The International ITG Workshop on Smart Antennas WSA 2011 will be organized by the Institute of High Frequency Technology at the RWTH Aachen University in cooperation with VDE/ITG and possibly some other professional association (application for technical co-sponsorship pending) on February 24-25, 2011 in Aachen, Germany.

It provides a forum for presentation of the most recent research on smart antennas. The objective is to continue, accelerate, and broaden the momentum already gained with a series of ITG Workshops held in since 1996 in different places in Germany, Switzerland and Austria, especially in Ulm 2006, in Vienna 2007, in Darmstadt 2008, in Berlin 2009 and Bremen 2010.
IMPACT OF TRANSMIT IMPAIRMENTS ON MULTIUSER MIMO NON–LINEAR TRANSCIEVERS

José González-Coma, Paula M. Castro, and Luis Castedo

Universidad de A Coruña, Campus de Elviña s/n, 15071. A Coruña, Spain
Email: jgonzalezcoma@gmail.com, {pcastro,luis}@udc.es

ABSTRACT

This paper analyzes the impact of residual transmit impairments on the performance of multiuser Multiple–Input Multiple–Output (MIMO) systems. We focus on the uplink with Decision Feedback (DF) non–linear receivers and the downlink with Tomlinson-Harashima non–linear precoders. We show that transmit noise severely affects the performance of both non–linear transmission schemes, specially for the downlink with THP. Nevertheless, this degradation can be significantly alleviated when including the transmit noise into the transceiver design.

1. INTRODUCTION

Practical transmitters in wireless communications generate noise due to the presence of a large number of residual impairments such as quantization noise, sampling, and carrier frequency offset, phase noise, I/Q imbalance, ... [1]. Such transmitter noise is normally not taken into account for the modeling of Multiple–Input Multiple–Output (MIMO) systems although it can significantly degrade predicted performance of practical systems.

More specifically, [2] analyzes the impact of residual transmit impairments on the MIMO channel capacity and also on the performance of several MIMO detection algorithms. It is demonstrated that Maximum-Likelihood (ML) and max-log A Posteriori Probability (APP) MIMO detection suffer from a substantial performance loss in the presence of weak transmit noise, whereas the performance of linear ZF detection is much less affected.

In this work, we analyze the impact of the above–mentioned residual transmit impairments on the performance of the uplink and the downlink of a multiuser MIMO wireless communication system. We focus on multiuser MIMO non–linear transmit/receive schemes not considered in [2], namely, Decision Feedback (DF) MIMO reception for the uplink [3, 4] and Tomlinson-Harashima Precoding (THP) for the downlink [5]. Both schemes are widely used because of its good trade–off between performance and complexity and because they offer significant advantages over linear reception and precoding. However, the influence of transmit impairments on their respective performances has not been analyzed so far in the literature.

We show that noise generated at the transmitter significantly affects the performance of these non–linear MIMO transceivers, specially that of the downlink with THP. The performance of the uplink with MIMO reception, on the contrary, is more robust against the presence of transmit noise. At first glance, this seems quite natural since for THP MIMO systems channel equalization is carried out by processing the signals at the transmitter. However, we will also show that this performance degradation can be greatly mitigated when the transmit noise is taken into account into the transceiver design.

The remaining of this paper is organized as follows. Section 2 describes the signal model of a MIMO system which takes into account the noise generated by practical transmitters. This model is useful for both the uplink and the downlink of a multiuser MIMO system. Section 3 focuses on the Minimum Mean Square Error (MMSE) design of a DF receiver for a multiuser MIMO uplink considering the transmit noise, while Section 4 does the same for the downlink with THP. Section 5 presents the results of computer simulations carried out to evaluate the impact of transmit noise on system performance and, finally, Section 6 is devoted to the conclusions.

2. SYSTEM MODEL WITH TX–NOISE

Let us consider a narrowband MIMO communication system with \( N_t \) transmit and \( N_r \) receive antennas. By considering only the receiver thermal-noise, this system can be represented by the following discrete-time model

\[
y[n] = Hx[n] + \eta_r[n] \in \mathbb{C}^{N_r}, \tag{1}
\]

where \( x[n] \in \mathbb{C}^{N_t} \) represents the transmit signals, \( \eta_r[n] \in \mathbb{C}^{N_r} \) is the noise vector introduced by the receivers (which will be referred to as the Rx–noise), \( H \in \mathbb{C}^{N_r \times N_t} \) is the MIMO channel matrix and \( y[n] \in \mathbb{C}^{N_r} \) is the received signal vector. Note that we assume a block fading channel, where \( H \) remains constant during the transmission of a data frame. Noise is complex-valued Gaussian with zero mean and covariance matrix \( C_{\eta_r} \), i.e. \( \eta_r[n] \sim \mathcal{N}_C(0, C_{\eta_r}) \). Transmit energy is
into account, a more accurate model for the transmit signal is constrained to a value \( E_{tx} = \text{tr}(C_x) \), where \( \text{tr}(\cdot) \) denotes the trace operator and \( C_x \) is the covariance matrix of the input symbols \( x[n] \). Accordingly, we define the \textit{Signal to Receiver Noise Ratio} (SRxNR) as

\[
\text{SRxNR} = \frac{E_{tx}}{\text{tr}(C_{\eta_r})}.
\]

When the residual impairments at the transmitter are taken into account, a more accurate model for the transmit signal is

\[
x_t[n] = x[n] + \eta_p[n] \in \mathbb{C}^N,
\]

(3)

where \( \eta_p[n] \in \mathbb{C}^N \) will be referred to as the \textit{Tx–noise}. This noise encompasses different practical effects such as phase-noise and nonlinearity of Radio Frequency (RF) amplifiers and/or quantization and clipping of digital to analog converters. As explained in [2], the Tx–noise is adequately modeled as an additive Gaussian noise since it results from the sum of a large number of residual transmit impairments. A zero mean Tx–noise is assumed with covariance matrix \( C_{\eta_p} \), so that the \textit{Signal to Transmitter Noise Ratio} (STxNR) is defined as

\[
\text{STxNR} = \frac{E_{tx}}{\text{tr}(C_{\eta_p})}.
\]

As an example, practical implementations of the IEEE 802.11 (WiFi) standard achieve STxNR values ranging from 22 dB to 32 dB (see [2] and references therein). The Tx–noise is also assumed to be statistically independent from the Rx–noise.

Eqs. (1) and (3) are also adequate to represent a multiantenna multiuser wireless system where a base station equipped with \( N \) antennas communicates with \( K \) single-antenna users. When considering the uplink \( N_t = K \) and \( N_r = N \), whereas \( N_t = N \) and \( N_r = K \) for the downlink. Although the channel model is the same, multiuser systems impose certain constraints on the signal processing that can be carried out to recover the transmitted information [6]. For the uplink, the base station collects the signals from all users, so that they can be separated using conventional MIMO reception methods. These methods, however, cannot be used in the downlink because normally users do not cooperate and each receiver does not know the signals from the other receivers. Therefore, for the downlink, signal separation can only be carried out by resorting to precoding or to any other form of transmit processing.

3. UPLINK WITH DECISION FEEDBACK MIMO RECEIVER

Fig. 1 depicts the block diagram of the uplink of a multiuser system with a Decision Feedback non–linear MIMO receiver. Information symbols to be sent from the users will be represented by \( u[n] \in \mathbb{A}^K \), where \( \mathbb{A} \) denotes the modulation alphabet. These symbols are directly transmitted from the users without any additional transmit processing, i.e. \( x[n] = u[n] \).

We assume \( u[n] \) is zero mean with covariance matrix denoted by \( C_u \).

It is apparent from Eqs. (1) and (3) that the input signal at the receiver can be written as

\[
y_i[n] = y[n] + H\eta_i[n] \in \mathbb{C}^N,
\]

(5)

where \( y[n] = Hu[n] + \eta_p[n] \) would be the received signal if there were no Tx–noise.

In DF reception, received signals are passed through the feedforward filter \( F \), which forces the \textit{Intersymbol Interference} (ISI) to be spatially causal and the error to be spatially white (i.e. minimum variance). By means of the feedback filter \( B \) and of the feedback loop depicted in Fig. 1, ISI can be recursively canceled without changing the statistical properties of the noise providing that the noise variance is sufficiently small, so that the symbol detector (represented by \( Q(\cdot) \) in Fig. 1) produces correctly detected symbols.

By elaborating the signal model according to Fig. 1, the estimated signal \( \hat{u}[n] \) can be written as

\[
\hat{u}_t[n] = F y_t[n] + (I - B) \hat{u}_p[n],
\]

(6)

where \( y_t[n] \) is defined as in Eq. (5) and \( \hat{u}_p[n] \in \mathbb{A}^N \) denotes the detected symbols after the threshold quantizer.

The ordering in which symbols are detected significantly has an influence on the performance of DF MIMO receivers. In the system model shown in Fig. 1, the ordering is obtained with the multiplication of the detected symbols, \( \hat{u}_p[n] \), by the permutation matrix \( P^T \). This multiplication produces \( \hat{u}_p[n] \), which constitutes the vector of detected symbols conveniently sorted. Having in mind that \( PP^T = I \), we have that \( \hat{u}_t[n] = Pu_t[n] \) and, hence, \( \hat{u}_t[n] \) can be rewritten as

\[
\hat{u}_t[n] = F y_t[n] + (I - B) P\hat{u}_p[n].
\]

The MMSE design of the DF MIMO receiver searches for the filtering and permutation matrices that minimize the variance of the error vector

\[
\epsilon_{t,p}[n] = Pu_t[n] - \hat{u}_t[n].
\]

Assuming correct decisions (i.e. \( \hat{u}_p[n] = u[n] \)) and using Eq. (5), this error vector can be rewritten as

\[
\epsilon_{t,p}[n] = BPu[n] - Fy_t[n] = \epsilon_p[n] - FH\eta_t[n],
\]

where \( \epsilon_p[n] = BPu[n] - Fy_t[n] \) is the error vector when there is no Tx–noise. Since Tx–noise is independent from Rx–noise and transmit signals, the \textit{Mean Square Error} (MSE) cost function to be minimized can be written as

\[
E[\Vert\epsilon_{t,p}[n]\Vert_2^2] = E[\Vert\epsilon_p[n]\Vert_2^2] + \text{tr}(FHC_{\eta_t}H^HI^H),
\]

(7)

where \( E[\Vert\epsilon_p[n]\Vert_2^2] \) is the MSE with no Tx–noise. Notice that \( E[\Vert\epsilon_p[n]\Vert_2^2] \) is the cost function that is minimized in
conventional MMSE design, whereas the additional term \( \text{tr} \left( F H C_{\eta} H^H F^H \right) \) is the MSE improvement caused by Tx–noise.

An MMSE design of the multiuser MIMO uplink that accounts for the Tx–noise should minimize the MSE given by Eq. (7). Similarly to the scenario without Tx–noise [7], minimization of Eq. (7) is readily accomplished from the Cholesky factorization with symmetric permutation of

\[
\Phi_t = (H^H (H C_{\eta} H^H + C_{\eta})^{-1} H + C_u^{-1})^{-1}.
\]

That factorization is given by \( P \Phi_t P^H = L D L^H \), where \( L \) is a unit lower triangular matrix and \( D \) is a diagonal matrix. After this decomposition, it can be demonstrated that the filters \( F \) and \( B \) for the MMSE DF non–linear MIMO receiver solution are

\[
F_{\text{MMSE}}^{\text{DF}} = D L^H P H (H C_{\eta} H^H + C_{\eta})^{-1} \\
B_{\text{MMSE}}^{\text{DF}} = L^{-1}.
\]

The minimum value of the MSE cost function is obtained plugging \( F_{\text{MMSE}}^{\text{DF}} \) and \( B_{\text{MMSE}}^{\text{DF}} \) into Eq. (7). Hence, the MMSE value is

\[
\text{MMSE}_{\text{DF}} = \text{tr} \left( D \right),
\]

where \( D \) is the diagonal matrix obtained from the Cholesky factorization with symmetric permutation of \( \Phi_t \).

Notice that the MMSE expression given by Eq. (8) depends on the permutation matrix \( P \). Brute force optimization of \( P \) can be carried out by computing the MMSE for all the \( K! \) possible permutation matrices and choosing that matrix that provides the minimum value of (8). Alternatively, more efficient ordering algorithms (such as the one described in [7]) can be used.

From the MMSE design of the DF receiver, it is straightforward to obtain the expressions for the Zero–Forcing (ZF) DF receiver: it is the limiting case when \( \text{tr}(H C_{\eta} H^H + C_{\eta})/E_{\text{tx}} \to 0 \). The final expressions for the ZF DF filters are exactly the same as before but \( L \) and \( D \) should be obtained from the Cholesky decomposition of

\[
\Phi_t = (H^H (H C_{\eta} H^H + C_{\eta})^{-1} H)^{-1}.
\]

3.1. Noise Whitening

Noise whitening is an alternative way to incorporate the Tx–noise into the MMSE design of DF MIMO receivers. Let us rewrite the received signal model given by Eq. (5) in the following way

\[
y_t[n] = H (u[n] + \eta_t[n]) + \eta_r[n] = H u[n] + \eta[n],
\]

where \( \eta[n] = H \eta_t[n] + \eta_r[n] \) is the overall additive Gaussian noise caused by either Tx– or Rx–noise. The covariance matrix of \( \eta[n] \) is

\[
C_{\eta} = H C_{\eta} H^H + C_{\eta}.
\]

Notice that the additive Gaussian noise \( \eta[n] \) is always correlated even for the case in which both Tx– and Rx–noise are white because the channel always correlates the Tx–noise.

The additive noise appeared in the received signal can be forced to be white by passing it through the whitening filter

\[
W = \sigma_{\eta} C_{\eta}^{-1/2},
\]

where \( C_{\eta}^{-1/2} \) results from the Cholesky factorization of \( C_{\eta}^{-1} \) and \( \sigma_{\eta}^2 \) is the variance of the whitened noise. The signal at the output of the whitening filter can be expressed as

\[
y_w[n] = W y_t[n] = H_w u[n] + \eta_w[n],
\]

where \( H_w = W H \) is the effective MIMO channel matrix and \( \eta_w[n] \) is an additive spatially white Gaussian noise with covariance matrix \( C_{\eta_w} = \sigma_{\eta}^2 I \). Hence, the conventional MMSE design of a DF MIMO receiver for spatially white Gaussian noise could now be applied assuming that the receiver input is the output of the whitening filter, \( y_w[n] \), that the MIMO channel matrix is \( H_w \), and that the channel noise variance is \( \sigma_{\eta}^2 \).

4. DOWNLINK WITH TOMLINSON–HARASHIMA MIMO PRECODING

Fig. 2 shows the block diagram of the downlink of a multiuser MIMO system that employs Tomlinson–Harashima precoding. THP is a non–linear precoding technique made up of a feedforward filter \( F \in \mathbb{C}^{N \times K} \), a feedbackward filter \( I - B \in \mathbb{C}^{K \times K} \), and a modulo operator represented in Fig. 2 by \( M(\cdot) \). The modulo operator is introduced to avoid the increase in transmit power due to the feedback loop [5]. Data symbols sent from the base station to the users will be represented
by $u[n] \in \mathbb{A}^K$, where $\mathbb{A}$ denotes the modulation alphabet. Similarly to DF MIMO reception, the ordering considerably affects the performance of THP and for this reason, transmit symbols are passed through a permutation filter $P$. The output of the modulo operator will be represented by $v[n]$, which are the inputs to the feedforward filter to produce the transmit signal

$$x[n] = Fv[n] \in \mathbb{C}^N.$$ (14)

A total transmit energy constraint is imposed on the transmit signal, so that $E[\|x[n]\|^2] = E_t$.

From Eqs. (1) and (3), the input signal at the receiver when there is Tx−noise is

$$y_t[n] = y[n] + H\eta_t[n] \in \mathbb{C}^K,$$ (15)

where $y[n] = Hx[n] + \eta_t[n]$ would be the received signal if there were no Tx−noise. At reception, users are equipped with a single antenna and do not cooperate, so the only signal processing they can be carried out is to multiply their respective incoming signal by a complex−valued scalar magnitude. We will assume that all users apply the same scalar value denoted by $g$. This assumption is necessary in order to arrive at closed-form solutions for the MMSE−THP design. That scaling is represented in Fig. 2 by the receive filter $gL$. The signal corrected by the factor $g$ will be denoted by $\tilde{d}_t[n] = gy_t[n]$. Using Eqs. (14) and (15), this signal can be rewritten as

$$\tilde{d}_t[n] = d[n] + gH\eta_t[n] \in \mathbb{C}^K,$$ (16)

where $d[n] = gHFv[n] + g\eta_t[n]$ is the received signal when there is no Tx−noise. At the receivers, the modulo operator is applied again to invert the effect of this operator at the transmitter and the resulting signal is passed through a symbol detector (represented by $Q(\bullet)$ in Fig. 2) to produce the detected symbols $\hat{u}[n] \in \mathbb{A}^K$.

As explained in [8], the MMSE design of the MIMO downlink with THP searches for the filtering and permutation matrices that minimize the variance of the error vector

$$e_t[n] = P^T Bv[n] - d_t[n] = e[n] - gH\eta_t[n],$$

where $e[n] = P^T Bv[n] - gy[n]$ is the error vector when there is no Tx−noise. Minimization is carried out under the restriction of $B$ being a spatially causal filter and the transmit energy constraint $E[\|x[n]\|^2] = E_t$.

Since the transmit noise is independent of the transmitted signal and the receiver noise, the MSE can be decomposed as

$$E[\|e_t[n]\|^2] = E[\|e[n]\|^2] + |g|^2 \text{tr}(HC\eta_t H^H),$$ (17)

where $E[\|e[n]\|^2]$ is the MSE when there is no Tx−noise, which constitutes the cost function that is minimized in the conventional MMSE design of THP.

Following similar derivations as in [9, 10], the minimization of the MSE cost function in Eq. (17), subject to the mentioned constraints, can be carried out from the factorization of

$$\Phi_t = (HH^T + \xi_t I)^{-1},$$

where

$$\xi_t = \xi + \text{tr}(HC\eta_t H^H)/E_{\text{tx}},$$ (18)

with $\xi$ being the inverse of the SRxNR. The symmetrically permuted Cholesky decomposition of this matrix is

$$P\Phi_t P^T = L^H D L,$$ (19)

where $L$ and $D$ are, respectively, unit lower triangular and diagonal matrices. Finally, the MMSE solution for the THP filters that account for the Tx−noise is given by

$$F_{\text{MMSE}}^{\text{THP}} = g_{\text{MMSE}}^{-1} H^H P^T L^H D$$

$$B_{\text{MMSE}}^{\text{THP}} = L^{-1}.$$ (20)

The receive scalar weight $g_{\text{MMSE}}^{\text{THP}}$ is directly obtained from the transmit energy constraint. Assuming that it is real and positive, it is obtained that

$$g_{\text{MMSE}}^{\text{THP}} = \sqrt{\text{tr}(H^H P^T L^H D^2 C_v LPH)/E_{\text{tx}}},$$

where $C_v$ is the covariance matrix of $v[n]$, which is diagonal with entries depending on the modulation alphabet [11].

The minimum value for the MSE cost function given by Eq. (17) can be obtained by substituting the expressions obtained for the optimum filters $F_{\text{MMSE}}^{\text{THP}}$ and $B_{\text{MMSE}}^{\text{THP}}$ and for the gain factor $g_{\text{MMSE}}^{\text{THP}}$. It is easy to show that the final MMSE under the presence of Tx−noise is given by

$$\text{MMSE}_{\text{THP}} = \xi_t \text{tr}(C_v D),$$ (20)

where $\xi_t$ is given by Eq. (18) and $D$ is the diagonal matrix that results from the permuted Cholesky factorization of Eq. (19).
As done in [12], instead of testing all possible permutation matrices to find the one that minimizes the cost function of Eq. (20), the ordering optimization is included into the computation of the Cholesky decomposition of Eq. (19). Contrary to the ordering algorithm in DF reception, the latter detection stages in THP are more constrained since after each ordering iteration, one additional transmit signal is subject to precoding. As a consequence, precoding filter optimization is carried out in the direction opposite to the precoding ordering (see [12]). Again, it is straightforward to obtain the expressions for the ZF–THP design as the limiting case when $\xi_t \to 0$. The expressions for the filters $F_{\text{THP}}$ and $B_{\text{THP}}$ are equal to that obtained for $F_{\text{MMSE}}$ and $B_{\text{MMSE}}$, respectively, although the matrices $P$, $L$, and $D$ should be obtained from the symmetrically permuted Cholesky factorization of $\Phi_t = (H H^H)^{-1}$. (21)

It is important to point out that the alternative way of incorporating the Tx–noise into the precoder design by means of a whitening filter is not feasible in the downlink of an uncoordinated multiuser system. Notice that practical implementation of a whitening filter requires full knowledge of the signals at all the receiving antennas. This assumption may be realistic in a single user point–to–point MIMO system with a centralized receiver but it is not realistic in the downlink of a multiuser system, where users do not cooperate and, hence, the received signals are not exchanged among them.

5. SIMULATION RESULTS

Computer simulations were carried out to evaluate the impact of Tx–noise on the performance of the uplink and the downlink of a MIMO multiuser communication system using Decision Feedback reception and Tomlinson-Harashima precoding, respectively. We considered a multiuser MIMO scenario formed by a base station with $N = 4$ antennas and $K = 4$ single–antenna users. Symbols are transmitted with QPSK modulation and are grouped into frames of 100 symbols. We assumed an spatially white MIMO Rayleigh block fading channel where the entries to the channel matrix $H$ are circularly symmetric complex–valued zero mean independent and identically distributed Gaussian random variables with unit variance. The channel remains constant during the transmission of each 100 symbol frame but is statistically independent from one frame to another.

As usual, the Rx–noise is assumed to be spatially white (i.e. $C_{\eta_r} = \sigma_r^2 I$) and the expression for the SRxNR given by Eq. (2) reduces to $\frac{E_{\text{tx}}}{N_r \sigma_r^2}$. We have also assumed that the Tx–noise is spatially white (i.e. $C_{\eta_t} = \sigma_t^2 I$) in which case the expression for the STxNR is $\frac{E_{\text{tx}}}{N_r \sigma_t^2}$ (cf. Eq. (4)). As explained in [2], when there is sufficient decoupling between the Tx–RF chains, impairments are statistically independent across transmit antennas. This hypothesis has been confirmed through experimental measurements. For the computer experiments we set the STxNR to 25 dB, a representative value since in practical implementations STxNR ranges between 22 and 32 dB [2].

Figs. 3 to 7 show the results obtained from these computer experiments. System performance is measured in terms of uncoded Bit Error Rate (BER) against SRxNR. Each figure plots the BER curves corresponding to the following three different situations: i) system performance for the ideal case in which there is no Tx–noise; ii) system performance when there is Tx–noise, which has been ignored for the transceiver design;
Fig. 5. BER vs SRxNR for the uplink with MMSE MIMO DF reception. STxNR is 25 dB.

Fig. 6. BER vs SRxNR for the uplink with MMSE MIMO DF reception and whitening filter. STxNR is 25 dB.
iii) system performance when there is Tx–noise, which has been taken into account for the transceiver design.

Fig. 3 plots the BER vs. SRxNR curves obtained for the uplink of our multiuser MIMO system when a ZF–DF MIMO non–linear receiver is used. It is apparent from this figure that TX–noise has a small impact on system performance: there is a small degradation at high SRxNR when the Tx–noise is not considered into the receiver design but if the Tx–noise is considered, the performance is the same as that achieved without Tx–noise.

Fig. 4 plots the BER vs. SRxNR curves obtained for the downlink of the multiuser MIMO system when a ZF–THP is used. Contrarily to the ZF–DF receiver, the performance of ZF–THP is severely affected by the presence of Tx–noise. It can be seen from Fig. 4 that Tx–noise causes an error floor of $10^{-3}$ for SRxNR values above 25 dB. Moreover, incorporating the Tx–noise into the precoder design does not produce any performance improvement since the matrix $\Phi_t$ given by Eq. (21), whose Cholesky decomposition yields the precoder filters, does not depend on the Tx–noise.

Fig. 5 shows the BER curves for the uplink when an MMSE DF receiver is used. As expected, much lower BER values are reached (cf. Fig. 3) when using MMSE receivers. On the other hand, MMSE receivers are more sensitive to Tx–noise than ZF receivers. Indeed, the Tx–noise causes an error floor of $6 \times 10^{-6}$ for SRxNR values above 30 dB. However, Fig. 5 also shows that this degradation can successfully avoided if the Tx–noise is incorporated into the receiver design.

Fig. 6 compares the uplink performance for the two possible ways of incorporating the Tx–noise into the DF receiver design described in Section 3: without and with whitening filter. For the MMSE design, both approaches exhibit a similar performance. However, this is not so for the ZF design, where the whitening approach suffers from a severe degradation at SRxNR values above 20 dB.

Finally, Fig. 7 shows the BER curves for the downlink when an MMSE–THP is applied. First of all, notice the superior performance of MMSE–THP with respect to ZF–THP. Second, observe the important degradation that conventional MMSE–THP suffers due to the presence of Tx–noise: BER increases when SRxNR exceeds 20 dB to converge to the performance achieved by the ZF–THP setup. Third, notice how this malfunction can be greatly alleviated by incorporating the Tx–noise into the THP design in which case an error floor arises although at the low value of about $10^{-7}$.

6. CONCLUSIONS

The impact of noise generated by practical transmitters on the performance of multiuser multiantenna systems has been investigated. Attention has been focused on the uplink with non–linear Decision Feedback (DF) reception and the downlink with Tomlinson–Harashima non–linear Precoding (THP). We have shown that the performance of the uplink with DF MIMO reception is relatively robust to the
presence of transmit noise. On the other hand, transmit noise severely degrades the performance of the downlink with MIMO THP. Nevertheless, we have also shown that this performance degradation can be significantly mitigated if Tx–noise is included into the transceiver design.

7. ACKNOWLEDGMENTS

The authors would like to thank A. Burg for fruitful discussions on residual transmit RF-impairments. This work was supported by Xunta de Galicia, Ministerio de Educación y Ciencia, Ministerio de Ciencia e Innovación of Spain and FEDER funds of the European Union under grants number 09TIC008105PR, TEC2010-19545-C04-01 and CSD2008-00010.

8. REFERENCES


