JOINT TRANSMITTER / RECEIVER DESIGN FOR MULTICARRIER DATA TRANSMISSION WITH LOW LATENCY TIME

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ABSTRACT

In this paper a design method for low latency multicarrier transmission is presented. It can be considered as a generalization of the Trailing-Zeros Transmitter approach in [1]. The generalization mainly consists of using FIR redundant filter banks for the transmitter and receiver instead of pure block transforms and allowing to choose the guard interval independently of the channel impulse response length. Thanks to the latter, we can design a multicarrier transmission system with a low latency time, which is a critical parameter for online applications, even for the case that the channel has a long impulse response, as e.g. a twisted-pair copper wire line of several miles length. The design of the transmitter and receiver is based on a Smith decomposition of the channel. Advantages as well as limitations of the new algorithm are discussed.

1. INTRODUCTION

Multicarrier modulation finds its application in recently standardized high rate data transmission systems as e.g. Digital Audio Broadcasting (DAB), Digital Video Broadcasting (DVB-T), Wireless LAN (IEEE 802.11, HIPERLAN 2), Asymmetric Digital Subscriber Lines (ADSL), Very High Data Rate Digital Subscriber Lines (VDSL). Instead of transmitting one symbol after the other as in single-carrier data transmission, a block of M symbols is transmitted in parallel and each symbol is assigned $1/M$ of the available bandwidth. The most common algorithms are Orthogonal Frequency Division Multiplexing (ODFM) for wireless transmission and Digital Multi-Tone (DMT) Modulation for transmission over twisted pair copper wires. Both algorithms are based on Fast Fourier Transform (FFT). In order to obtain a simple equalizer at the receiver a so called guard interval of L samples is introduced at the end of each block of M symbols where L has to be at least as long as the channel impulse response. However, the introduction of the guard interval reduces the bandwidth efficiency by a factor of $M/(M + L)$. This is particularly severe, if the channel impulse response and thus the guard interval L is long compared to the block length M. Consequently, in high bitrate applications over short distances as ADSL or VDSL the block length M is chosen in the range from 256 to 2048 to maintain a reasonable bandwidth efficiency. This, on the other hand results in a large latency time of the transmission system due to the large block length.

For time invariant transmission channels the transmitter and receiver can be jointly optimized to obtain a higher data rate for a given channel. The guard interval here is replaced by the more general idea of the introduction of redundancy. This is very similar to the idea of channel coding where redundancy is introduced in order to reduce the bit error rate during transmission at the receiver. The following approaches have already been treated in literature:

For IIR transmitter and receiver filters Yang et al. propose an iterative minimum mean squared error (MMSE) transmitter / receiver optimization in [2] and a closed form MMSE solution is provided by Li et al. in [3]. For FIR joint transmitter / receiver optimization a zero-forcing solution as well as well as an MMSE solution are provided in [1]. In [4] a method using a now widely used cyclic extension is presented. However, in these approaches the number of redundant samples that is introduced has to be greater than the length of the channel impulse response. Thus, this approach cannot be applied when both high bandwidth efficiency as well as low latency time have to be satisfied for a transmission channel with long impulse response.

We here take the idea from [1] for joint FIR transmitter / receiver optimization and generalize it such that the amount of redundancy to be introduced is independent of the channel impulse response length.

The outline of the paper is as follows: In section 2 we describe the transmission system as a multiple-input multiple-output (MIMO) system. We then review the Trailing-Zeros Transmitter from [1] in section 3 and propose a generalization that allows low latency transmission also for long channel impulse responses in section 4. Finally, section 5 draws conclusions and shows limitations of the proposed algorithm.

2. DESCRIPTION OF THE TRANSMISSION SYSTEM

The general transmission scheme is shown in Figure 1.

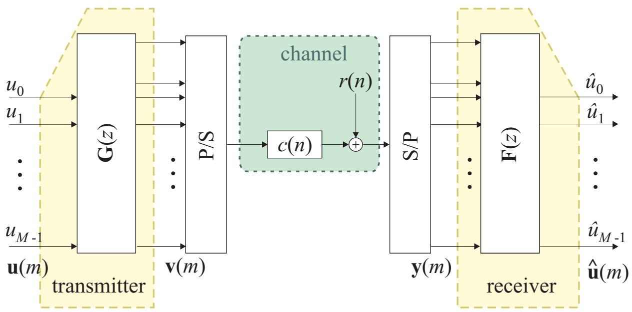

Figure 1: General MIMO transmission scheme with FIR transmitter $\mathbf{G}(z)$ and FIR receiver $\mathbf{F}(z)$

The transmitter is given by $\mathbf{G}(z)$ and transforms the vector $\mathbf{u}(m)$ of M input signals into a vector $\mathbf{v}(m)$ of $M+L$ output values. These are then parallel to serial transformed and transmitted over the channel which consists of the channel impulse response $c(n)$ as well as additive white gaussian noise $r(n)$. At the receiver the incoming data are serial to parallel transformed. The FIR receiver matrix $\mathbf{F}(z)$ then equalizes the incoming $P = M + L$ values in $\mathbf{y}(m)$ into the vector $\hat{\mathbf{u}}(m)$. In the case of perfect equalization, $\hat{\mathbf{u}}(m)$ is a delayed copy of the input vector $\mathbf{u}(m)$:

$$\hat{\mathbf{u}}(m) = \mathbf{u}(m - d) \tag{1}$$

where d describes the system delay in number of blocks. Note that basically, the system delay can be any integer delay, but for simplicity we focus on delays of integer numbers of blocks. For a low-latency system, d shoul be chosen as a small integer value.

The input / output relationship of the transmission scheme in the z domain can be derived as [1]:

$$\hat{\mathbf{u}}(z) = \mathbf{F}(z)z^{-1}\mathbf{C}(z)\mathbf{G}(z)\mathbf{u}(z) + \mathbf{F}(z)\mathbf{r}(z) \tag{2}$$

where $\mathbf{r}(z)$ is a vector containing the P polyphase components of the additive channel noise and $\mathbf{C}(z)$ describes the channel matrix

$$\mathbf{C}(z) = \begin{bmatrix} C_0(z) & z^{-1}C_{P-1}(z) & \cdots & z^{-1}C_1(z) \\ C_1(z) & C_0(z) & & z^{-1}C_2(z) \\ \vdots & & \ddots & \vdots \\ C_{P-1}(z) & C_{P-2}(z) & \cdots & C_0(z) \end{bmatrix}$$

whose entry $C_i(z)$, $i = 0, \ldots, P-1$, is the i-th type-I polyphase component [5] of the channel impulse response $C(z)$:

$$C(z) = \sum_{i=0}^{P-1} z^{-i} C_i(z^P) \tag{3}$$

3. TRAILING-ZEROS TRANSMITTER

In [1] a Trailing-Zeros Transmitter for joint transmitter / receiver optimization was proposed. The authors make the assumption that both, transmitter and receiver, are pure block transforms described by matrices $\mathbf{G}_0$ and $\mathbf{F}_0$, respectively, and that the order of channel impulse response does not exceed length of guard interval. Thus, due to the latter, this algorithm cannot be applied for low latency transmission with high bandwidth efficiency if the channel has a long impulse response. As redundant samples L zeros are introduced at the end of each block of M data symbols. The resulting transmission scheme is a special case o the general MIMO setting and is depicted in Figure 2. The input / output relationship writes:

$$\hat{\mathbf{u}}(z) = \mathbf{F}_0 \cdot \mathbf{C}(z)z^{-1} \begin{bmatrix} \mathbf{G}_0 \\ \mathbf{0}_{L\times M} \end{bmatrix} \mathbf{u}(z) + \mathbf{F}_0 \cdot \mathbf{r}(z) \tag{4}$$

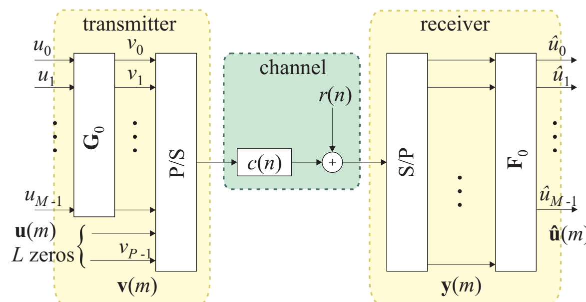

Figure 2: Trailing-Zeros Transmitter from [1]

In case that the above mentioned assumptions are met, the the entries of the channel matrix reduce to polynomials of a maximal degree of one:

$$\mathbf{C}(z) = \mathbf{C}_0 + z^{-1}\,\mathbf{C}_1 \tag{5}$$

and the input / output relationship in the time domain writes:

$$\hat{\mathbf{u}}(m) = \mathbf{F}_0 \cdot \mathbf{C}_0 \cdot \begin{bmatrix} \mathbf{G}_0 \\ \mathbf{0}_{L\times M} \end{bmatrix} \cdot \mathbf{u}(m-1) + \mathbf{F}_0 \cdot \mathbf{r}(m) \tag{6}$$

Observe that $\mathbf{C}_1$ vanishes because its non-zero entries are met by $\mathbf{0}_{L\times M}$. From the above equation it can be seen that no intersymbol interference occurs. A zero-forcing solution for $\mathbf{G}_0$ and $\mathbf{F}_0$ that minimizes the noise variance as well as an MMSE solution are provided in [1].

4. GENERALIZATION OF TRAILING-ZEROS TRANSMITTER

In the following we propose a generalization of the Trailing-Zeros Transmitter that allows for low latency time independently of the channel impulse response length. We make

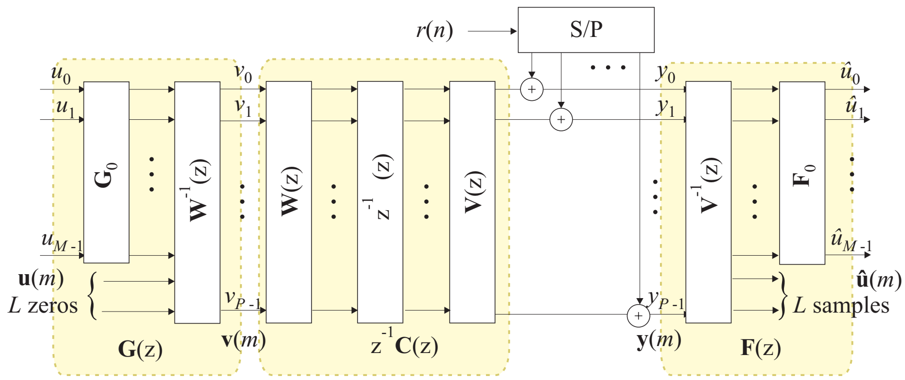

Figure 3: Smith decomposition of channel matrix; special structure of transmitter and receiver

the assumptions that both, transmitter and receiver, are FIR redundant filter banks described by $\mathbf{G}(z)$ and $\mathbf{F}(z)$, respectively, in Figure 1 and that the number of redundant samples L can be chosen independently of the filter length.

The input / output relationship in the z domain is thus given by (2). In order to find optimal transmitter and receiver, we apply the Smith decomposition [5] to the channel matrix $\mathbf{C}(z)$:

$$\mathbf{C}(z) = \mathbf{V}(z) \cdot \Lambda(z) \cdot \mathbf{W}(z)$$

with $\mathbf{V}(z)$ and $\mathbf{W}(z)$ being $P \times P$ unimodular matrices and $\Lambda(z)$ being a diagonal matrix, see Figure 3. Recall that a unimodular matrix is a matrix with FIR entries that has an FIR inverse. This can be seen as similar to channel equalization. Unlike the general Smith decomposition, we would like to include a delay z^{-n_0}, where n_0 is a small integer, in the diagonal matrix $\Lambda(z)$. We now choose the transmitter and receiver such that they diagonalize the channel matrix:

$$\mathbf{G}(z) = \mathbf{W}^{-1}(z) \begin{bmatrix} \mathbf{G}_0 \\ \mathbf{0}_{L\times M} \end{bmatrix}, \ \mathbf{F}(z) = [\mathbf{F}_0 \, \mathbf{0}_{M\times L}] \, \mathbf{V}^{-1}(z) \tag{7}$$

This choice can be interpreted as performing a channel optimized precoding at the transmitter (using $\mathbf{W}^{-1}(z)$) and equalization at the receiver (using $\mathbf{V}^{-1}(z)$). $\mathbf{G}_0$ and $\mathbf{F}_0$ are $M \times M$ block transforms that will be optimized jointly. For this transmitter and receiver the input / ouput relationship writes [5]:

$$\begin{aligned} \hat{\mathbf{U}}(z) &= [\mathbf{F}_0 \, \mathbf{0}_{M\times L}] \, z^{-1} \Lambda(z) \begin{bmatrix} \mathbf{G}_0 \\ \mathbf{0}_{L\times M} \end{bmatrix} \mathbf{U}(z) \\ &\quad + [\mathbf{F}_0 \, \mathbf{0}_{M\times L}] \, \tilde{\mathbf{R}}(z) \end{aligned} \tag{8}$$

with $\tilde{\mathbf{R}}(z) = \mathbf{V}^{-1}(z) \, \mathbf{R}(z)$. Observe that $\mathbf{V}(z)$ is not necessarily paraunitary. That means that a channel noise amplification can appear. To avoid or limit this effect, we included the delay n_0 in the diagonal matrix $\mathbf{\Lambda}(z)$, such that we choose the n_0 with the least noise amplification. In the Smith decomposition context this can be interpreted as choosing the biggest coefficient as pivot element for the decomposition. Its corresponding exponent of z then determines the delay.

If the P polyphase components $C_i(z)$ in the channel matrix $\mathbf{C}(z)$ do not contain common zeros, $\Lambda(z)$ writes:

$$\Lambda(z) = z^{-n_0} \text{diag}(1, \ldots, 1, \det(\mathbf{C}(z))) \tag{9}$$

and if $L \geq 1$, the input / output relationship in (8) becomes:

$$\hat{\mathbf{U}}(z) = z^{-1} \mathbf{F}_0 \, \mathbf{G}_0 \, \mathbf{U}(z) + \mathbf{F}_0 \, \tilde{\mathbf{R}}_\mathbf{0}(z) \tag{10}$$

and writes in the time domain

$$\hat{\mathbf{u}}(m) = \mathbf{F}_0 \, \mathbf{G}_0 \, \mathbf{u}(m-1) + \mathbf{F}_0 \, \tilde{\mathbf{r}}_\mathbf{0}(m) \tag{11}$$

$\tilde{\mathbf{R}}_\mathbf{0}(z)$ and $\tilde{\mathbf{r}}_\mathbf{0}(m)$ are column vectors of length M in the z and time domain, respectively, and contain the first M polyphase components (rows) of the noise vectors $\tilde{\mathbf{R}}(z)$ and $\tilde{\mathbf{r}}(m)$, respectively. Figure 4 shows the equivalent transmission scheme. Similar to (6), no intersymbol interference occurs in (11). Thus, the same algorithms as in [1] can be applied to jointly optimize the block matrices $\mathbf{G}_0$ and $\mathbf{F}_0$ of the FIR transmitter and receiver filter banks, respectively.

Maximizing the output SNR under a zero-forcing (ZF) constraint (perfect equalization in absence of additive noise), the optimization criterion writes:

$$\max_{\mathbf{G}_0, \mathbf{F}_0} \text{SNR} \qquad \text{subject to} \quad \mathbf{F}_0 \, \mathbf{G}_0 = \mathbf{I}_M \tag{12}$$

In [1] the following expression for the SNR has been derived:

$$\text{SNR} = \frac{trace(\mathbf{F}_0 \, \mathbf{G}_0 \mathbf{R}_{uu} \, \mathbf{G}_0^H \, \mathbf{F}_0^H)}{trace(\mathbf{F}_0 \, \mathbf{R}_{\tilde{r}_0 \tilde{r}_0} \, \mathbf{F}_0^H))} \tag{13}$$

where $\mathbf{R}_{uu}$ and $\mathbf{R}_{\tilde{r}_0 \tilde{r}_0}$ denote autocorrelation matrices

$$\mathbf{R}_{uu} = \mathcal{E}\{\mathbf{u}(m)\mathbf{u}^H(m)\}, \quad \mathbf{R}_{\tilde{r}_0 \tilde{r}_0} = \mathcal{E}\{\tilde{\mathbf{r}}_\mathbf{0}(m)\tilde{\mathbf{r}}_\mathbf{0}^H(m)\} \tag{14}$$

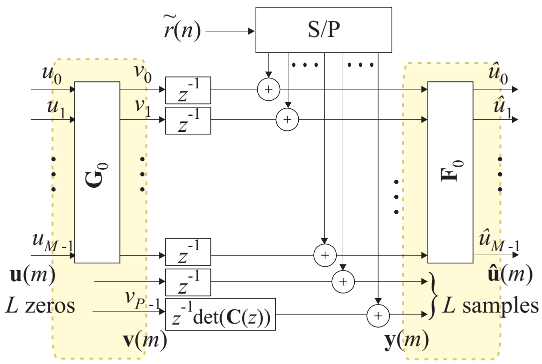

Figure 4: Equivalent transmission scheme obtained from simplifying Figure 3

respectively, and $\mathbf{u}^H(m)$ denotes the transpose hermitian of $\mathbf{u}(m)$. Under the ZF constraint $\mathbf{G}_0 \cdot \mathbf{F}_0 = \mathbf{I}_M$ where $\mathbf{I}_M$ denotes the $M \times M$ identity matrix, (13) writes:

$$\text{SNR} \stackrel{ZF}{=} \frac{trace(\mathbf{R}_{uu})}{trace(\mathbf{F}_0\, \mathbf{R}_{\tilde{r}_0\tilde{r}_0}\, \mathbf{F}_0^H)} \tag{15}$$

A solution for white input sequence $u(n)$ with power σ_u^2, which is a common case in communications for modulated input signals, is [1]:

$$\mathbf{G}_0 = \frac{1}{\sigma_u}\mathbf{V}_{\tilde{r}_0}\Lambda_{\tilde{r}_0}^{-1/2}, \quad \mathbf{F}_0 = \sigma_u\,\Lambda_{\tilde{r}_0}^{1/2}\mathbf{V}_{\tilde{r}_0}^H \tag{16}$$

with $\mathbf{R}_{\tilde{r}_0\tilde{r}_0}^{-1} = \mathbf{V}_{\tilde{r}_0}\,\Lambda_{\tilde{r}_0}\,\mathbf{V}_{\tilde{r}_0}^H$

5. CONCLUSIONS AND LIMITATIONS

In this paper we have provided a generalization of the Zero-Trailing Transmitter proposed in [1]. The generalization mainly consists of using redundant filter banks instead of block transforms at the transmitter and receiver. This allows us to choose the amount of redundancy inserted independently of the filter length. Thus, even for channels with a long channel impulse response the overall latency time can be small. Although the transmitter and receiver consist of redunandant FIR filter banks we only optimize coefficients of the $M \times M$ block transforms $\mathbf{G}_0$ and $\mathbf{F}_0$ in Figure 3. The remaining part of the transmitter is used to diagonalize the channel using Smith form for the channel matrix. On one hand, this procedure has the nice feature that for channels where the P polyphase components do not have common zeros, the Smith form not only diagonalizes the channel but also equalizes $P-1$ elements of the decoupled impulse responses in $\Lambda(z)$. On the other hand, if the channel impulse response has small coefficients at the beginning, synchronizing on these coefficients results in a significant SNR loss and a high range of values for the unimodular matrices $\mathbf{W}(z)$ and $\mathbf{V}(z)$. For that reason we introduced the delay z^{-n_0} in the diagonal matrix $\Lambda(z)$. Furthermore, these matrices are not unique and different realizations can differ significantly in length and numerical stability.

Future research has to focus on optimizing $\mathbf{W}(z)$, $\mathbf{V}(z)$, and $\mathbf{G}_0$, $\mathbf{F}_0$ together. Also, a better SNR can be expected if not only optimizing block matrices but FIR filters in the transmitter and receiver.

6. REFERENCES

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