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Equalization Design for Dispersive MIMO Channels subject to Channel Variability

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***Abstract*—Multiple-input multiple-output (MIMO) equalizers combat the large amount of intersymbol interference and crosstalk generated in high-speed communication over dispersive MIMO channels. Unfortunately, fully adjusting these equalizers to the specific channel could induce intolerably large complexity. This contribution proposes an equalization strategy in which (part of) the equalizers are fixed and only depend on the channel statistics instead of the specific channel. The numerical results confirm that a hybrid strategy could indeed be a practical low-complexity alternative to the fully adjustable equalizers.**

***Index Terms*—MIMO Equalization, Frequency-Selective Channel, Channel Variability, mean square error methods**

I. Introduction

The continuous demand for larger communication rates requires equalization schemes that properly manage the intersymbol interference and crosstalk generated by the dispersive multiple-input multiple-output (MIMO) channel. Due to the extremely high complexity of the optimal maximum-likehood sequence detector, less complex linear equalizers at the transmitter (TX) and/or receiver (RX) are more attractive in practice [1]. To improve these linear equalizers, they can be combined with a nonlinear precoder at the TX [2] or a decision feedback equalizer (DFE) at the RX [3].

In the literature, equalizers are generally adjusted to the specific channel realization. Although this strategy naturally results in the best performance, a high computational and implementational complexity is induced, which can be infeasible for some applications. However, only a limited performance degradation is expected when considering fixed equalizers when channel variability is small.

This contribution discusses the main results presented in [4], where the design of a general MIMO transceiver scheme is investigated. More precisely, the less-complex hybrid strategy, in which the TX is fixed for all channels and the RX is adjusted to the specific channel realization, is analyzed and compared with the fully adjustable and fully fixed equalizers in terms of mean square error (MSE) and symbol error rate (SER).

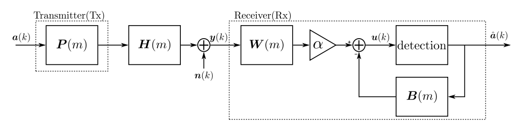

Figure 1. MIMO channel $\mathbf{H}(m)$ equalized with a pre-equalizer $\mathbf{P}(m)$ at the TX and a DFE at the RX comprising a feedforward filter $\mathbf{W}(m)$ and a feedback filter $\mathbf{B}(m)$.

II. System model

Fig. 1 presents a MIMO baseband communication link, in which the TX first applies L streams of independent M-PAM symbols $\boldsymbol{a}(k)$ to a finite impulse response (FIR) $L \times L$ pre-equalizer $\mathbf{P}(m)$ with L_w taps, i.e., $\mathbf{P}(m) = 0\,\forall m \notin [L_w^{(1)}, L_w^{(2)}]$. Furthermore, the pre-equalizer is selected such that the averaged energy per symbol interval is constraint to LE_s. Subsequently, the pre-equalizer's output is transmitted over the dispersive and mutually coupled $L \times L$ channel $\mathbf{H}(m)$, which is obtained by sampling the convolution of the impulse responses of the transmit filter, channel and receive filter. Moreover, stationary noise represented by the noise samples $\boldsymbol{n}(k)$ affects the channel output. To compute a decision $\hat{\boldsymbol{a}}(k)$, the RX first equalizes the received signal using a DFE, and then obtains $\hat{\boldsymbol{a}}(k)$ from a symbol-by-symbol detector. This DFE comprises a FIR $L \times L$ feedforward filter $\mathbf{W}(m)$ of length $L_p (= L_p^{(1)} + L_p^{(2)} + 1)$ with scaling factor α, combined with a FIR feedback filter $\mathbf{B}(m)$ operating on the previously detected symbols. When the past symbols are assumed to be correctly detected, the decision variable $\boldsymbol{u}(k)$ can be decomposed as

$$\begin{aligned}\boldsymbol{u}(k) =& \alpha \sum_{m \in \boldsymbol{\Phi}_{\mathbf{G}}} \mathbf{W}\mathbf{G}(m)\mathbf{P}\boldsymbol{a}(k-m) + \alpha \mathbf{W}\tilde{\boldsymbol{n}}(k) \\ &- \sum_{m \in \boldsymbol{\Phi}_{\mathbf{b}}} \mathbf{B}(m)\boldsymbol{a}(k-m), \end{aligned} \tag{1}$$

where $\boldsymbol{\Phi}_{\mathbf{G}}$ and $\boldsymbol{\Phi}_{\mathbf{b}}$ are the sets of time instants on which $\mathbf{W}\mathbf{G}(m)\mathbf{P}$ and the feedback filter, respectively, are nonzero. The matrix $L \times (LL_w)$ $\mathbf{W}$ is constructed by rearranging all

$\mathbf{W}(m)$ into a row block matrix, and $(LL_p) \times L$ $\mathbf{P}$ is obtained by stacking all $\mathbf{P}(m)$ into a column block matrix. Further, the $(LL_w) \times (LL_p)$ matrix $\mathbf{G}(m)$ is defined as follows

$$\mathbf{G}(m) = \begin{bmatrix} \mathbf{H}_m(-L_w^{(1)}, -L_p^{(1)}) & \dots & \mathbf{H}_m(-L_w^{(1)}, L_p^{(2)}) \\ \vdots & & \vdots \\ \mathbf{H}_m(L_w^{(2)}, -L_p^{(1)}) & \dots & \mathbf{H}_m(L_w^{(2)}, L_p^{(2)}) \end{bmatrix}, \tag{2}$$

where $\mathbf{H}_m(m_1, m_2) = \mathbf{H}(m - m_1 - m_2)$.

III. MMSE MIMO EQUALIZER

The MIMO channel $\mathbf{H}(m)$ is presumed to be stochastic; the impulse response is static, but variability is present among the different realizations due to, for example, manufacturing tolerances. Ideally, all equalization coefficients are computed with respect to the specific channel realization. However, equalization algorithms that consider (part of) the equalization structure completely independent of the specific channel realization are of particular interest, since they can considerably lower the high complexity associated with the fully adjustable equalizers. Hence, this contribution not only presents the all-adjustable strategy S-A (adjustable TX, adjustable RX), but also the fixed strategy S-F (fixed TX, fixed RX) and the hybrid strategy S-H (fixed TX, adjustable RX). As for the adjustable parts, the equalization coefficients are obtained such that the $\text{MSE}_{\mathbf{G}}$ of a specific MIMO channel is minimized. This $\text{MSE}_{\mathbf{G}}$ is a function of all equalization coefficients and is defined as

$$\text{MSE}_{\mathbf{G}}(\mathbf{P}, \mathbf{W}, \alpha, \mathbf{B}) \triangleq \frac{\mathbb{E}\left[\|\boldsymbol{u}(k) - \boldsymbol{a}(k)\|^2\right]}{\mathbb{E}\left[\|\boldsymbol{a}(k)\|^2\right]}. \tag{3}$$

In contrast, the fixed parts are obtained by minimizing $\text{MSE}_{\text{avg}}(\mathbf{P}, \mathbf{W}, \alpha, \mathbf{B}) = \mathbb{E}_{\mathbf{G}}[\text{MSE}_{\mathbf{G}}]$. To compute the equalization coefficients, first the optimal feedback filter $\mathbf{B}^\star$ is written as a function of the other optimization parameters by equating the derivative of $\text{MSE}_{\mathbf{G}}$ or MSE_{avg} with respect to $\mathbf{B}$ to zero for S-A and S-H, or S-F, respectively. Unfortunately, the direct optimization of the resulting $\text{MSE}_{\mathbf{G}}(\mathbf{P}, \mathbf{W}, \alpha, \mathbf{B}^\star)$ and $\text{MSE}_{\text{avg}}(\mathbf{P}, \mathbf{W}, \alpha, \mathbf{B}^\star)$ is infeasible and thus an iterative algorithm similar to [5] is proposed. More precisely, the optimal TX ($\mathbf{P}^\star$,$\alpha^\star$) and optimal RX ($\mathbf{W}^\star$) are alternatingly computed with the assumption that the other one is fixed. As for the adjustable TX in S-A, the optimization of $\text{MSE}_{\mathbf{G}}(\mathbf{P}, \mathbf{W}^\star, \alpha, \mathbf{B}^\star)$ is performed for each realization individually, whereas the optimal fixed TX in S-H and S-F is determined by minimizing $\text{MSE}_{\text{avg}}(\mathbf{P}, \mathbf{W}^\star, \alpha, \mathbf{B}^\star)$. Similarly, the adjustable RX in S-A and S-F is obtained by optimizing $\text{MSE}_{\mathbf{G}}(\mathbf{P}^\star, \mathbf{W}, \alpha^\star, \mathbf{B}^\star)$, while the fixed RX in S-F results from the minimization of $\text{MSE}_{\text{avg}}(\mathbf{P}^\star, \mathbf{W}, \alpha^\star, \mathbf{B}^\star)$.

IV. NUMERICAL RESULTS

The equalization concepts introduced in Section III are applied to an electrical MIMO chip-to-chip interconnect. Due to manufacturing tolerances, variability among the various realizations of the interconnect exists. This is modeled by considering 6 geometrical and material parameters as stochastic variables with standard deviation σ_r. For each σ_r, 1000 realizations are generated according to the approach in [6].

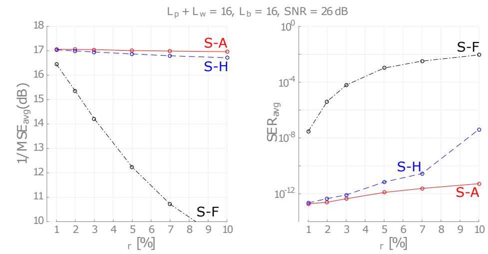

Figure 2. $1/\text{MSE}_{\text{avg}}$ and SER_{avg} performance of different strategies as a function of σ_r. At low σ_r, S-H is satisfactory alternative for S-A, since only limited degradation is induced by the fixed TX.

Fig. 2 examines $1/\text{MSE}_{\text{avg}}$ and SER_{avg} for the three equalization strategies. The pre-equalization, feedforward and feedback filter have respectively $L_p = 11$, $L_w = 5$, and $L_b = 16$ taps; the signal-to-noise ratio (SNR) is defined as $\text{SNR} = \frac{LE_s}{\|\boldsymbol{n}(k)\|^2}$ and equals 26 dB. As can be observed, S-A performs the best and maintains this performance constantly for all considered σ_r. More interestingly, S-H is suitable as a less complex alternative for S-A at low σ_r as only limited performance degradation is visible. Only at high σ_r, the difference between S-A and S-H becomes intolerably large. On the other hand, S-F cannot handle any channel variability, making it not attractive in practice.

V. CONCLUSION

This contribution proposes three equalization strategies for a dispersive MIMO channel subject to channel variability. Optimal performance is reached when both TX and RX are adjustable to the specific channel realization. However, simulations confirm that the hybrid strategy, with a fixed TX that is computed based on the channel statistics, is a suitable low-complexity alternative. Finally, the fully fixed strategy cannot handle any channel variability and thus achieves a seriously degraded performance.

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