Robust Linear MIMO Receivers: A Minimum Error-Rate Approach

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Abstract

This paper looks at the linear reception of spatially multiplexed signals across MIMO channels. We address the problem of robustness in the presence of detrimental effects such as correlation and Ricean components. We consider the error-rate performance of MIMO linear filters, as these can be used in purely linear receivers, or as part of each stage in successive interference canceling (SIC) receivers. We know from multiuser detection theory that minimum error-rate (MER) linear receivers significantly outperform minimum mean-square error (MMSE) receivers when correlation is high, however no direct method exist to design the MER receiver simply. We derive a scheme allowing a closed-form approximate solution to this problem. The solution is a good approximation to the true MER receiver upon fulfillment of a certain, easily checkable, channel-related condition. The algorithms are derived first for the two-input many-output case. A generalized scheme is provided for the case of arbitrary number of inputs and outputs. The performance gain compared to that MMSE is evaluated for various correlated and Ricean channels and transmit power allocation strategies.

I. Introduction

Multiple-input multiple-output (MIMO) antenna systems have shown great potential in their ability to increase dramatically the data rate of wireless communications. Since pioneering work such as [1], [2], [3], progress has been made to try and realize the large capacity gains promised by MIMO information theory with practical transmit and receive architectures. Space-time coding [4] and spatial multiplexing [5] are the prominent techniques so far. In spatial multiplexing approaches such as BLAST[1] type architectures [6], one typically considers the transmission of multiple (N_t) signals over the N_t-transmitting antennas N_r-receiving antennas MIMO channel. The signals can be independent as in BLAST, in which case we assume $N_r \geq N_t$, or jointly encoded. Both linear and non-linear (eg. maximum likelihood (ML)) methods have been proposed in the literature to separate the spatially multiplexed inputs at the receiver. For complexity reduction reasons, there has been great interest in the class of linear receivers such as zero-forcing (ZF) and MMSE filters. Even in simplified non-linear methods such as the successive interference canceling (SIC) approach, used in V-BLAST [6], one exploits a linear receiver at each stage of detection. Existing linear techniques work well when the signatures of the various inputs are linearly separable. However, receivers such as MMSE and, to an even greater extent ZF

[1] Bell Labs Layered Architecture for Space-Time

have degraded error rate performance when attempting to invert a channel matrix that is ill-conditioned.

In practice moderately to severely ill-conditioned MIMO channels can arise due to a number of reasons, including fading correlation and/or the presence of a Ricean component. Antenna correlation occurs due to a lack of a rich scattering environment or limited path angle spread. High correlation levels (> 0.8) can also be obtained *by design* when building compact MIMO arrays with very small antenna spacing. Ricean channels (i.e. having a non-zero line-of-sight (LOS) component) also represent a serious challenge to spatial multiplexing algorithms. It has been shown that LOS part of the MIMO channel is usually extremely ill conditioned (see eg. [7]). Therefore the inversion of the overall MIMO matrix can lead to severe noise enhancement when the LOS part is significant.

Although the adverse effects of channel correlation (in a general sense) have been studied in terms of the MIMO capacity (see eg. [8], [9]) or algorithm performance degradation (e.g. [10]), it remains unclear how to best *address* the problem in terms of improved algorithm design. At any rate, the derivation of a class of linear receivers which are more robust to Ricean components and high correlation levels appears practically important and desirable.

In this paper we propose a new family of narrow-band spatial multiplexing linear receivers based on *error-rate minimization* rather than on mean-square noise or interference minimization. Notably, minimum-error-rate (MER) receivers have been investigated previously in areas such as channel equalization and CDMA multiuser detection (MUD). In particular, comparisons between MMSE and MER criteria are performed in [11], where it is shown that MER and MMSE detection provide similar results in most CDMA cases of interest. The results there, however, assume good code orthogonality properties between users. When that is not the case, and especially when combined with near-far effects, significant differences appear between MMSE and MER multiuser detectors as was noted in eg. [12]. Note however that unlike in CDMA where good code designs secures some minimum orthogonality between users, correlation between the MIMO inputs can be difficult to avoid for reasons mentioned above. Therefore we expect a MIMO receiver design based on MER to provide the needed robustness to this problem.

Unfortunately, general solutions to derive the MER filter in MUD are not really tractable due the complexity of the bit-error rate expression, even in the 2 user case. Several algorithms have been proposed in the recent past to tackle MER detection. All of these rely on adaptive descent/gradient approaches such as [12], [13] to cite a few examples. Although interesting, iterative/adaptive methods have convergence issues such as speed and local minima. In this paper we propose a first (up to the author's knowledge) closed-form algorithm based on an certain approximation of the error rate expression valid for the the case of two inputs and we give a strategy allowing to exploit the idea further for the case of arbitrary number of inputs. Clearly, our approach is not limited to MIMO reception and can, in turn, be used in MUD problems.

Contributions: We introduce the problem on MER-based linear receivers for MIMO spatial multiplexing systems. Then we present a strategy for MER receiver estimation based on expressing the receiver as a combination of ZF receivers. The MER receiver can be found via numerical optimization of the error rate expression. More interestingly however, this paper presents a simple strategy to obtain a closed-form solution to the approximate MER receiver estimation problem. We show the closed-form solution is realizable upon a certain condition that depends on the channel realization and that can be easily checked. For those channels for which the condition is not met, the algorithm simply falls back to a ZF or MMSE receiver. The solutions are presented in the $N_t = 2$ input case. Finally we present a scheme allowing to exploit the proposed techniques in the general case with arbitrary N_t. The algorithms are shown for a QPSK modulation but the principle can be readily extended to other QAM levels. The proposed approach for $N_t > 2$ is based on the idea of *cascaded receivers*, in which the input symbol is first estimated as part of a group of two symbols, and then finally extracted in the minimum error rate sense. Our simulations show that the proposed receivers surpass the MMSE receiver, with benefits becoming very significant in the case of correlated and/or Ricean channels together with unequal transmit power allocation. In particular, our simulations indicate that the MER approach helps get rid of near-error-flooring effects occurring with MMSE receivers at high SNR in Ricean channels. The unequal power scenario is consistent with a V-BLAST approach where streams decoded in early detection stages should carry more power than others to compensate for the lack of diversity in those stages, or else the performance is strongly limited by the first stream. We simulate various power allocation strategies based on channel correlation values allowing to optimize performance, however, the specific problem of optimal transmit power allocation is addressed elsewhere [14].

This paper is organized as follows: In Section II we present the MIMO signal and receiver model. In III we introduce the error rate expressions and approximations used in the rest of the paper. In IV we describe the construction of approximated closed-form MER receivers in the $N_t = 2$, arbitrary N_r, case. The extension to arbitrary N_t is shown in V. Sections VI illustrates the behavior of the proposed algorithms in correlated and Ricean channel scenarios, compared to MMSE receivers. We conclude in VII.

Notations: Throughout the paper, we adopt the following notations: $\mathbf{x}$ (lower-case, bold face): vectors. $\mathbf{X}$ (capitals, bold face): matrices. $\mathbf{X}^T$ and $\mathbf{X}^*$ denote the transpose and transpose conjugate respectively. $E(.)$ expectation operator. $\|.\|$ Frobenius norm of a vector or matrix. $Q(.)$ standard Q function. $\mathbf{I}_k$ identity of size $k \times k$. iid: identically independent distributed. diag(.) is the matrix with arguments on the main diagonal and zero elsewhere. Finally, $\mathbf{e}_n$ denotes a vertical vector with 1 at the n-th entry and zeros elsewhere, the size is determined from context.

II. MIMO signal and receiver model

We consider a narrow-band MIMO system shown in Fig. 1, with N_t inputs and N_r outputs. Input signals stacked in a vector notation as $\mathbf{s} = [s_1, .., s_{N_t}]^T$. Because we focus on spatial multiplexing systems, we assume the signals to be statistically independent rather than jointly encoded. However some level of joint (space-time) encoding is admissible provided that the second-order signal decorrelation between signals ($E(s_k s_l^*) = 0, k \neq l$) is preserved. We normalize the input power so that ($E|s_k|^2 = 1$). At the receiver, we observe the vector $\mathbf{y} = [y_1, y_2, .., y_{N_r}]^T$ given by:

$$\mathbf{y} = \mathbf{H}\mathbf{s} + \mathbf{n} \tag{1}$$

$$\mathbf{H} = \mathbf{H}_0 \Lambda \tag{2}$$

where $\mathbf{H}$ is the $N_r \times N_t$ MIMO flat-fading channel matrix. $\mathbf{H}_0$ is a matrix with, possibly correlated, complex Gaussian entries of unit power. A non-zero mean can be used to obtain a Ricean channel (see Section VI). $\mathbf{n}$ is a white Gaussian noise vector, with iid entries of variance σ_n^2. Λ is the transmit power normalization matrix:

$$\Lambda = \text{diag}[\sqrt{p_1}, \sqrt{p_2}, .., \sqrt{p_{N_t}}], \quad \sum_{k=1}^{N_t} p_k = 1 \tag{3}$$

where p_i is the power alloted to the i-th input. Because we focus on the narrow-band case, the system has no memory and we have omitted the time index in both the channels and signals. At the receive side, we consider filters in the class of linear receivers and SIC receivers. Note that also in the case of SIC receivers, each stage takes typically the form of a linear detection (followed by a decision and subtraction).

A. Arbitrary linear detectors

We consider the linear estimation of an arbitrary symbol s_k, from $\mathbf{y}$, using the $N_r \times 1$ vector filter $\mathbf{w}_k$:

$$\mathbf{w}_k^* \mathbf{y} = \hat{s}_k \tag{4}$$

where $\hat{s}_k$ is an estimate of s_k in any given sense. With this notation, and for a known channel, the ZF and MMSE receivers are given respectively by:

$$\begin{aligned}
\mathbf{w}_k^{zf} &= (\mathbf{e}_k^T \mathbf{H}^{\#})^* \quad \textbf{(ZF)} && (5)\\
\mathbf{w}_k^{mmse} &= \mathbf{R}_y^{-1} \mathbf{r}_{ys_k} \quad \textbf{(MMSE)} && (6)\\
\text{where } \mathbf{R}_y &= E(\mathbf{y}\mathbf{y}^*) && (7)\\
\text{and } \mathbf{r}_{ys_k} &= E(\mathbf{y}s_k^*) && (8)
\end{aligned}$$

where $\mathbf{H}^{\#}$ is the pseudo-inverse of $\mathbf{H}$. Since the input symbols are decorrelated and independent from the noise, we have the standard results:

$$\begin{aligned}
\mathbf{R}_y &= \mathbf{H}\mathbf{H}^* + \mathbf{R}_n && (9)\\
\mathbf{r}_{ys_k} &= \mathbf{H}\mathbf{e}_k && (10)
\end{aligned}$$

where $\mathbf{R}_n$ is the noise covariance, here we assume $\mathbf{R}_n = \sigma_n^2 \mathbf{I}_{N_r}$. By design, the ZF receiver is orthogonal to the signatures of all inputs except that of s_k and thus cancels out all interference while the MMSE receiver lets a small amount of interference in the interest of minimizing the mean-square error. In general the combined response of an arbitrary linear receiver $\mathbf{w}_k$ is given by:

$$\mathbf{w}_k^* \mathbf{H} = [\alpha_1, \alpha_2, .., \alpha_{N_t}] \tag{11}$$

where the vector $[\alpha_1, .., \alpha_{k-1}, \alpha_{k+1}, .., \alpha_{N_t}]^T$ constitutes the *residual* response of the receiver.

A.1 Re-parameterizing linear MIMO detectors

In this section, we show that linear detectors of interest can be expressed uniquely in terms of a linear combination of ZF detectors. This re-parameterization turns out extremely useful when optimizing the error rate expression. Without loss of generality, we can normalize $\mathbf{w}_k$ by α_k^* so that we adopt the following notation:

$$\mathbf{w}_k^* \mathbf{H} = \mathbf{t}^T = [\alpha_1, .., \alpha_{k-1}, 1, \alpha_{k+1}, .., \alpha_{N_t}] \tag{12}$$

Clearly the residual for the ZF receiver is zero. Given the *combined response* $\mathbf{t}$ in (12) it is possible to express the corresponding *minimum norm*[2] receiver $\mathbf{w}_k$ in the form of a linear combination of ZF receivers:

$$\mathbf{w}_k = \mathbf{H}^{\#^*} \mathrm{conj}(\mathbf{t}) = \mathbf{w}_k^{zf} + \sum_{l \neq k} \mathbf{w}_l^{zf} \alpha_l^* \tag{13}$$

where conj means complex conjugation. We choose to optimize the receiver in terms of the coefficients of $\mathbf{t}$. Note that, when optimizing the normalized receiver in the MMSE sense, it appears that one selects $\mathbf{t}$ such that:

$$\mathbf{t}_{mmse} = \mathrm{argmin}_{\mathbf{t}} \{ (\sum_{l \neq k} |\alpha_l|^2 + \sigma_n^2 \| \mathbf{w}_k^{zf} + \sum_{l \neq k} \mathbf{w}_l^{zf} \alpha_l^* \|^2) \} \tag{14}$$

[2]The min norm receiver is defined by the solution $\mathbf{w}_k$ to (12) such that $\|\mathbf{w}_k\|$ is minimum. We restrict ourselves to those solutions because they can be easily shown to minimize noise amplification among all other solutions to (12).

while the conventional (non normalized) MMSE receiver in (6) has its combined response optimizing:

$$\text{argmin}_{\mathbf{t}}\{|\alpha_k - 1|^2 + (\sum_{l\neq k}|\alpha_l|^2 + \sigma_n^2\|\mathbf{w}_k^{zf} + \sum_{l\neq k}\mathbf{w}_l^{zf}\alpha_l^*\|^2)\} \tag{15}$$

III. Error rate performance

The MMSE receiver is doing the best compromising between noise amplification and noise enhancement by selecting an appropriate linear combination of ZF receivers. However, it does not so in view of error-rate minimization but merely in order to minimize the total interference plus noise power. To closely approach the behavior of the MER receiver, the residual noise plus interference should be Gaussian distributed. This assumption is acceptable in MUD with a large number of users but not in typical MIMO antenna systems. In fact the implicit Gaussian assumption made in the MMSE receiver is a pessimistic one, since Gaussian noise is generally known to be the worst possible additive noise [15].

The analysis of error rate performance of linear MIMO receivers has been initiated in the case of MUD for CDMA signals in eg. [11]. It is very difficult (or impossible) to optimize the general error-rate expressions in closed-form though. The strategy that we adopt is to two-fold:
1. Address the problem in the $N_t = 2$ case where optimization is simplest and, following remarks above, MER receivers have the largest advantage over MMSE ones. Unfortunately there is no available closed-form solution for a MER receiver in that case either. However, by using the equivalent representation of $\mathbf{w}_k$ in (13) and a judicious approximation, we are able to propose a simple solution that mimics the behavior of the true MER receiver.
2. In the $N_t > 2$ case, break the problem down into a set of simpler two-input problems, where the above solutions apply, following a philosophy reminiscent of [16].

A. Error-rate for $N_t = 2$

Now we consider $N_t = 2$ inputs and arbitrary $N_r \geq 2$ outputs. Without loss of generality we consider the estimation the first input s_1 from the $N_r \times 1$ receiver $\mathbf{w} = \mathbf{w}_1$ (index 1 is dropped below). From (13), we write our receiver as:

$$\mathbf{w} = \mathbf{w}_1^{zf} + \mathbf{w}_2^{zf}\alpha_2^* = \mathbf{w}_1^{zf} + \mathbf{w}_2^{zf}\rho e^{-j\phi} \tag{16}$$

where α_2 is expressed in terms of its amplitude and phase by $\alpha_2 = \rho e^{j\phi}$. At the output of $\mathbf{w}$ in (16) we observe:

$$\hat{s}_1 = \mathbf{w}^*\mathbf{y} = s_1 + \rho e^{j\phi} s_2 + \mathbf{w}^*\mathbf{n} \tag{17}$$

The output of the receiver can be seen as the superposition of the original constellation with another scaled down and rotated version of it. ρ measures the scale of the interfering constellation (clearly $\rho < 1$ is required) and ϕ its rotation angle. This is illustrated in Fig.2 in the case of QPSK signals.

We now derive the probability $P(\hat{s}_1 \rightarrow \tilde{s}_1)$ of deciding in favor of a wrong symbol $\tilde{s}_1$. We use a standard technique in assuming that $\tilde{s}_1$ is confined to be a nearest neighbor of s_1. Making use of elementary geometry in Fig.2, we extend classical results of [17] to find, for QPSK signals in white Gaussian noise:

$$\begin{aligned} P_e(\rho,\phi) = P(\hat{s}_1 \rightarrow \tilde{s}_1) \;&=\; \frac{1}{2}\left[Q\left(\frac{1-\sqrt{2}\rho\sin(\phi+\frac{\pi}{4})}{\sigma_n\|\mathbf{w}\|_{\rho,\phi}}\right) + Q\left(\frac{1-\sqrt{2}\rho\cos(\phi+\frac{\pi}{4})}{\sigma_n\|\mathbf{w}\|_{\rho,\phi}}\right)\right. \\ &+\; \left. Q\left(\frac{1+\sqrt{2}\rho\sin(\phi+\frac{\pi}{4})}{\sigma_n\|\mathbf{w}\|_{\rho,\phi}}\right) + Q\left(\frac{1+\sqrt{2}\rho\cos(\phi+\frac{\pi}{4})}{\sigma_n\|\mathbf{w}\|_{\rho,\phi}}\right)\right] \end{aligned} \tag{18}$$

where $\|\mathbf{w}\|_{\rho,\phi} = \|\mathbf{w}_1^{zf} + \rho e^{-j\phi}\mathbf{w}_2^{zf}\|$. The MER receiver can now be defined by $(\rho_{opt}, \phi_{opt}) = \text{argmin}_{\rho,\phi}\{P_e(\rho,\phi)\}$. The expression above does not admit closed-form minima to our knowledge. It can be optimized numerically for ρ, ϕ but it may be difficult to avoid local minima in the present form however. In the following we present a way to tackle this problem based on a leading-term approximation of (18). Although the presentation is made for QPSK signals, a similar approach can be used to deal with other QAM modulation with regular grids such as binary, 16-QAM, 32 QAM etc.

IV. Approximate Minimum-error based receivers

To simplify the optimization for interference amplitude ρ and phase ϕ, we follow the strategy below:

1. We solve for ρ and ϕ separately. Clearly relaxation-type approaches could be used here, however we found our strategy to provide a simple closed-form algorithm with good performance. The main explanation for this is that we find the role of ϕ to be mainly limited to the minimization of the noise enhancement factor $\sigma_n\|\mathbf{w}\|_{\rho,\phi}$ in (18), whose solutions turn out to be independent of ρ, as we will see.

2. We ignore cases where the optimal amplitude ρ_{opt} returned by the closed-form algorithm is not significantly larger than zero (when that happens we simply pick the ZF or, better, MMSE receiver instead). The intuition for this is two-fold: first, the case when ρ is close to zero (defined in an algorithmic sense via $\rho < \rho_{min}$, where ρ_{min} is a chosen threshold) is the case when the computation of the MER receiver (over, say, the ZF receiver) is least justified, since for $\rho \rightarrow 0$, $\mathbf{w} \rightarrow \mathbf{w}^{zf}$. Second, when ρ is significantly larger than 0 ($\rho \geq \rho_{min}$ in the algorithm below), which is the most meaningful case for a MER approach, considerable simplifications of the error-rate expression are obtained.

We now describe the closed-form optimization. For exposition purposes, we present the optimization of ρ first, however, in practice, the algorithm will estimate the phase ϕ first.

A. Optimization of ρ

Assume $\rho > \rho_{min}$: we define $\xi_\phi = \max\{|\sin(\phi + \frac{\pi}{4})|, |\cos(\phi + \frac{\pi}{4})|\}$. We now use the exponential decay property of the Q function and we approximate the probability of error by its leading term. Indeed, for an appropriate choice of the threshold ρ_{min}, we have:

$$P_e \approx \hat{P}_e = \frac{1}{2} Q(\frac{1 - \sqrt{2}\rho\xi_\phi}{\sigma_n \|\mathbf{w}\|_{\rho,\phi}}) \tag{19}$$

Remarks on the approximation. We note that the approximation above will be poor in the case when $\phi \approx n\pi/2$. Since for that case $|\sin(\phi + \frac{\pi}{4})| \approx |\cos(\phi + \frac{\pi}{4})| \approx 1/\sqrt{2}$, and we obtain for that special case:

$$P_e(\phi \approx n\pi/2) \approx Q(\frac{1 - \rho}{\sigma_n \|\mathbf{w}\|_{\rho,\phi}}) \tag{20}$$

Interestingly, however, the expressions in (20) and (19) only differ by a constant factor 1/2 when $\phi \approx n\pi/2$. Therefore we argue that it is reasonable to use (19) for the optimization of ρ for any arbitrary phase ϕ. This fact is confirmed by our simulations.

To minimize the error rate in (19) is equivalent to maximizing the argument of the Q function in its positive region. Equivalently we select the optimal ρ_{opt} in $[0,\ 1/(\sqrt{2}\xi_\phi)]$ such that:

$$\begin{aligned}\rho_{opt} &= \text{argmax}_{\rho}\left(\frac{1-\sqrt{2}\rho\xi_{\phi}}{\sigma_n\|\mathbf{w}\|_{\rho,\phi}}\right)^2 \\ &= \text{argmax}_{\rho}J(\rho) = \text{argmax}_{\rho}\frac{(1-\sqrt{2}\rho\xi_{\phi})^2}{\|\mathbf{w}_1^{zf}+\rho e^{-j\phi}\mathbf{w}_2^{zf}\|^2} \qquad (21)\end{aligned}$$

Setting the gradient of $J(\rho)$ to 0 and eliminating the minimum point in $\rho = 1/(\sqrt{2}\xi_{\phi})$, we easily find an equation which admits one *single* solution, given by:

$$\rho_{opt} = -\frac{\eta+\sqrt{2}\xi_{\phi}\|\mathbf{w}_1^{zf}\|^2}{\sqrt{2}\xi_{\phi}\eta+\|\mathbf{w}_2^{zf}\|^2} \qquad (22)$$

$$\text{where } \eta = \text{Real}(\mathbf{w}_1^{zf*}\mathbf{w}_2^{zf}e^{-j\phi}) \qquad (23)$$

Validity of the closed-form solution: The closed-form solution above remains valid as long as the optimal ρ returned by (22) lies in the interval of positivity of $1-\sqrt{2}\rho\xi_{\phi}$. In addition, the algorithm must check that $\rho_{opt} > \rho_{min}$ to justify our approximation of P_e. Our simulations indicate that these conditions are most often satisfied for a proper choice of ρ_{min} (typically $\rho_{min} \approx 0.1$). This condition is the price for replacing minimization of the error-rate in (19) by the well-behaved cost in (21). Note that for those channels for which ρ_{opt} falls outside the validity interval, then the MER receiver estimation may not have a simple closed-form solution. In that case, a possible strategy is to fall back to a MMSE receiver instead. Our experience is this does not result in a significant performance loss in practice, due to the fact the gain brought by the MER receiver over the MMSE one is obtained mostly from those cases when $\rho \in [\rho_{min}, 1/(\sqrt{2}\xi_{\phi})]$.

B. Optimization of phase ϕ

To optimize the phase ϕ in closed-form from (19) independently of ρ is not possible. However we make the following argument from (19):

$$\frac{1}{2}Q(\frac{1-\rho}{\sigma_n\|\mathbf{w}\|_{\rho,\phi}}) \leq \hat{P}_e \leq \frac{1}{2}Q(\frac{1-\sqrt{2}\rho}{\sigma_n\|\mathbf{w}\|_{\rho,\phi}}) \qquad (24)$$

Here it is possible to minimize *both* the upper bound and lower bound of $\hat{P}_e$ by selecting ϕ_{opt} such that

$$\begin{aligned}\phi_{opt} &= \text{argmax}_{\phi}\{\frac{1-\rho}{\sigma_n\|\mathbf{w}\|_{\rho,\phi}}\} \text{ or equivalently} \\ \phi_{opt} &= \text{argmin}_{\phi}\{\|\mathbf{w}\|^2_{\rho,\phi}\} = \text{argmin}_{\phi}\|\mathbf{w}_1^{zf} + \rho e^{-j\phi}\mathbf{w}_2^{zf}\|^2 \end{aligned} \quad (25)$$

The solution to (25) is obtained from a standard gradient approach and is given by:

$$e^{j\phi_{opt}} = -\frac{\mathbf{w}_1^{zf^*}\mathbf{w}_2^{zf}}{|\mathbf{w}_1^{zf^*}\mathbf{w}_2^{zf}|} \quad (26)$$

Note that, as desired, this solution *does not* depend on ρ. Thus it can be carried out first and independently of ρ.

C. Summary of algorithm for $N_t = 2$

The approximate MER receiver algorithm goes as follows, given selected threshold ρ_{min} and given channel $\mathbf{H}$:

1. Compute ZF receivers $\mathbf{w}_1^{zf}, \mathbf{w}_2^{zf}$ from (5)
2. Compute ϕ_{opt} from (26).
3. Compute ρ_{opt} from (22).
4. If $\rho_{opt} \in [\rho_{min}, 1/(\sqrt{2}\xi_{\phi})]$, compute $\mathbf{w} = \mathbf{w}_1^{zf} + \rho_{opt}e^{-j\phi_{opt}}\mathbf{w}_2^{zf}$. If not, compute MMSE receiver $\mathbf{w}^{mmse}$ from (6), let $\mathbf{w} = \mathbf{w}^{mmse}$.

Once signal s_1 is estimated, the second signal s_2 can be obtained by following a similar procedure, or by subtracting the contribution s_1, followed by maximum ratio combining, in the standard SIC/V-BLAST fashion [6].

V. Receiver algorithms for arbitrary N_t

Error rate expressions for linear receivers and $N_t > 2$ are available, however their optimization in closed-form is difficult (or impossible) up to our knowledge. To extend the closed-form algorithm to arbitrary N_t, we propose an approach that reduces the N_t-input problem to a 2-input problem where the closed-form MER-based receiver can be used. Importantly, this technique is also consistent with exploiting the MER-based receiver where it yields the largest advantage over other known linear receivers, i.e. the case of two strongly correlated signals.

The algorithm consists in two stages for the estimation of the k-th input s_k, $k = 1..N_t$, which is reminiscent of the approach used for simplified space-time decoding in [16]. In the first stage, we select one other input index l (along with k) and use a *group* receiver to estimate the contribution of the *pair* of inputs (s_k, s_l) in the MMSE sense. In the second step, we design a 2-input MER-based receiver to estimate s_k from the output of the group receiver. The MER-based receiver is slightly modified to take into account colored noise statistics at the output of the group receiver, but otherwise follows the design shown in the previous section.

Importantly, the index l in the group (s_k, s_l) can be selected so as to minimize the probability of error of s_k, as shown in Sec.V-E. The 2-stage decomposition is illustrated in Fig.3.

A. Optimality of a two-stage MMSE receiver

Interestingly, the optimality of the decomposition into two cascaded receivers, where each receiver is MMSE-based, can be demonstrated through the result below:

Lemma 1. *Let input signal indexes $k, l \in [0, .., N_t], k \neq l$, be arbitrary. Let $\mathbf{w}_k^{mmse}$ be the MMSE receiver for s_k as in (6). Let $\mathbf{s}_{kl} = [s_k, s_l]^T$. Let $\mathbf{H}_{kl} = [\mathbf{H}(:, k), \mathbf{H}(:, l)]$ be the submatrix of $\mathbf{H}$ only consisting of the signatures of s_k, s_l. We define $\mathbf{W}_{kl}$, the $N_r \times N_r$ MMSE* group *receiver matrix for s_k, s_l, by:*

$$\mathbf{W}_{kl} = \text{argmin}_{\mathbf{W}} \{E\|\mathbf{W}^*\mathbf{y} - \mathbf{H}_{kl}\mathbf{s}_{kl}\|^2\} \tag{27}$$

Finally, let $\mathbf{w}_{kl\to k}$ be the $N_r \times 1$ receiver allowing the estimation of s_k from $\mathbf{z}_{kl} = \mathbf{W}_{kl}^*\mathbf{y}$ in the MMSE sense. Then $\mathbf{w}_k^{mmse}$ is equal to the cascaded filters $\mathbf{W}_{kl}$,$\mathbf{w}_{kl\to k}$:

$$\mathbf{w}_k^{mmse} = \mathbf{W}_{kl}\mathbf{w}_{kl\to k} \tag{28}$$

Proof. See Appendix 1. □

The result above shows that nothing is lost in decomposing a MMSE receiver into the product of a $N_r \times N_r$ MMSE *group* receiver for symbols s_k, s_l and a $N_r \times 1$ MMSE receiver for symbol s_k.

In fact the result above is valid for any group size $\leq N_t$. However, when the group size is two, the second stage receiver can be replaced by a MER-based receiver similar to that developed in the previous section. Only, colored noise statistics must now be taken into consideration. We now describe the construction of the cascaded receiver.

B. MMSE group receiver

We're interested in detecting input s_k, where k is arbitrary $k = 1, .., N_t$. The MMSE group receiver for (s_k, s_l), where s_l is another input distinct from s_k, is defined in (27). From standard derivative analysis, the solution to this problem is obtained from:

$$\mathbf{R}_y \mathbf{W}_{kl} = \mathbf{H}_{kl} \mathbf{H}_{kl}^* \tag{29}$$

which can be seen as direct generalization of the usual case when the group size is one (such as (6)). At the output of the group receiver, we observe:

$$\mathbf{z}_{kl} = \mathbf{W}_{kl}^* \mathbf{y} = \mathbf{W}_{kl}^* \mathbf{H}_{kl} \mathbf{s}_{kl} + \mathbf{W}_{kl}^* \tilde{\mathbf{H}}_{kl} \tilde{\mathbf{s}}_{kl} + \mathbf{W}_{kl}^* \mathbf{n} \tag{30}$$

where $\tilde{\mathbf{H}}_{kl}$ is the $N_r \times (N_t - 2)$ submatrix of $\mathbf{H}$ consisting of the signatures of s_i, $i \neq k, l$. Correspondingly, $\tilde{\mathbf{s}}_{kl}$ is obtained by removing entries s_k, s_l from $\mathbf{s}$.

B.1 Noise covariance

We merge all terms of (30), except the contribution of desired inputs s_k, s_l, into a single composite noise term. The covariance of this composite noise at the output of the group receiver, denoted by $\mathbf{R}_{\bar{n}}$, is no longer identity, yet is obtained easily:

$$\mathbf{R}_{\bar{n}} = \mathbf{W}_{kl}^* (\tilde{\mathbf{H}}_{kl} \tilde{\mathbf{H}}_{kl}^* + \sigma_n^2 \mathbf{I}_{N_r}) \mathbf{W}_{kl} \tag{31}$$

C. MER-based receiver

The MER-based is applied at the output $\mathbf{z}_{kl}$ of the group receiver. For convenience, we rewrite (30) into

$$\mathbf{z}_{kl} = \mathbf{G}_{kl} \mathbf{s}_{kl} + \tilde{\mathbf{G}}_{kl} \tilde{\mathbf{s}}_{kl} + \mathbf{W}_{kl}^* \mathbf{n} \tag{32}$$

where $\mathbf{G}_{kl} = \mathbf{W}_{kl}^* \mathbf{H}_{kl}$ and $\tilde{\mathbf{G}}_{kl} = \mathbf{W}_{kl}^* \tilde{\mathbf{H}}_{kl}$.

Note that this time we can assume the interference term $\tilde{\mathbf{G}}_{kl}\tilde{\mathbf{s}}_{kl}$ to be close to Gaussian, this approximation being well justified when the number of antennas N_t becomes large. It is then possible to exploit the two-input MER-based receiver developed in Section IV, with slight modifications since the channel matrix $\mathbf{H}$ is replaced by the *equivalent* channel $\mathbf{G}_{kl}$ here, and the additive noise is no longer white.

Let $\mathbf{w}_k^{mer}$ be a $N_r \times 1$ vector receiver acting on $\mathbf{z}_{kl}$. Let $\mathbf{w}_{Gk}^{zf}$ and $\mathbf{w}_{Gl}^{zf}$ the ZF filters such that $\mathbf{w}_{Gk}^{zf*}\mathbf{G}_{kl} = [1, 0]$, and $\mathbf{w}_{Gl}^{zf*}\mathbf{G}_{kl} = [0, 1]$. Then, similarly to results in (26), (22), the MER-based receiver is found from an appropriate linear combination of $\mathbf{w}_{Gk}^{zf}$ and $\mathbf{w}_{Gl}^{zf}$:

$$\mathbf{w}_k^{mer} = \mathbf{w}_{Gk}^{zf} + \rho_{opt} e^{-j\phi_{opt}} \mathbf{w}_{Gl}^{zf} \tag{33}$$

C.1 Phase optimization

As earlier, the phase is found from minimizing the (here colored) noise power $\sigma_n^2 \|\mathbf{w}_k^{mer}\|^2_{\mathbf{R}_{\bar{n}}} = \sigma_n^2 \mathbf{w}_k^{mer*}\mathbf{R}_{\bar{n}}\mathbf{w}_k^{mer}$. We find for this:

$$e^{j\phi_{opt}} = -\frac{\mathbf{w}_{Gk}^{zf*}\mathbf{R}_{\bar{n}}\mathbf{w}_{Gl}^{zf}}{|\mathbf{w}_{Gk}^{zf*}\mathbf{R}_{\bar{n}}\mathbf{w}_{Gl}^{zf}|} \tag{34}$$

C.2 Amplitude optimization

Defining again $\xi = \max\{|\sin(\phi_{opt} + \frac{\pi}{4})|, |\cos(\phi_{opt} + \frac{\pi}{4})|\}$, the amplitude is found by extending (19) to the colored noise case and minimizing:

$$\rho_{opt} = \text{argmin}_{\rho} \frac{1}{2} Q\left(\frac{1 - \sqrt{2}\rho\xi}{\sigma_n \sqrt{\mathbf{w}_k^{mer*}\mathbf{R}_{\bar{n}}\mathbf{w}_k^{mer}}}\right) \tag{35}$$

Following the same strategy as in Sec. IV, a closed-form optimum for (35), subject to the condition on the interval of validity, can be found by

$$\rho_{opt} = -\frac{\eta + \sqrt{2}\xi \mathbf{w}_{Gk}^{zf*}\mathbf{R}_{\bar{n}}\mathbf{w}_{Gk}^{zf}}{\sqrt{2}\xi\eta + \mathbf{w}_{Gl}^{zf*}\mathbf{R}_{\bar{n}}\mathbf{w}_{Gl}^{zf}} \tag{36}$$

$$\text{with } \eta = \text{Real}(\mathbf{w}_{Gk}^{zf*}\mathbf{R}_{\bar{n}}\mathbf{w}_{Gl}^{zf} e^{-j\phi_{opt}}) \tag{37}$$

D. Algorithm summary for $N_t > 2$

To estimate input s_k, we proceed as follows. given selected threshold ρ_{min}, channel $\mathbf{H}$, selected index l:

1. Compute $\mathbf{W}_{kl}$ from (29) and $\mathbf{R}_{\bar{n}}$ from (31)
2. Compute ZF receivers $\mathbf{w}_{Gl}^{zf}, \mathbf{w}_{Gk}^{zf}$ from $\mathbf{G}_{kl}$
3. Compute ϕ_{opt} from (34)
4. Compute ρ_{opt} from (36)
5. If $\rho_{opt} \in [\rho_{min}, 1/(\sqrt{2}\xi_\phi)]$, compute $\mathbf{w}_k^{mer}$ from (33). Then let $\mathbf{w}_k = \mathbf{W}_{kl}\mathbf{w}_k^{mer}$. If not, compute the MMSE receiver $\mathbf{w}_k^{mmse}$ from (6), let $\mathbf{w}_k = \mathbf{w}_k^{mmse}$.

E. Selection of optimum group

Note that, in the first stage (Sec.V-B) the group receiver does not attempt to separate the inputs selected in the group (or, here, pair) (s_k, s_l), as this task is let to the MER-based receiver used in the second stage. Because the MER receiver is more robust to correlated inputs than the MMSE receiver, a reasonable strategy in the first stage is, for any given input s_k, to select input s_l, $l \neq k$, such that the signatures of s_k, s_l are simultaneously the 'most' correlated with each other and the 'most' linearly separable as a group from all other signatures, so that both the MMSE and MER stages give best performance. This approach is also consistent with practical arrays, where, for instance, pairs of neighboring antennas tend to exhibit higher correlation than more distant antennas in the array.

The optimum way to realize this is to select $l = l_k$ so as to optimize the error rate performance for the detection of s_k. For a given l, the error rate can be again approximated and l_k found by:

$$l_k = \text{argmin}_l \hat{P}_e(l) = \text{argmin}_l \frac{1}{2} Q\left(\frac{1-\sqrt{2}\rho_{opt}\xi}{\sigma_n\sqrt{\mathbf{w}_k^{mer*}\mathbf{R}_{\bar{n}}\mathbf{w}_k^{mer}}}\right) \tag{38}$$

where the dependence in l is implicit in the expression above. It is also possible to use the more complex expression found in (18), modified according to the colored noise's covariance, to get more accurate results. In our simulations we limit ourselves to the approach shown in (38).

F. Detection of remaining inputs

Once s_k has been estimated, other inputs may be obtained following the same procedure. For better performance however, the contribution of s_k may be subtracted first before estimation of s_{k+1}. This procedure can be iterated in way analogous to V-BLAST.

VI. Simulations

We test the error-rate performance of the proposed algorithms (with $\rho_{min} = 0.1$) in the case of MIMO channels with severe correlation and/or significant Ricean components. All plots including averaging over 1000 Monte-Carlo runs each featuring one independent channel realization.

The realization for a (transmit-) correlated Ricean MIMO channel is obtained through a channel matrix of the form

$$\mathbf{H}_0 = \sqrt{\frac{\beta}{1+\beta}}\mathbf{H}_{los} + \sqrt{\frac{1}{1+\beta}}\mathbf{H}_r\mathbf{R}^{1/2} \tag{39}$$

where $\mathbf{R}$ is the standard correlation matrix (here for transmit side), $\mathbf{H}_r$ is zero mean i.i.d. complex Gaussian, $\mathbf{H}_{los}$ is the LOS channel matrix and β is the Ricean factor. The LOS matrix is built simply from specifying antenna positions of (broadside) linear arrays at transmit and receive and the distance between transmitter and receiver, see for instance [7]. We choose a compact antenna spacing of 0.3 wavelengths and a propagation distance of 1000 wavelengths. This results in condition number of about 3000 for $\mathbf{H}_{los}$.

We consider two cases. In the first one we take $N_t = 2$, $N_r = 4$ and compare the MER-based algorithm of Sec. IV to a MMSE receiver. We assume perfect channel knowledge. The antenna correlation is 0.9 (this corresponds to a severe lack of angular spread or closely spaced antennas) [7]. The Ricean factor is 8dB, within the typical range for suburban MIMO channels [18]. In Fig. 4 we plot the symbol error rate for the first input (the order is pre-determined randomly). We consider several power allocation strategies between the first and second input, with the first input having a power ratio of 0dB, 3dB, 5dB, 7dB, respectively, above the second input. The second input is detected after subtraction of the first one in a SIC fashion. Unequal power allocation can occurs for e.g. by design to compensate for lack of diversity in the first input, or by propagation differences when the input do not originate from the same physical transmitter (as in space division multiple access). We do not attempt to optimize the power allocation as this separate problem is treated in [14]. However, we also plot the performance on the second input to test the impact of the power allocation.

We found the closed-form condition to be met in about 90% cases, with some variation depending on the power allocation. The closed-form MER-based always outperform the MMSE receiver. The difference is small when we use equal power allocation, which confirms the predictions of [11], [12], but it becomes extremely significant for unequal powers (equivalent to near-far effects in MUD). From Figs. 4, 5, we see that a power ratio of up to 7 dB improves on the performance of *both* the first *and* second input, regardless of the receiver used. That is because although the second gets less power than the first, it benefits from a better error rate in s_1 in the SIC algorithm. This effects reverses for more than 7dB of power ratio (not shown here). We also see that the MER approach helps suppress the a "near error flooring" phenomenon occurring at high SNR with MMSE for both inputs. This slow decreasing of the error rate for MMSE is due to the high antenna correlation and the irreducible interference coming from ill-conditioned LOS component, which causes the curve to differ dramatically from an ideal $4-2+1=3$-order diversity performance.

In Fig.6, we compare for the same channel parameters as above and for a power ratio of 5dB, the closed form approximate MER detector with an "exact" MER detector. As done in previous papers such as [12], the exact MER receiver has no closed form solution and is obtained by e.g. running a two-variable gradient descent on the exact error rate expression (18) for every channel realization. The results show that little is lost in the approximations used in (26) and (22).

Let us stress here that the closed form algorithm is derived independently of the power allocation. However it is known from MUD theory that the *exact* MER detector outperforms the MMSE receiver significantly (only) in the case of unequal power allocation, which makes this latter case the interesting one to study for our algorithm.

In the second case, we take $N_t = 3$, $N_r = 6$ and use the algorithm shown in Sec. V. We plot the error performance in Fig. 7 for the same channel conditions as earlier: Correlation is 0.9 between neighboring antennas and decays exponentially after that (0.81,0.73..). The power allocation specifies the power ratio between input k and $k+1$, $k = 1, 2$. Once again, the MER-based receiver outperforms the MMSE receiver in all cases.

Finally we measure the increased robustness of the MER-based receiver to various levels of correlation (with Ricean factor 0) and Ricean factors (with correlation 0) in the 2×4

case, in Fig. 8 and 9 respectively. The SNR is fixed at 15dB in this case. Unlike for MMSE, the performance keeps improving with β for the MER-based solution and power ratio 7dB. Once again the MER-based receiver is helpful, with the greatest impact on Ricean channels.

VII. Conclusions

This paper presents new algorithms for the construction of approximate minimum-error-rate linear MIMO receivers. The key contribution is a simple closed-form solution to this problem, shown for the $N_t = 2$ case. In the general case of N_t we extend the method using cascaded MMSE and MER-based receivers. The MER-based receiver is applied on selected pairs of inputs so as to minimize the probability of error. The advantage of the proposed approach over existing linear MMSE-type receivers is a greater robustness with respect to ill-conditioned MIMO channel matrix arising in eg. correlated fading scenarios or Ricean environments. The advantage of the MER approach is increased in the case of unequal input power allocation, as predicted from MUD theory.

Acknowledgments

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Appendix:Proof of Lemma 1

The second stage receiver $\mathbf{w}_{kl\to k}$ is defined by:

$$\mathbf{w}_{kl\to k} = \text{argmin}_{\mathbf{w}}\{E|\mathbf{w}^*\mathbf{W}_{kl}^*\mathbf{y} - s_1|^2\} \tag{40}$$

hence can is obtained from :

$$(\mathbf{W}_{kl}^*\mathbf{R}_y\mathbf{W}_{kl})\mathbf{w}_{kl\to k} = \mathbf{W}_{kl}^*\mathbf{h}_k \tag{41}$$

where $\mathbf{h}_k$ is the k-th column of $\mathbf{H}$ Therefore we have:

$$\mathbf{W}_{kl}^*(\mathbf{R}_y\mathbf{W}_{kl}\mathbf{w}_{kl\to k} - \mathbf{h}_k) = 0 \tag{42}$$

Let us denote $\mathbf{x} = \mathbf{R}_y\mathbf{W}_{kl}\mathbf{w}_{kl\to k} - \mathbf{h}_k$ and show it is 0. Then this will prove our result since $\mathbf{W}_{kl}\mathbf{w}_{kl\to k}$ will be equal to the single stage MMSE receiver.

We have also $\mathbf{x} = \mathbf{H}_{kl}\mathbf{H}_{kl}^*\mathbf{w}_{kl\to k} - \mathbf{h}_k$ which shows that $\mathbf{x}$ lies in the span of $\mathbf{H}_{kl}$. Then, let us denote the 2×1 vector $\mathbf{u}$ such that $\mathbf{x} = \mathbf{H}_{kl}\mathbf{u}$. From (42),(29), we get:

$$\mathbf{W}_{kl}^*\mathbf{H}_{kl}\mathbf{u} = 0 = \mathbf{R}_y^{-1}\mathbf{H}_{kl}\mathbf{H}_{kl}^*\mathbf{H}_{kl}\mathbf{u} \tag{43}$$

We can assume $\mathbf{H}_{kl}$ is of full rank 2 (happens with probability one). Then $\mathbf{H}_{kl}^*\mathbf{H}_{kl}$ has full rank 2. Since $\mathbf{R}_y^{-1}$ also has full rank, then the only solution above is $\mathbf{u} = 0$, hence $\mathbf{x} = 0$.

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LIST OF FIGURE CAPTIONS:

- Fig. 1. Diagram of a MIMO spatial multiplexing system.
- Fig. 2. Superposed QPSK constellations at the output of linear MIMO receiver with $N_t = 2$. The interfering constellation has amplitude ρ and phase ϕ.
- Fig. 3. Diagram of cascaded receiver for source s_k: a group MMSE receiver extracts the contribution of s_k, s_l. A MER receiver is then applied to estimate s_k.
- Fig. 4. Error rate in 2×4 case for first input. Correlation is 0.9. Ricean factor is 8dB.
- Fig. 5. Error rate in 2×4 case for second input. Correlation is 0.9. Ricean factor is 8dB.
- Fig. 6. Comparison with exact MER receiver in 2×4 case for first input. Correlation is 0.9. Ricean factor is 8dB. Power ratio is 5dB.
- Fig. 7. Error rate in 3×6 case for first input. Correlation is 0.9. Ricean factor is 8dB.
- Fig. 8. Robustness to correlation in 2×4 case. Ricean factor is 0dB. SNR is 15db.
- Fig. 9. Robustness to Ricean channels in 2×4 case. Antenna correlation is 0. SNR is 15db.

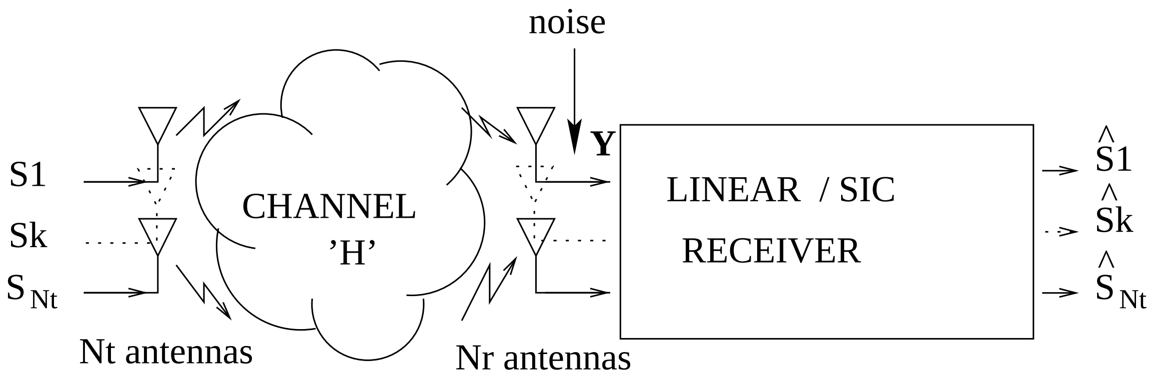

Fig. 1. Diagram of a MIMO spatial multiplexing system.

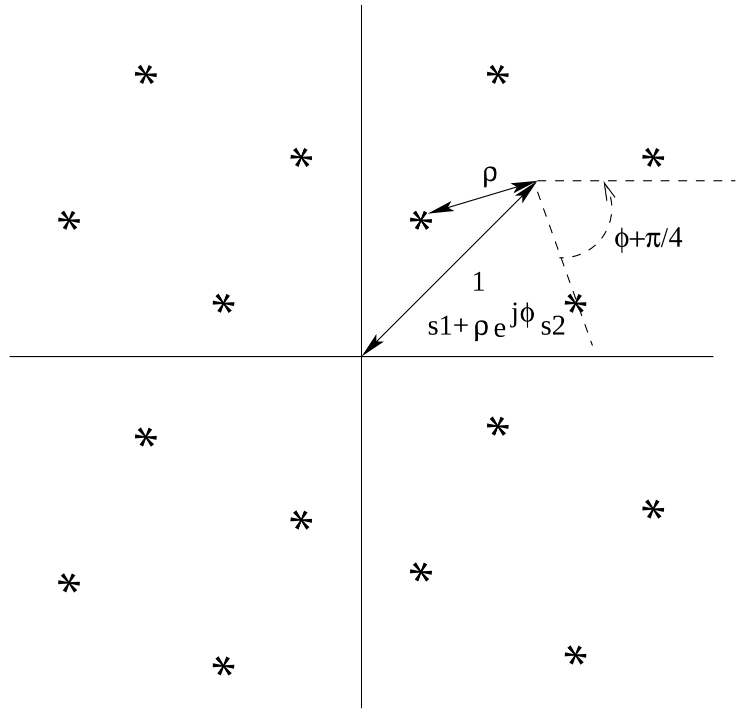

Fig. 2. Superposed QPSK constellations at the output of linear MIMO receiver with $N_t = 2$. The interfering constellation has amplitude ρ and phase ϕ.

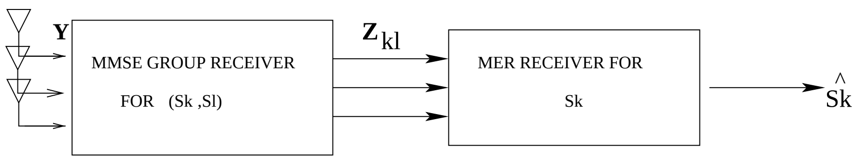

Fig. 3. Diagram of cascaded receiver for source s_k: a group MMSE receiver extracts the contribution of s_k, s_l. A MER receiver is then applied to estimate s_k.

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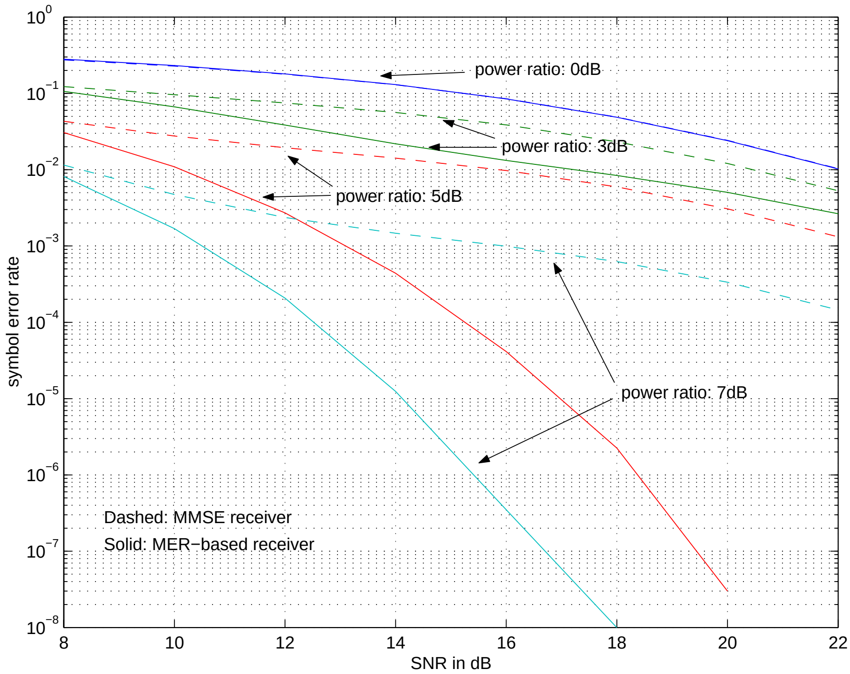

Fig. 4. Error rate in 2×4 case for first input. Correlation is 0.9. Ricean factor is 8dB.

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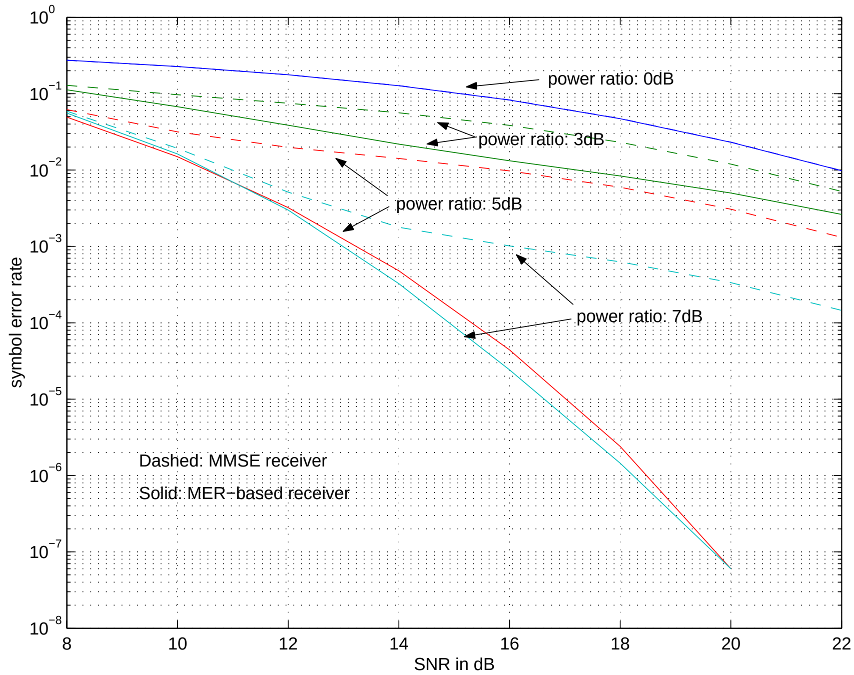

Fig. 5. Error rate in 2×4 case for second input. Correlation is 0.9. Ricean factor is 8dB.

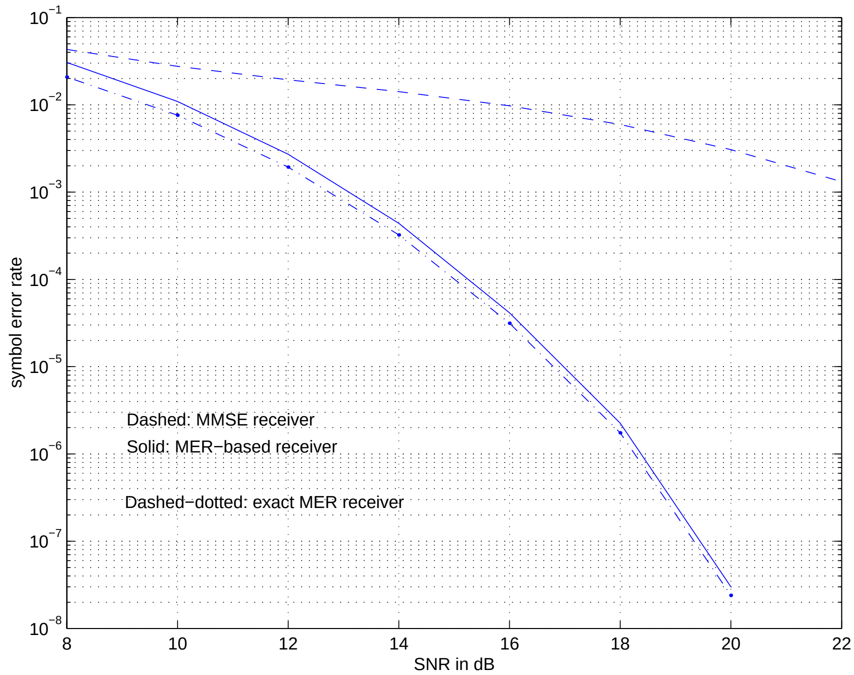

Fig. 6. Comparison with exact MER receiver in 2×4 case for first input. Correlation is 0.9. Ricean factor is 8dB. Power ratio is 5dB.

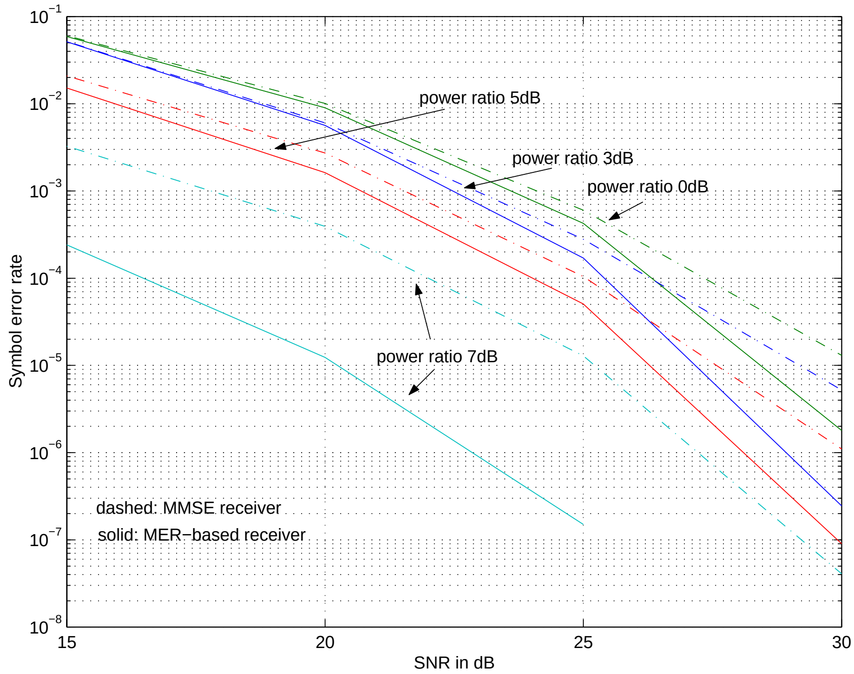

Fig. 7. Error rate in 3×6 case for first input. Correlation is 0.9. Ricean factor is 8dB.

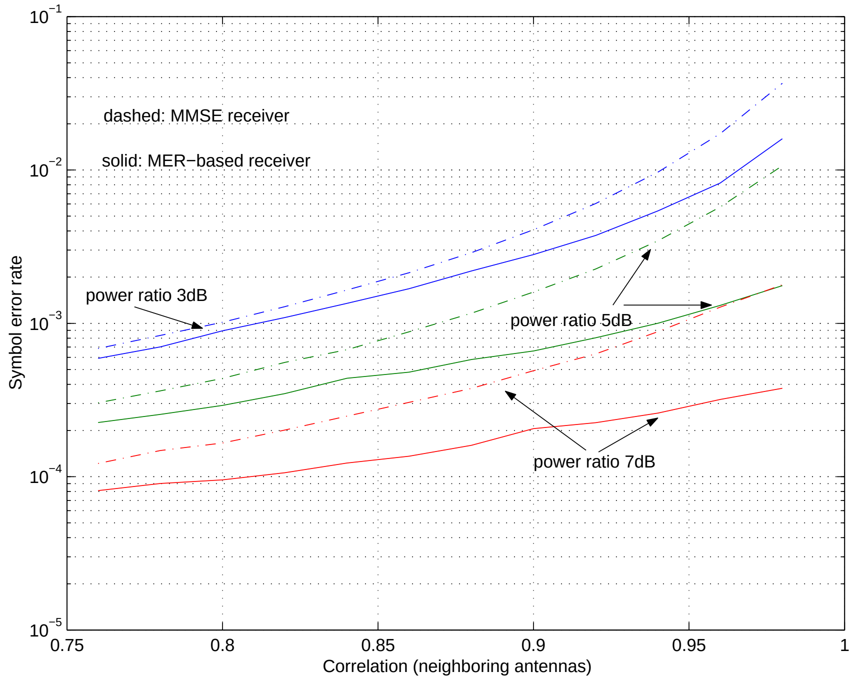

Fig. 8. Robustness to correlation in 2×4 case. Ricean factor is 0dB. SNR is 15db.

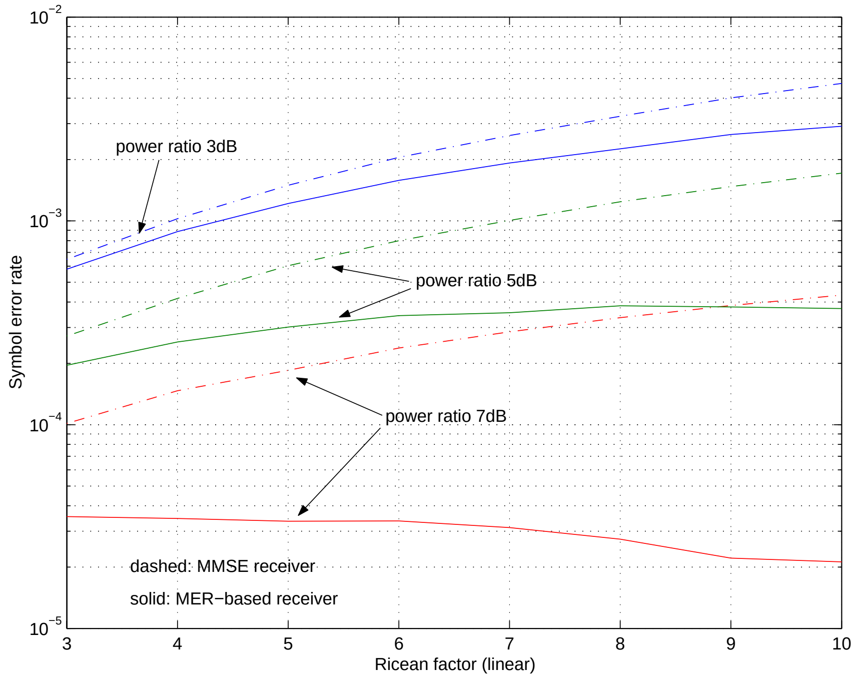

Fig. 9. Robustness to Ricean channels in 2×4 case. Antenna correlation is 0. SNR is 15db.