Power Loading in MIMO Multicarrier Transmission Systems for Multi-Pair Cables

Lange, Christoph; Ahrens, Andreas

Abstract—*Crosstalk between neighbouring wire pairs is one of the major impairments in digital transmission via multi-pair copper cables. For high-rate transmission, often the strong near-end crosstalk (NEXT) disturbance is avoided or suppressed and only the far-end crosstalk (FEXT) remains as crosstalk influence. If FEXT is present, signal parts are transmitted via the FEXT paths from the transmitter to the receiver in addition to the direct transmission paths. Therefore transmission schemes are of great practical interest, which take advantage of the signal parts transmitted via the FEXT paths. Here a SVD (singular-value decomposition) equalized MIMO (multiple-input multiple-output) multicarrier system is investigated. Based on the Lagrange multiplier method an optimal power allocation scheme is considered in order to reduce the overall bit-error rate at a fixed data rate and fixed QAM constellation sizes. For high FEXT couplings between neighbouring wire pairs considerable gains are possible and the importance of FEXT exploitation becomes obvious.*

Index Terms—*Twisted-Pair Cable, OFDM, Power Allocation, Multiple-Input Multiple-Output System, Lagrange Multiplier Method, Singular-Value Decomposition.*

1. Introduction

THE local cable network substantially ensures the fixed subscriber access to telephone and data services. For most parts, this local cable network consists of multi-pair symmetric copper cables. Originally, these copper cables were installed for analogue telephone services in the lower frequency range. In addition, nowadays these copper-based local exchange networks are widely used for digital transmission, too.

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OFDM (orthogonal frequency division multiplex) is a widely accepted multicarrier transmission scheme in both, wireline and wireless transmission. Examples of it's application include digital subscriber line techniques (DSL) [4], [5], digital video broadcast (DVB), digital audio broadcast (DAB) and wireless local area networks (WLAN) such as 802.11a and HIPERLAN/2. A lot of articles have been published in the literature where the resilience of multicarrier transmission systems against typical channel distortions in wireline and wireless transmission is highlighted [5], [6].

In local cable networks, crosstalk is one of the most limiting disturbances, whereby near-end crosstalk (NEXT) and far-end crosstalk (FEXT) occur in bidirectional driven cables [7]. Since the NEXT is a very strong impairment several techniques have been developed in order to avoid or suppress it [8]. In this case only the FEXT remains as crosstalk influence. Often optical fibre transmission is used up to a building's entrance and the last few hundred metres within the building are bridged by copper cables. For such short cables used in high-data rate systems in the local cable area, the FEXT is particularly strong [7] and as a result heavy FEXT interferences between the signals on neighbouring wire pairs arise [4], [5].

In different publications, e. g, in [9], it was theoretically shown, that gains are possible by FEXT exploitation. In this contribution an interesting approach for the practical exploitation of the FEXT signal parts is presented: On each wire pair the multicarrier technique OFDM [10] is used and in addition the mutual impact of the wire pairs in a cable binder via far-end crosstalk is taken into account. Therefore the n-pair cable is modelled as a (n, n) MIMO transmission system and the combination of singular-value decomposition (SVD) and optimal power allocation using the Lagrange Multiplier method is considered with the aim of a bit-error rate minimization at a given data rate. Contrary to other publications considering a similar topic, here the focus lies on the combination

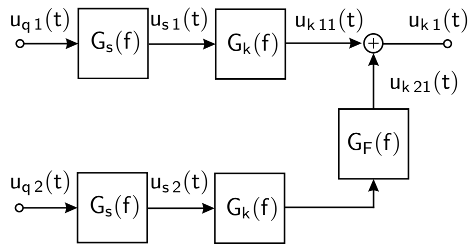

Fig. 1. Model of transmitter and cable with FEXT (example: $N = 2$)

of singular-value decomposition and power allocation. Thereby the optimal power allocation solution is presented for given boundary conditions (fixed QAM constellation size and limited total transmit power).

The remaining part of this contribution is organized as follows: In section 2 the cable characteristics are given. Section 3 introduces the considered system model including the MIMO-OFDM transmission system with SVD-based equalization. In section 4 possible optimization objectives for MIMO transmission systems are discussed and the underlying optimization criteria are briefly reviewed. In section 5 the transmit power allocation scheme is explained and in section 6 the obtained results are presented and discussed. Finally, section 7 provides some concluding remarks.

2. Cable characteristics

The distorting influence of the cable on the wanted signal is modelled by the cable transfer function

$$G_{\rm k}(f) = {\rm e}^{-l\sqrt{{\rm j}\frac{f}{f_0}}} \ , \tag{1}$$

where l denotes the cable length (in $\rm km$) and f_0 represents the characteristic cable frequency (in $\rm MHz \cdot km^2$) [11]. The far-end crosstalk coupling is covered by the transfer function $G_{\rm F}(f)$ with

$$|G_{\rm F}(f)|^2 = K_{\rm F} \cdot l \cdot f^2 \ , \tag{2}$$

whereby $K_{\rm F}$ is a coupling constant of the far-end crosstalk, which depends on the cable properties such as the type of isolation, the number of wire pairs and the kind of combination of the wire pairs within the binders [7].

The far-end crosstalk signal $u_{\rm k\,21}(t)$ arises, when the transmit signal at the output of the transmit filter $G_{\rm s}(f)$ of the disturbing transmitter passes through the cascade of FEXT coupling transfer function and cable transfer function $G_{\rm k}(f) \cdot G_{\rm F}(f)$ (see Fig. 1). Therefore the effective FEXT

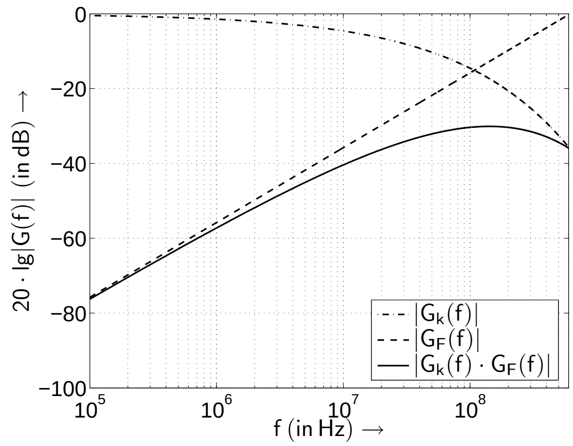

Fig. 2. Cable and FEXT coupling transfer functions for a cable length of $l = 0.1\,\rm km$ with $K_{\rm F} = 2.6248 \cdot 10^{-17}\,({\rm Hz}^2 \cdot {\rm km})^{-1}$

influence is characterized by the overall FEXT transfer function $G_{\rm k}(f) \cdot G_{\rm F}(f)$. This transfer function attenuates the FEXT coupled signal much more in longer cables than in shorter ones, which becomes obvious by comparing Fig. 2 and Fig. 3. Therefore it can be stated, that the FEXT impact is much stronger in short cables than in longer cables [7].

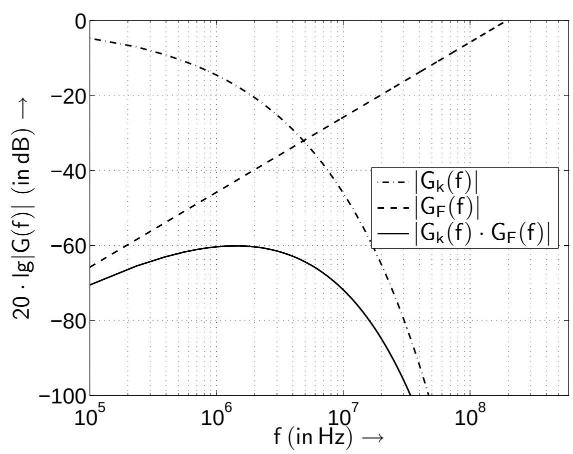

Fig. 3. Cable and FEXT coupling transfer functions for a cable length of $l = 1.0\,\rm km$ with $K_{\rm F} = 2.6248 \cdot 10^{-17}\,({\rm Hz}^2 \cdot {\rm km})^{-1}$

3. System model

The considered cable binder consists of n wire pairs and therefore a (n, n) MIMO transmission system arises. The mapping of the transmit signals $u_{{\rm s}\,\mu}(t)$ onto the received signals $u_{{\rm k}\,\mu}(t)$ (with $\mu = 1, \ldots, n$) can be described accordingly to Fig. 4. On each wire pair of the cable binder OFDM (orthogonal frequency division multiplexing) is used as multicarrier transmission technique to combat the effects of the frequency-selective

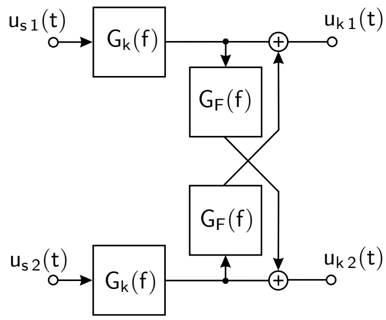

Fig. 4. MIMO cable transmission model system with FEXT ($n=2$)

channel [5], [12]. Altogether a multicarrier system with N subchannels is considered. In combination with a sufficient guard interval length, a subcarrier specific description can be obtained in the following form as shown in [2] and [3]:

$$\widetilde{\boldsymbol{u}}_\kappa = \mathbf{R}_\kappa \cdot \widetilde{\boldsymbol{a}}_\kappa + \boldsymbol{n}_\kappa \ . \tag{3}$$

The matrices $\mathbf{R}_\kappa$ (with $\kappa = 1, \ldots, N$) describe the subcarrier specific distortions and can be defined according to

$$\mathbf{R}_\kappa = \begin{bmatrix} r_{1\,1}^{(\kappa)} & \cdots & r_{1\,n}^{(\kappa)} \\ \vdots & \ddots & \vdots \\ r_{n\,1}^{(\kappa)} & \cdots & r_{n\,n}^{(\kappa)} \end{bmatrix} \ , \tag{4}$$

with the matrix elements describing the couplings between the data symbols on the subchannel κ. Based on the symmetry of the considered transmission system $r_{\nu\,\mu}^{(\kappa)}$ (for $\nu = \mu$) can be determined by taking the FFT of $g_{\mathrm{k}}(t) = \mathcal{F}^{-1}\left\{G_{\mathrm{k}}(f)\right\}$ into account. The elements $r_{\nu\,\mu}^{(\kappa)}$ (for $\nu \neq \mu$) consider the coupling between neighbouring wire pairs and can be ascertained calculating the FFT of $g_{\mathrm{k\,fn}}(t) = \mathcal{F}^{-1}\left\{G_{\mathrm{F}}(f) \cdot G_{\mathrm{k}}(f)\right\}$. The κth value of this vector represents $r_{\nu\,\mu}^{(\kappa)}$. The elements $r_{\nu\,\mu}^{(\kappa)}$ (for $\nu \neq \mu$) are assumed to be identical for each κ, although in practical systems the coupling between the wire pairs is slightly different and it depends on their arrangement in the binder [7]. The data vector $\widetilde{\boldsymbol{a}}_\kappa$ contains now the data symbols that are transmitted via the same subcarrier and results in

$$\widetilde{\boldsymbol{a}}_\kappa = \left(a_{\kappa\,1}, \ldots, a_{\kappa\,\mu}, \ldots, a_{\kappa\,n}\right)^{\mathrm{T}} \ . \tag{5}$$

Additionally a white Gaussian noise with power spectral density Ψ_0 is assumed, which results after receive filtering in the vector $\boldsymbol{n}_\kappa$ and can be defined similar to (5) as

$$\boldsymbol{n}_\kappa = \left(n_{\kappa\,1}, \ldots, n_{\kappa\,\mu}, \ldots, n_{\kappa\,n}\right)^{\mathrm{T}} \ . \tag{6}$$

Thereby it is assumed that the noise components are independently from each other, which can be justified by the rectangular shape of the receive filter functions in the time domain.

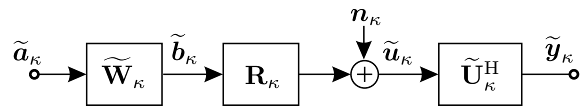

Fig. 5. Resulting block oriented MIMO-OFDM transmission system for one subcarrier

The remaining interferences on each subcarrier can now be eliminated by an efficient equalization strategy. A popular strategy is represented by the singular value decomposition (SVD), which can be done on each subcarrier separately [3]. The SVD of the matrix $\mathbf{R}_\kappa$ can be written as

$$\mathbf{R}_\kappa = \widetilde{\mathbf{U}}_\kappa \cdot \widetilde{\mathbf{V}}_\kappa \cdot \widetilde{\mathbf{W}}_\kappa^{\mathrm{H}} \ , \tag{7}$$

where $\widetilde{\mathbf{U}}_\kappa$ and $\widetilde{\mathbf{W}}_\kappa^{\mathrm{H}}$ are unitary matrices and $\widetilde{\mathbf{V}}_\kappa$ is a real diagonal matrix [13]. The data vector $\widetilde{\boldsymbol{a}}_\kappa$ (5) is multiplied by the matrix $\widetilde{\mathbf{W}}_\kappa$ and results in the transmit data vector $\widetilde{\boldsymbol{b}}_\kappa$. The received vector $\widetilde{\boldsymbol{u}}_\kappa = \mathbf{R}_\kappa \cdot \widetilde{\boldsymbol{b}}_\kappa + \boldsymbol{n}_\kappa$ is multiplied by the matrix $\widetilde{\mathbf{U}}_\kappa^{\mathrm{H}}$. Thereby neither the transmit power nor the noise power is enhanced. The overall transmission relationship for the subcarrier κ (with $\kappa = 1, \ldots, N$) is defined as (see Fig. 5)

$$\begin{aligned} \widetilde{\boldsymbol{y}}_\kappa &= \widetilde{\mathbf{U}}_\kappa^{\mathrm{H}} \cdot \widetilde{\boldsymbol{u}}_\kappa = \widetilde{\mathbf{U}}_\kappa^{\mathrm{H}} \cdot \left(\mathbf{R}_\kappa \cdot \widetilde{\mathbf{W}}_\kappa \cdot \widetilde{\boldsymbol{a}}_\kappa + \boldsymbol{n}_\kappa\right) \\ &= \widetilde{\mathbf{V}}_\kappa \cdot \widetilde{\boldsymbol{a}}_\kappa + \widetilde{\mathbf{U}}_\kappa^{\mathrm{H}} \cdot \boldsymbol{n}_\kappa \ . \end{aligned} \tag{8}$$

The SVD is done per subchannel, because due to a sufficient guard interval length the subchannels are assumed to be independent from each other. A global system matrix can be constituted of the subchannel system matrices (4) according to [1]–[3]: Then the SVD can also be applied to the global system matrix as shown e. g. in [14].

4. Optimization objectives and quality criteria

The optimization strategies for MIMO systems typically fall into two categories: data throughput maximization at a given bit-error rate or bit-error rate minimization at a fixed data rate [15], [16]. In this contribution we have restricted ourselves to the BER minimization at a fixed data rate.

The bit-error probability in the general case of M-ary quadrature amplitude modulation (QAM) is given by

$$P_{\mathrm{f}} = \frac{2}{\mathrm{ld}(M)} \left(1 - \frac{1}{\sqrt{M}}\right) \operatorname{erfc}\left(\frac{U_{\mathrm{A}}}{\sqrt{2} U_{\mathrm{R}}}\right) \ , \tag{9}$$

where U_{A} denotes the half vertical eye opening and U_{R}^2 is the noise disturbance power per quadrature component, respectively [17], [18].

5. Transmit power allocation

Power allocation has been widely investigated in the literature. Thereby optimal but highly complex or suboptimal solutions with reduced complexity can be found, e. g. [14]–[16], [19]–[21].

The SVD-based equalization on each subcarrier leads to a different half vertical eye opening

$$U_{\mathrm{A}}^{(\mu)} = \sqrt{\xi_\mu} \cdot U_{\mathrm{s}} \qquad (10)$$

for each data symbol. Here, U_{s} denotes the half-level transmit amplitude (see Fig. 6) and $\sqrt{\xi_\mu}$ are the positive square roots of the eigenvalues of the matrix $\mathbf{R}_\kappa^{\mathrm{H}}\,\mathbf{R}_\kappa$, describing the distortions on each subcarrier. Therefore all eye openings are in general different from each other. Assuming an identical noise power for all symbol positions, the symbol positions with the smallest half vertical eye openings dominate the bit-error rate. Here a transmit power partitioning scheme would be necessary in order to minimize the overall bit-error rate under the constraint of a limited total transmit power. In the following all N_{b} symbols which are

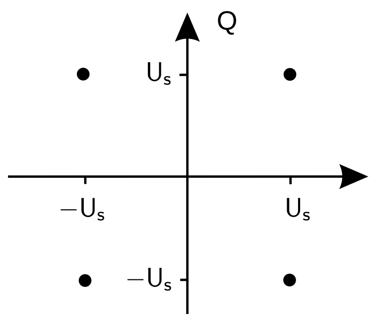

Fig. 6. Definition of the half-level transmit amplitude U_{s} for QPSK ($M = 4$)

simultaneously transmitted over the n wire pairs of the binder are taken into account by the power allocation. The power allocation evaluates the half-level amplitude U_{s} of the μth symbol by the factor $\sqrt{p_\mu}$. This causes in general a modified transmit amplitude $U_{\mathrm{s}}\,\sqrt{p_\mu}$ for each symbol of the transmit data vector and the signal constellation is modified according to Fig. 7; the half vertical eye opening changes to

$$U_{\mathrm{A,PA}}^{(\mu)} = \sqrt{p_\mu} \cdot \sqrt{\xi_\mu} \cdot U_{\mathrm{s}} \qquad (11)$$

per symbol after power allocation. Furthermore, each symbol is disturbed by a noise with identical disturbance power in the quadrature components, which is assumed to be uncorrelated with power U_{R}^2 each. Using (9) and (11), together with the

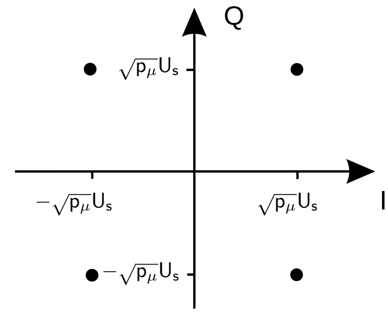

Fig. 7. QPSK constellation of the μth symbol after power allocation

noise disturbance per quadrature component a BER per symbol and block can be calculated:

$$P_{\mathrm{f}\,\mu} = \frac{2\left(1 - \frac{1}{\sqrt{M}}\right)}{\mathrm{ld}(M)}\,\mathrm{erfc}\left(\sqrt{\frac{p_\mu\,\xi_\mu}{2}} \cdot \frac{U_{\mathrm{s}}}{U_{\mathrm{R}}}\right) \ . \qquad (12)$$

The aggregate bit-error probability per block

$$P_{\mathrm{f}} = \frac{2\left(1 - \frac{1}{\sqrt{M}}\right)}{\mathrm{ld}(M)\,N_{\mathrm{b}}} \sum_{\mu=0}^{N_{\mathrm{b}}-1} \mathrm{erfc}\left(\sqrt{\frac{p_\mu\,\xi_\mu}{2}} \cdot \frac{U_{\mathrm{s}}}{U_{\mathrm{R}}}\right) \qquad (13)$$

is obtained by averaging over the error probabilities of all N_{b} symbols of the data block, which are simultaneously transmitted over the n wire pairs of the binder.

In the subchannels of the multicarrier system investigated in this contribution M-ary square QAM with transmit power [18]

$$P_{\mathrm{s\,QAM}} = \frac{2}{3}\,U_{\mathrm{s}}^2\,(M-1) \qquad (14)$$

is used [18]. Using a parallel transmission over N subchannels the overall mean transmit power per wire pair yields to

$$P_{\mathrm{s}} = N \cdot P_{\mathrm{s\,QAM}} = N\,\frac{2}{3} U_{\mathrm{s}}^2 (M-1) \ , \qquad (15)$$

and results in a total transmit power of $n\,P_{\mathrm{s}}$ by taking n wire pairs per binder into account.

Considering now generally different half-level amplitudes $U_{\mathrm{s}}\,\sqrt{p_\mu}$ after power allocation on the symbol layers, it follows

$$P_{\mathrm{s}\,\mu} = \left(\sqrt{p_\mu}\right)^2 \cdot \frac{2}{3} U_{\mathrm{s}}^2 (M-1) = p_\mu\,\frac{2}{3} U_{\mathrm{s}}^2 (M-1) \qquad (16)$$

for the μth symbol position.

If now a block of N_{b} data symbols, transmitted over N parallel subchannels and n wire pairs, is analyzed with these generally different half-level amplitudes $U_{\mathrm{s}}\,\sqrt{p_\mu}$ after power allocation, the mean transmit power of the block becomes

$$P_{\mathrm{s,PA}} = n\,N\,\frac{2}{3} U_{\mathrm{s}}^2 (M-1)\,\frac{1}{N_{\mathrm{b}}} \sum_{\mu=0}^{N_{\mathrm{b}}-1} \left(\sqrt{p_\mu}\right)^2 \ . \qquad (17)$$

From the requirement

$$P_{\mathrm{s,PA}} - n\,P_{\mathrm{s}} = 0 \tag{18}$$

that the overall mean transmit power for the whole binder consisting of n wire pairs is limited to $n\,P_{\mathrm{s}}$ it follows, that the auxiliary condition

$$B_{N_{\mathrm{b}}} = \sum_{\mu=0}^{N_{\mathrm{b}}-1} p_\mu - N_{\mathrm{b}} = 0 \tag{19}$$

has to be maintained [3].

In order to find the optimal $\sqrt{p_\mu}$ the Lagrange multiplier method is used. Contrary to other publications (e. g. [15]), here the power allocation has not been carried out on each subcarrier independently from each other, although this might be possible. To consider the subcarrier specific distortions in a best possible way, here all subcarrier singular values are combined in one vector, in order to smooth out the distortions. The Lagrangian cost function $J(p_0, \cdots, p_{N_{\mathrm{b}}-1})$ may be expressed as

$$J(\cdots) = \frac{A}{N_{\mathrm{b}}} \sum_{\mu=0}^{N_{\mathrm{b}}-1} \mathrm{erfc}\left(\sqrt{\frac{p_\mu\,\xi_\mu}{2}} \cdot \frac{U_{\mathrm{s}}}{U_{\mathrm{R}}}\right) + \lambda \cdot B_{N_{\mathrm{b}}} \;, \tag{20}$$

with the Lagrange multiplier λ [15] and

$$A = \frac{2}{\mathrm{ld}(M)} \left(1 - \frac{1}{\sqrt{M}}\right) \;. \tag{21}$$

Differentiating the Lagrangian cost function $J(p_0, \cdots, p_{N_{\mathrm{b}}-1})$ with respect to the p_μ and setting it to zero, leads to the optimal set of power allocation coefficients. As solution for the p_μ (a computer algebra system such as MAPLE or MATLAB may come in handy)

$$p_\mu = \frac{1}{\xi_\mu} \frac{U_{\mathrm{R}}^2}{U_{\mathrm{s}}^2}\, \mathrm{W}\left(\frac{A^2\,\xi_\mu^2}{2\,\pi\,N_{\mathrm{b}}^2\,\lambda^2} \frac{U_{\mathrm{s}}^4}{U_{\mathrm{R}}^4}\right) \tag{22}$$

is obtained, where $\mathrm{W}(x)$ describes the Lambert W function [22]. The parameter λ can be calculated by insertion of (22) in (19) and numeric analysis. With calculated λ the optimal p_μ can be determined using (22). Figure 8 shows the Lambert W function for positive arguments. From Fig. 8 it becomes obvious, that very small p_μ are possible. Therefore it can be stated that optimal power allocation could behave like a waterfilling, since more power is spent to the less attenuated data symbols.

Power allocation with lower complexity can be achieved by suboptimal methods, which can on the one hand rely on an approximation for the $\mathrm{erfc}(x)$ function or which ensure on the other hand equal signal-to-noise ratios per symbol [16].

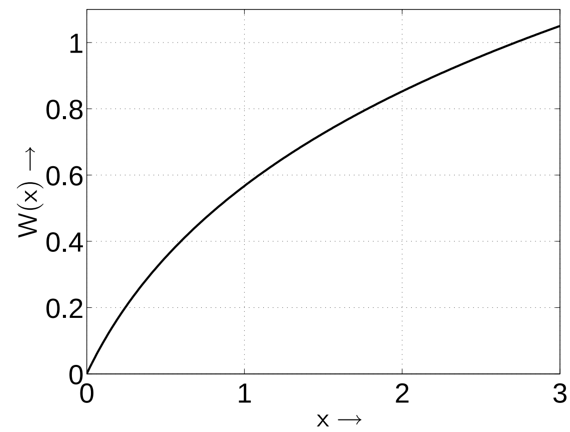

Fig. 8. Lambert W function for positive argument values

6. Results

The FEXT impact is in particular strong for short cables [7]. Therefore for numerical analysis an exemplary cable of length $l = 0.4\,\mathrm{km}$ with $n = 10$ wire pairs is chosen. The wire diameter is $0.6\,\mathrm{mm}$ and hence a characteristic cable frequency of $f_0 = 0.178\,\mathrm{MHz} \cdot \mathrm{km}^2$ is assumed. On each of the wire pairs a multicarrier system with $N = 10$ subcarriers was considered. The actual crosstalk circumstances are difficult to acquire and they vary from cable to cable. Therefore an exemplary mean FEXT coupling constant of $K_{\mathrm{F}} = 10^{-13}\,(\mathrm{Hz}^2 \cdot \mathrm{km})^{-1}$ is employed [7], [23]. The average transmit power on each wire pair is supposed to be $P_{\mathrm{s}} = 1\,\mathrm{V}^2$ and as an external disturbance a white Gaussian noise with power spectral density Ψ_0 is assumed. Identical systems on all wire pairs are presumed (multicarrier symbol duration $T_{\mathrm{s}} = 2\,\mu\mathrm{s}$, M-ary QAM, a block length of $n_{\mathrm{b}} = 10$ and a guard interval length of $T_{\mathrm{g}} = T_{\mathrm{s}}/2$). Furthermore, the baseband channel of the multicarrier system is excluded from the transmission in order to provide this frequency range for analogue telephone transmission.

For a fair comparison the ratio of symbol energy to noise power spectral density at the cable output is defined for the MIMO case $(n > 1)$ according to

$$\frac{E_{\mathrm{s}}}{\Psi_0} = (T_{\mathrm{s}} + T_{\mathrm{g}})\; \frac{P_{\mathrm{k}} + (n-1)P_{\mathrm{k\,fn}}}{\Psi_0} \;, \tag{23}$$

with P_{k} as mean power of the signal on the direct paths at the cable output and $P_{\mathrm{k\,fn}}$ as mean FEXT signal power at the cable output [3].

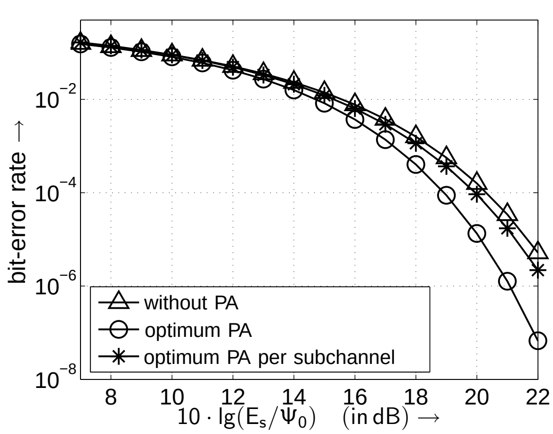

Fig. 9. BER comparison analyzing MIMO-OFDM ($M = 4, n = 10, K_{\mathrm{F}} = 10^{-13}\,(\mathrm{Hz}^2 \cdot \mathrm{km})^{-1}$) with and without PA

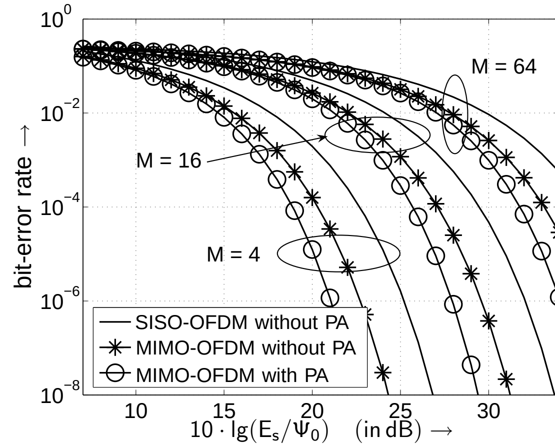

Fig. 10. BER comparison analyzing SISO-OFDM ($n = 1$) and MIMO-OFDM ($n = 10, K_{\mathrm{F}} = 10^{-13}\,(\mathrm{Hz}^2 \cdot \mathrm{km})^{-1}$) with and without PA for different QAM constellation sizes

A. Cooperative versus non-cooperative design

The power allocation to the wire pairs of the binder can be done for each subchannel separately (non-cooperative) or jointly for all subchannels of all wire pairs (cooperative). The performance of the investigated power allocation schemes is depicted in Fig. 9 in order to show the efficiency of the proposed carrier-cooperative setup (combined power allocation for all subchannels of all wire pairs) against the non-cooperative one (power allocation for the subchannels on all wire pairs individually). The performance gains are clearly recognizable if all subcarriers are jointly taken into account. The analysis of the subcarrier specific bit-error rates under the boundary condition of fixed QAM constellation sizes leads to an increased bit-error rate with increasing subcarrier number: The cable attenuation increases with increasing frequency and is linearly equalized; this leads to an enhanced noise power with increasing frequency. The overall bit-error characteristic will be mainly determined by the largest subcarrier bit-error rate. Therefore based on an increasing cable attenuation with increasing frequency under the restriction of fixed QAM constellation sizes a cooperative system design seems to be very advantageous in cable transmission systems.

B. Results for different QAM constellation sizes

In Fig. 10 numerical BER results are presented with the QAM constellation sizes M as parameter. For purposes of simplicity and in order to obtain meaningful results, the QAM constellation sizes are chosen to be equal in all subchannels of the multicarrier systems. This seems to be reasonable in the example considered here, since short cables do not have very strong frequency-selective characteristics. For general cable transmission the optimization of combined power and bit loading with low complexity in the MIMO context remains open for further investigations. Furthermore it seems to be worth mentioning that in case of different QAM constellation sizes M also different overall bit rates can be achieved. This could be used to adapt the bit rate to the user's needs. In case of MIMO-OFDM the signal parts, which are transmitted via the FEXT paths are no longer disturbance: Now they are exploited as useful signal parts. Therefore the transmission quality is improved compared to the SISO-OFDM case (OFDM transmission over a (fictive) perfectly shielded single wire pair). Similar results are known from MIMO radio transmission with multiple transmit and/or receive antennas, where multiple transmission paths are exploited, too [24], [25].

The results show that under severe FEXT influence it is worth taking the FEXT signal paths into account (Fig. 10). At smaller FEXT couplings no significant gains are possible by MIMO-OFDM without power allocation compared to a perfectly shielded wire pair (SISO-OFDM), because the FEXT coupled signal parts are very small [2].

The results in Fig. 10 further show the potential of appropriate power allocation strategies. The absolut achievable gains depend on the actual cable type, on the isolation of the wire pairs and on the arrangement of the wire pairs inside the cable binder.

7. Conclusion

In this contribution, the practical exploitation of the FEXT paths for improving the signal transmission quality is investigated in terms of an exemplary SVD-equalized multicarrier transmission system on a symmetric copper cable. It is shown, that the MIMO-OFDM cable transmission enables gains in the BER performance especially under severe FEXT influence. Thereby it could be shown that power allocation is necessary to achieve a minimum bit-error rate. In the exemplary system considered here some restrictions are made, which directly lead to some open points for further investigations: In order to use MIMO-OFDM for cables of any length the most important open point is the optimization of bit loading in combination with the power allocation in the MIMO-OFDM context.

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