A PRACTICAL METHOD FOR WIRELESS CHANNEL RECIPROCITY EXPLOITATION THROUGH RELATIVE CALIBRATION

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ABSTRACT

We present a relative calibration method for a wireless TDD link, which, after a calibration phase involving feedback, lets the transmitter acquire knowledge of the downlink channel state from the uplink channel estimates, through proper modeling and estimation of the RF circuitry impulse responses. Contrarily to previous methods, relative calibration does not require specific calibration hardware. Experimental results confirm the validity of the proposed linear reciprocity model, and of the calibration approach.

1. INTRODUCTION

Knowledge of the channel state by the transmitter of a communications system has been demonstrated to improve the efficiency of wireless communications. Various methods have been proposed to shape the transmitted signal according to the channel state information (CSI) available at the transmitter, for both single-antenna and multiple-antenna settings. Besides the classical method using continuous estimation and feedback to bring CSI to the transmitter, the use of channel reciprocity is usually suggested for Time-Division Duplex (TDD) systems. The reciprocity principle is based on the property that electromagnetic waves traveling in both directions will undergo the same physical perturbations (*i.e.* reflection, refraction, diffraction, etc…). Therefore, if the link operates on the same frequency band in both directions, the impulse response of the channel observed between any two antennas should be the same regardless of the direction. Application of the reciprocity principle lifts the requirement for a continuous feedback of the channel estimates while still allowing to make use of CSIT in order to optimize the transmission.

Despite the fact that the electromagnetic foundations of the reciprocity principle, due to H. A. Lorentz, have been known since 1896 and extensively explored (see for instance [1] and references therein), applications in the field of wireless communications have been scarce. This is due to the general understanding that the non-symmetric characteristics of the radio-frequency (RF) electronic circuitry would break the reciprocity property. Various solutions to this issue have been recently proposed. One is to calibrate each transmitter and receiver, *i.e.* to let them learn and compensate for the characteristics of their own circuitry [2]. We refer to this method as absolute calibration. This method has been is use in the radar community for a long time, since absolute calibration is necessary to determine the direction of arrival of an electromagnetic wave. It requires an external reference source with tight requirements, and is therefore expensive to implement in the context of a communications system. Another method [3] aims at ensuring the reciprocity of the electronic circuitry through a specially crafted transceiver where the same op-amp is used for both transmitting and receiving, thus lifting the requirement of calibration at the expense of design complexity.

Contrarily to these methods relying on hardware solutions, we propose a signal-space calibration (or *relative* calibration) method. It relies on a calibration phase to establish the relationship between the channel as measured in both directions. We deem this a relative calibration since it takes place entirely in signal space, and no external reference source, nor any other hardware, are necessary.

In the sequel, we introduce a linear reciprocity model, and the concept of relative calibration. We propose a class of algorithm to estimate the reciprocity parameters, and assess the validity of the proposed model through experimental results obtained for the case of a Single-Input Single-Output (SISO) system.

2. SYSTEM MODEL

Let us consider a bidirectional TDD MIMO link, between station A and station B, using respectively M and N antennas. We model the channel as seen during the baseband processing as the cascade of three linear filters, and some additive white Gaussian noise (AWGN), as represented in Fig. 1. The upper part of the diagram represents a transmission from A to B, whereas the lower part represents a transmission from B to A. T_A denotes the M-input M-output equivalent filtering operation of the transmit circuitry of A, $C(t)$ is a $N \times M$ matrix containing the impulse responses (one per Tx-Rx antenna pair) of the

*Institut Eurécom's research is partially supported by its industrial members: Swisscom, Thales, SFR, France Telecom, Hitachi Europe, Texas Instruments, STMicroelectronics, Bouygues Telecom, and Sharp.

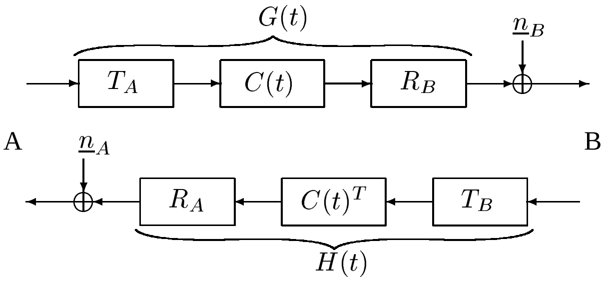

Fig. 1. Reciprocity model for a MIMO FDD frequency-selective channel

electromagnetic channel at time t, and R_B is the N-input N-output equivalent filter modeling the receive circuitry of B. Symmetrically, T_B and R_A denote the equivalent filters corresponding respectively to the transmit circuitry of B and the receive circuitry of A. Note that the characteristics of the circuitry do not depend on t, since their variation is usually much slower than the channel variation.

In a TDD setting, assuming that the transmissions in both directions take place within a time frame shorter than the channel coherence time, the reciprocity of the electromagnetic channel guarantees that the impulse responses between each antenna pair is the same in both directions, therefore the filter $C(t,\tau)$ is common to both directions (it needs to be transposed to respect the order of the antennas). Let us denote by

$$G(t,\tau) = R_B(\tau) * C(t,\tau) * T_A(\tau) \tag{1}$$

the compound impulse response (τ is the index in the lag domain, and $*$ denotes the convolution) of the (noiseless) channel from A to B, and by

$$H(t,\tau) = R_A(\tau) * C(t,\tau)^T * T_B(\tau) \tag{2}$$

its counterpart when the transmission takes place from B to A. For the sake of simplicity, the noise $\underline{n}_A$, $\underline{n}_B$ is supposed to be injected after the cascade of the three filters, although it really appears between the electromagnetic channel and the receive circuitry. Knowledge of $G(t,\tau)$ is easily available to station B, using any classical channel estimation method, and similarly station A can estimate $H(t,\tau)$.

Note that $T_A(f)$, $T_B(f)$, $R_A(f)$ and $R_B(f)$ are all square matrices, and we will work under the assumption that they have no singularities for any f in the considered frequency band. This should be a reasonable requirement, since the design target for the circuitry is usually to have unit diagonal gains over the desired band, and as little crosstalk as possible. A strictly diagonal structure can be assumed if little or no crosstalk is present between antenna channels in the circuitry, thus further simplifying the model.

3. RELATIVE CALIBRATION

As stated before, although it is common practice for station B to estimate G in order to perform coherent detection of the received signal, the knowledge of G is desirable to A, since it would enable the use of CSIT-exploiting methods. Instead of relying on continuous feedback of CSI from B to A, we link the channel estimates in both directions by eliminating C in eqs. (1) and (2). This yields the frequency-domain expression $G(t,f) = R_B(f)T_B(f)^{-T}H(t,f)^T R_A(f)^{-T}T_A(f)$. Note that T_A, T_B, R_A and R_B are generally not known individually, since this would constitute absolute calibration on both sides of the link. Nevertheless, it can be seen that only $P_A(f) \triangleq R_A(f)^{-T}T_A(f)$ and $P_B(f) \triangleq R_B(f)T_B(f)^{-T}$ are necessary in order to infer $G(t)$ from $H(t)$ through

$$G(t,\tau) = P_B(\tau) * H(t,\tau)^T * P_A(\tau). \tag{3}$$

Estimating the (matrix) filters $P_A(\tau)$ and $P_B(\tau)$ constitutes a relative calibration between A and B, and can be realized entirely through the use of classical channel estimation and feedback techniques, as will be shown in the sequel.

4. RECIPROCITY PARAMETERS ESTIMATION

In this section, we present a method to estimate the reciprocity parameters $P_A(\tau)$ and $P_B(\tau)$ from one or several pairs of (simultaneous, uplink and downlink) channel measurements.

SISO case: In this particular case, the product in eq. (3) commutes because all factors are 1×1 matrices. Therefore, letting $P(\tau) = P_B(\tau) * P_A(\tau)$, we rewrite (3) as

$$G(t,\tau) = H(t,\tau) * P(\tau). \tag{4}$$

Let us consider K pairs of measurements of the discretized complex channel impulse responses in both directions, $\underline{\mathbf{g}}_k \triangleq \left(g_1^{(k)}, \ldots, g_L^{(k)}\right)^T \in \mathbb{C}^L$, $\underline{\mathbf{h}}_k \triangleq \left(h_1^{(k)}, \ldots, h_{L'}^{(k)}\right)^T \in \mathbb{C}^{L'}$. Under the finite length assumption, the convolution in the reciprocity condition (4) can be written as

$$\mathbf{H}_k \underline{\mathbf{p}} = \underline{\mathbf{g}}_k, \tag{5}$$

with $\underline{\mathbf{p}} \triangleq (p_1, \ldots, p_{L'-L+1})^T$ and

$$\mathbf{H}_k \triangleq \begin{bmatrix} h_{L'-L+1}^{(k)} & h_{L'-L}^{(k)} & \cdots & h_1^{(k)} \\ h_{L'-L+2}^{(k)} & h_{L'-L+1}^{(k)} & \cdots & h_2^{(k)} \\ \vdots & & \ddots & \vdots \\ h_{L'}^{(k)} & \cdots & & h_{L'-L+1}^{(k)} \end{bmatrix}.$$

(note that the length of the impulse response of the filter, $L' - L + 1$, must be a sensible value.) It is possible to solve for $\underline{\mathbf{p}}$ in eq. (5), through *e.g.* least-squares if the system is overdetermined, however this method assumes that only $\underline{\mathbf{g}}_k$ is noisy, and that $\underline{\mathbf{h}}_k$ is known perfectly. Since in practice, only the noisy versions $\widehat{\underline{\mathbf{g}}}_k$ and $\widehat{\underline{\mathbf{h}}}_k$ are known, we look for $\underline{\mathbf{p}}$ as the solution of the optimization problem

$$\min_{\mathbf{E}_k, \underline{\mathbf{x}}} \underline{\alpha}_k^H \underline{\alpha}_k + \underline{\beta}_k^H \underline{\beta}_k, \quad \text{s.t.} \left(\widehat{\mathbf{H}}_k + \mathbf{E}_k\right) \underline{\mathbf{x}} = \widehat{\underline{\mathbf{g}}}_k + \underline{\beta}_k \tag{6}$$

where $\mathbf{E}_k$ is Toeplitz and contains the coefficients of $\underline{\alpha}_k$:

$$\mathbf{E}_k \triangleq \begin{bmatrix} \alpha^{(k)}_{L'-L+1} & \alpha^{(k)}_{L'-L} & \cdots & \alpha^{(k)}_1 \\ \alpha^{(k)}_{L'-L+2} & \alpha^{(k)}_{L'-L+1} & \cdots & \alpha^{(k)}_2 \\ \vdots & & \ddots & \vdots \\ \alpha^{(k)}_{L'} & \cdots & & \alpha^{(k)}_{L'-L+1} \end{bmatrix}.$$

The vectors $\underline{\alpha}_k \in \mathbb{C}^{L'}$ and $\underline{\beta}_k \in \mathbb{C}^{L}$ represent the corrections of the noise present on $\underline{\mathbf{h}}_k$ and $\underline{\mathbf{g}}_k$ respectively. Under the assumption that the noise is Gaussian i.i.d., the ML (Maximum-Likelihood) solution of (6) can be obtained numerically, since this formulation defines a Structured Total Least-Squares (STLS) problem, as recognized by Mastronardi [4].

In order to guarantee the identifiability of $\underline{\mathbf{p}}$, and since the measurements are noisy, it is preferable to over-determine the problem. To this aim, the linear system (5) can be extended by concatenating the successive channel measurements, since $\underline{\mathbf{p}}$ is assumed to remain constant over all measurements: $\begin{bmatrix} \mathbf{H}_1 \\ \vdots \\ \mathbf{H}_K \end{bmatrix} \underline{\mathbf{p}} = \begin{bmatrix} \underline{\mathbf{g}}_1 \\ \vdots \\ \underline{\mathbf{g}}_K \end{bmatrix}$. The STLS algorithm can be straightforwardly extended to estimate $\underline{\mathbf{p}}$ using all the measurements, by solving the optimization problem (with $\underline{\alpha} \triangleq [\underline{\alpha}_1^T \dots \underline{\alpha}_K^T]^T$)

$$\min_{\mathbf{E}_1,\dots,\mathbf{E}_K,\underline{\mathbf{x}}} \underline{\alpha}^H \underline{\alpha} + \underline{\beta}^H \underline{\beta}, \qquad (7)$$

$$\text{s.t.} \quad \begin{bmatrix} \widehat{\mathbf{H}}_1 + \mathbf{E}_1 \\ \vdots \\ \widehat{\mathbf{H}}_K + \mathbf{E}_K \end{bmatrix} \underline{\mathbf{x}} = \begin{bmatrix} \widehat{\underline{\mathbf{g}}}_1 \\ \vdots \\ \widehat{\underline{\mathbf{g}}}_K \end{bmatrix} + \underline{\beta}. \qquad (8)$$

SIMO and MISO cases: Consider now the situation where only one side of the link is equipped with multiple antennas, namely SIMO (Single-Input Multiple-Output, *i.e* $M = 1, N > 1$) and MISO (Multiple-Input Single-Output, *i.e.* $M > 1, N = 1$) channels. In both cases, either $P_A(\tau)$ or $P_B(\tau)$ is a 1×1 filter, and therefore the commutation property can still be used to transform eq. (3) into $G(t,\tau) = P_{SIMO} * H(t,\tau)^T$ (SIMO case) or $G(t,\tau) = H(t,\tau)^T * P_{MISO}$ (MISO case), where $P_{SIMO} \triangleq P_B(\tau) * (\mathbf{I}_N \otimes P_A(\tau))$ and $P_{MISO} \triangleq (\mathbf{I}_M \otimes P_B(\tau)) * P_A(\tau)$. In both cases, the reciprocity parameters are grouped into one single linear filter, which again can be efficiently (ML-)estimated using the STLS method. Furthermore, if no crosstalk is present on the side equipped with the multiple antennas, the problem merely degenerates into several parallel SISO channels whose reciprocity parameters can be estimated independently. This is evidenced by the fact that, for the respective SIMO and MISO cases, if P_B (resp. P_A) is diagonal, $P_B(\tau) * (\mathbf{I}_N \otimes P_A(\tau))$ (resp. $(\mathbf{I}_M \otimes P_B(\tau)) * P_A(\tau)$) are also diagonal filters.

MIMO case: Estimation of the reciprocity parameters in the MIMO case ($M > 1$ and $N > 1$) is less straightforward, since in this case eq. (3) is not jointly linear in the unknowns (P_A,P_B), and can not be made linear by commutation of the filters.

A first approach, applicable only in the case where P_A and P_B are both diagonal, is to over-parameterize the model. In order to do this, first note that eq. (3) is equivalent to the MN equations obtained from $[G(t,\tau)]_{i,j} = [P_B(\tau)]_{i,i} * [H(t,\tau)]_{j,i} * [P_A(\tau)]_{j,j}$ for $i = 1 \dots N, \quad j = 1 \dots M$ (where $[\cdot]_{i,j}$ denotes the $(i,j)^{th}$ element of a matrix). Since $[P_A(\tau)]_{j,j}$ and $[P_B(\tau)]_{i,i}$ represent SISO filtering operations, each $P_{(i,j)} \triangleq [P_B(\tau)]_{i,i} * [P_A(\tau)]_{j,j}$ can be estimated as in the SISO case since $[G(t,\tau)]_{i,j} = P_{(i,j)} * [H(t,\tau)]_{j,i}$. However, this method fails to take into account the fact that the MN linear filters $P_{(i,j)}$ are generated from only $M+N$ impulse responses.

The second proposed approach relies on alternating estimation: it is an iterative method whereby $P_A(\tau)$ and $P_B(\tau)$ are alternatively assumed to be known perfectly, while the other is estimated under this assumption (the STLS algorithm can be used again, since assuming that $P_A(\tau)$ is perfectly known makes the problem linear in $P_B(\tau)$, and vice-versa). The algorithm is initialized by assuming *e.g.* that $P_A(\tau) = \mathbf{I}_M$ at the first iteration. The proposed algorithm is therefore the following:

1. initialization: $\widehat{P}_A(\tau) = \mathbf{I}_M$
2. assuming $\widehat{P}_A(\tau)$ perfectly known, compute $\widehat{P}_B(\tau)$ as the TLS solution of a linear system identification problem expressed by equation (3).
3. assuming $\widehat{P}_B(\tau)$ perfectly known, compute $\widehat{P}_A(\tau)$, by applying TLS to eq. (3) again (note that the same equation describes different problems since the role of the unknowns and the constants have changed).
4. iterate (go to step 2) until convergence

This approach is applicable even for non-diagonal reciprocity parameters, and does not over-parameterize the system. Unfortunately, its convergence has not been proved, and no optimality claim can be made about the results. However, initial simulation results suggest that convergence is not a problem, and that the results constitute a valid estimate of $P_A(\tau)$ and $P_B(\tau)$.

5. EXPERIMENTAL RESULTS

In order to validate the linear system model and to assess the validity of the calibration process, channel measurements were performed in the case of a SISO system, and processed using the algorithm of section 4. The experimental setting is based on a prototype UMTS TDD indoor-to-outdoor link [5] operating on a 3.84MHz wide channel in the 1900-1920MHz IMT-2000 TDD band. Channel estimates were obtained in the framework of an actual UMTS connection, using conventional channel estimation techniques based on training sequences embedded in the UMTS traffic. The feedback link required by the calibration phase was assumed to be of infinite bandwidth (*i.e.* the estimates performed on both sides were made available to one single place for computation with no further degradation). One channel estimate for each direction was obtained every 10ms. The oscillators on both sides of the wireless link were synchronized through a

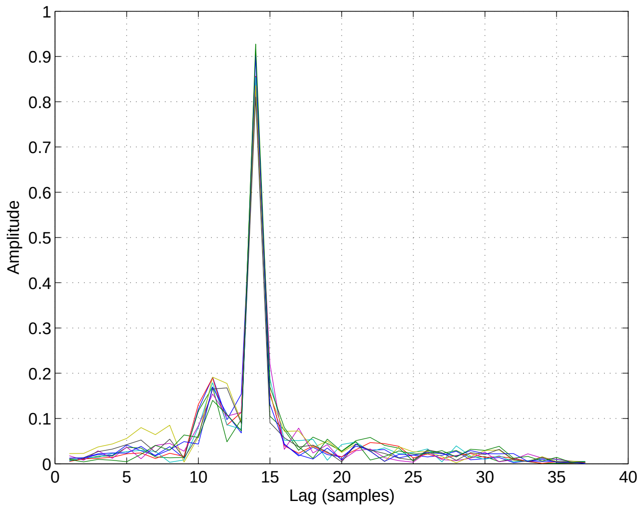

Fig. 2. $\widehat{P}(\tau)$ impulse response, fixed setting

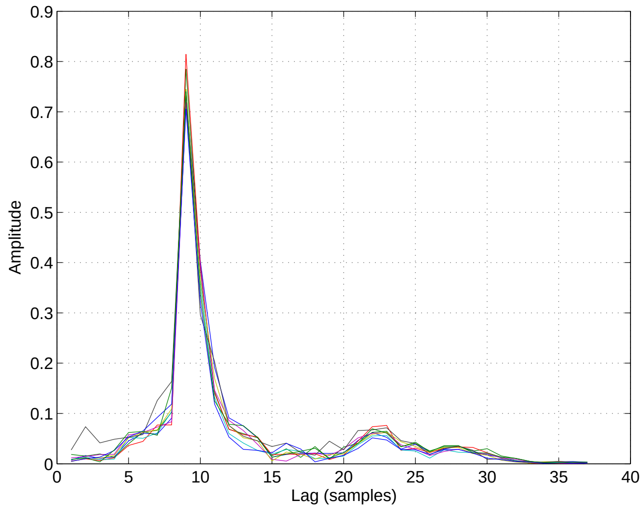

Fig. 3. $\widehat{P}(\tau)$ impulse response, moving setting

wired link. Fig. 2 and 3 show the time-domain representation of the estimated reciprocity function $\widehat{P}(\tau)$. Fig. 2 corresponds to a fixed setting, with the antenna corresponding to the mobile terminal (MT) lying on a table and no movement in the environment, whereas in the moving setting presented in Fig. 3, the MT antenna was hand-held and moved rapidly by a human operator. Every figure shows 9 curves, corresponding to successive estimates of $\widehat{P}(\tau)$. Each estimate is done over $K = 50$ successive channel estimates, *i.e.* over a 500ms time span.

The first noticeable characteristic of these figures is their deviation from the identity filter, which demonstrates the absence of overall reciprocity if no calibration is performed. The relative stability of the estimates of $P(\tau)$ in the series of 9 consecutive measurements indicates that the assumption that reciprocity parameters vary slowly is valid, and is therefore an encouraging sign that the proposed relative calibration process is possible. Comparing Figs. (2) and (3), and bearing in mind that both experiments used the same hardware and were separated by only a few minutes, the discrepancy in the reciprocity function can presumably be attributed to changes in the MT antenna coupling, due to the presence of the operator's hand near the antenna in the second case (moving setting). This points towards the necessity of frequent calibration cycles when the environment in close proximity of the antennas is potentially changing.

The same series of measurements was reprocessed in order to evaluate the accuracy of channel estimates obtained through reciprocity after relative calibration. After estimating $\widehat{P}$ over 50 2-tuples $(\widehat{G}(t), \widehat{H}(t))$, $\widehat{G}_{est}(t)$ was computed as $G_{est}(t,\tau) = \widehat{H}(t,\tau) * \widehat{P}(\tau)$ for the subsequent 422 measurements of $\widehat{H}(t,\tau)$ (*i.e.* over a 4.22s time span), and compared to $\widehat{G}(t)$. We use the noise amplification metric $\alpha = \frac{||G_{est} - \widehat{G}||^2}{\sigma^2_{\widehat{G}}}$, for which lower values are better, and $\alpha \geq \alpha' \triangleq 1 + \frac{SNR_{\widehat{G}}}{SNR_{\widehat{H}}}$. $\alpha = \alpha'$ if the reciprocity is perfect. The measured channels yield $\alpha_{fix} = 2.02$, very close to $\alpha'_{fix} = 1.96$, for the fixed measurements, which confirms the validity of the convolutive model. In the moving setting, $\alpha_{mob} = 3.72$ whereas $\alpha'_{mob} = 1.73$. The larger discrepancy in this case could be explained by the fact that the uplink and downlink channel measurements are only approximately simultaneous, and therefore the underlying channel $C(t,\tau)$ itself might have changed.

6. CONCLUSION

We introduced the concept of relative calibration in a wireless TDD communications system, whereby channel reciprocity can be exploited without specific calibration hardware, since the calibration takes place entirely in signal-space. We proposed algorithms to estimate the reciprocity functions in various cases. An experimental assessment of the proposed linear reciprocity model showed the validity of this model, and of the relative calibration concept.

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