be equalized within this cycle are sorted at the bottom. After their detection, the reconstructed signal is also subtracted from the receive vector. Thus, we obtain

$$\hat{\mathbf{x}}_1 = \hat{\mathbf{x}}_0 - \hat{d}_{\boldsymbol{\lambda}'_1} \mathbf{H}(:, \boldsymbol{\lambda}'_1)\,, \tag{19}$$

which can be evaluated in the subsequent cycle.

This procedure is repeated until all J blocks of subcarriers are equalized. The resulting computational effort can be rated with $\mathcal{O}\left(J(P + 2Q)^3\right)$. In case of an optimized adjustment of the parameters P and Q, this leads to an extreme complexity reduction compared to the $\mathcal{O}(N^3)$ operations of the full MMSE-SQRD approach.

V. Simulation Results

In this section, we present a performance comparison of the before introduced frequency domain equalization techniques for cancelling ICI caused by Doppler spread. Our simulated transmission scenario comprises an OFDM system with $N = 64$ subcarriers, a sampling frequency of $f_s = 20\,\mathrm{MHz}$, and a carrier frequency of $f_c = 60\,\mathrm{GHz}$. The channel coefficients were randomly created according to a 10-tap Rayleigh-fading channel model. The relative velocity between transmitter and receiver was set to $v_0 = 600\,\mathrm{m/s}$ resulting in a maximum Doppler frequency of $f_{D,max} = 120\,\mathrm{kHz}$. That corresponds to approx. 40% of the carrier spacing $\Delta f = 312.5\,\mathrm{kHz}$.

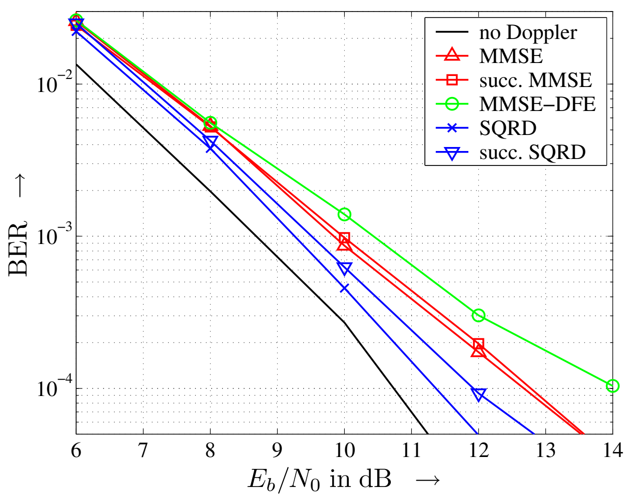

Fig. 1. Bit error rates: Comparison of ICI suppression techniques for QPSK-modulated subcarriers

The obtained bit error rates (BER) in case of QPSK-modulated subcarriers are presented in Fig. 1. At a BER of 10^{-4} the standard ('MMSE') as well as the low-complexity successive ('succ. MMSE') MMSE approaches show a loss of approx. 2 dB in E_b/N_0 compared to the Doppler-free case. The BER of the latter represents a lower bound that cannot be reached, even with the optimum performance of a maximum likelihood detector. Surprisingly, the MMSE-DFE technique performs 1 dB worse than the MMSE approaches. The cause of that loss is the suboptimum processing order of the subcarriers, which leads to wrong decisions and thereby provoked consecutive errors.

In contrast, the MMSE-SQRD algorithm with post-sorting ('SQRD') assures the optimum processing order, resulting in a performance close to the Doppler-free case and at a BER of 10^{-4} nearly 1.5 dB better than the MMSE approaches. For the cost of only 0.5 dB, the application of the successive MMSE-SQRD technique ('succ. SQRD'), with $J = 8$ blocks and $Q = 8$ additionally considered subcarriers to each side of a block, significantly reduces the complexity, comparable to the computational effort of the low-complexity MMSE approach.

VI. Conclusions

In the presented paper, we adapted the sorted QR decomposition for cancelling intercarrier interference caused by Doppler spread. We also derived a low-complexity version based on a successive processing of subblocks instead of equalizing a whole OFDM symbol at once. Both novel approaches show a superior performance over existing MMSE and DFE techniques, as we demonstrated in our simulation results.

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