



**HAL**  
open science

## Transmission Techniques and Channel Calibration for Spatial Interweave TDD Cognitive Radio Systems

Francesco Negro, Boris Kouassi, Irfan Ghauri, Luc Deneire, Dirk Slock

► **To cite this version:**

Francesco Negro, Boris Kouassi, Irfan Ghauri, Luc Deneire, Dirk Slock. Transmission Techniques and Channel Calibration for Spatial Interweave TDD Cognitive Radio Systems. 2012. hal-00845671

**HAL Id: hal-00845671**

**<https://hal.science/hal-00845671>**

Preprint submitted on 17 Jul 2013

**HAL** is a multi-disciplinary open access archive for the deposit and dissemination of scientific research documents, whether they are published or not. The documents may come from teaching and research institutions in France or abroad, or from public or private research centers.

L'archive ouverte pluridisciplinaire **HAL**, est destinée au dépôt et à la diffusion de documents scientifiques de niveau recherche, publiés ou non, émanant des établissements d'enseignement et de recherche français ou étrangers, des laboratoires publics ou privés.

# Transmission Techniques and Channel Calibration for Spatial Interweave TDD Cognitive Radio Systems

Francesco Negro, Boris Kouassi, Irfan Ghauri, Luc Deneire, and Dirk T.M. Slock

**Abstract**—We study the problem of beamforming design for a Cognitive Radio (CR) system in which a multiple-input multiple-output (MIMO) link, identified as secondary link, wants to opportunistically communicate without harming a licensed MIMO system, called primary.

In the proposed solution the opportunistic user designs its beamformer in order to span the noise subspace at the primary receiver, thus intertwining its signal with the primary's so that its signal lies within the spatial whitespace of the primary system, causing no interference to the latter. This is spatial interweave. To solve this beamforming design problem the knowledge of channel state information at the transmitter (CSIT) is crucial. In our model we do not require any *a priori* knowledge of the channel information but we rely on channel reciprocity in Time-Division Duplex (TDD) transmission. We provide beamformer design algorithms and channel estimation procedures which allow the secondary communication, without need of cooperation with the primary system. However, in practice uplink and downlink channels are not reciprocal due to non reciprocal Radio Frequency front-ends. To compensate the channel mismatch we introduce a new calibration algorithm that outperforms the performance of previously proposed solutions for MIMO channels. We prove that channel calibration is a crucial operation and, more importantly, that it can be done without cooperation between primary and secondary systems. Finally, after a model extension to multiple primary pairs, we show how it is possible to implement in practice the proposed CR settings using the frame structure of the recent cellular communication standard Long Term Evolution (LTE).

## I. INTRODUCTION AND STATE OF THE ART

Cognitive Radio [1] (CR) has been recently introduced to enhance the spectral efficiency of modern wireless communication systems. The basic idea behind CR is that an opportunistic system, usually called secondary system, can transmit using the same communication resources as a licensed system, called primary system, while keeping the interference caused to the primary system under control. The remarkable work in [2] has suggested the classification of Cognitive Radio in three categories namely, *Overlay*, *Underlay* and *Interweave* depending upon the extent of inter-system cooperation and interference tolerance of the legacy party.

F. Negro and D.T.M. Slock are with Mobile Communications Department of EURECOM, Sophia Antipolis, France. I. Ghauri is with Intel Mobile Communications, Sophia Antipolis, France. B. Kouassi and L. Deneire are with the I3S laboratory at the University of Nice Sophia Antipolis, France. EURECOM's research is partially supported by its industrial members: BMW Group Research & Technology, Swisscom, ORANGE, SFR, ST Microelectronics, Symantec, Monaco Telecom, SAP, iABG and by the EU FP7 projects CROWN, SACRA, WHERE2 and NEWCOM#. The research of Intel Mobile Communications is also supported in part by the EU FP7 FET projects CROWN and SACRA.

Overlay CR is a cooperative technique in which the secondary signals are designed to offset any degradation they may cause to primary communications, requiring a shared knowledge of the codebooks and modulation schemes.

Underlay CR allows coexistence of a primary and a secondary network, constraining interference caused by secondary transmitters on primary receivers to be under a certain threshold, usually called *Interference temperature constraint* [1].

Finally, Interweave (IW) CR exploits the unused communication resources, called *white spaces*, of the primary system in an opportunistic fashion. In this communication paradigm, secondary transmission can take place only if it does not cause any interference to the primary user. The unused primary resources can be time, frequency or, as recently introduced, space.

One of the first attempts to study how it is possible to exploit the spatial dimension at the secondary users in a cognitive radio setting is reported in [3]. In [3] the trade off between maximizing the secondary user's rate and controlling the interference caused at the primary receivers is studied for different CR settings. The authors exploit multiple antennas at the secondary transmitter to design the optimal transmit filter that effectively balances between spatial multiplexing for the secondary transmission and interference avoidance at the primary receivers. In this paper, the abbreviations Tx, Rx may stand for transmit, transmission, transmitter resp. receive, reception, receiver.

In [4] the authors considered a heterogeneous system in which the primary and secondary communications can coexist only if the latter keeps under control the interference generated at the primary Rx. They propose an iterative algorithm to design the secondary system transmission parameters to maximize the secondary user's rate while imposing a maximum interference constraint to the primary Rx or the more stringent constraint of zero interference to the primary system. The proposed algorithm is completely decentralized and is based on iterative water filling (IWF). The authors also provide an analytical description of the performance using a game theory framework. In [4] the authors do not investigate how the cognitive users can acquire all the necessary pieces of information on channels and primary communication parameters. In [5] a more practical spatial interweave CR setting is studied. Interestingly in this work no *a priori* knowledge is assumed at the secondary network but the necessary information is acquired during a learning phase that exploits reciprocity of the primary TDD communication strategy. During this phase also

partial knowledge of the primary signal subspace is acquired. They underline that the proposed scheme is better, in term of degrees of freedom (DoF), than the previously proposed solution in [3], because partial knowledge of the Rx subspace at the primary Rxs increases the number of streams that can be transmitted from the cognitive Tx. The authors call this opportunistic transmission, at the secondary system, as *opportunistic spatial sharing*.

The authors of [6] studied the same setting of [5] but with the objective of making their work more practical. A transmission scheme of three phases is introduced in which the primary-to-secondary channel is acquired, then the channel between secondary users is estimated and finally the transmission takes place. In the proposed analysis the secondary channel estimation errors are taken into account in the secondary beamformer (BF) design and the interference caused at the secondary Rx, due to primary communication, is reduced introducing a Rx filter at the secondary receiver.

In the spatial interweave scenario the secondary Tx can design its transmitted signal according to a form of *Interference Alignment* (IA) [7]. Using this transmission method the secondary signal is designed in such a way that it is constrained to lie in the received signal dimensions that are not occupied by the primary communications. As a result there is no degradation of the performance of the primary system. This beamforming technique has been proposed in [8] where it is called *opportunistic interference alignment*. The authors assume perfect knowledge of all channels without investigating how to obtain the necessary pieces of information. As shown in [9], acquisition of channel state information at the Tx (CSIT) is of crucial importance in all multi-user systems and is particularly non-trivial in a non cooperative system such as the spatial interweave (IW) considered here. The present work includes an inventory of tools needed to render coexistence of the two systems possible. In particular, the difficult problem of CSIT acquisition is addressed. It is shown that the solution relies on Time-Division Duplex (TDD) mode of operation. TDD is desirable since, in theory, it allows the exploitation of uplink (UL) - downlink (DL) reciprocity of the underlying radio propagation channel. Using this transmission strategy the transceiver can obtain DL (UL) channel knowledge using an estimate of the UL (DL) channel. In this work, we prove that TDD is not just a *possible* option, but that it is crucial for spatial IW CR to work if unrealistic overheads and communications between the two systems are to be avoided. Unfortunately in practice, even in TDD, the channel reciprocity assumption only holds for part of the overall channel, namely the propagation channel itself. More precisely, in order to exploit channel reciprocity one needs to compensate for the mismatch between the analog Tx/Rx circuitry at both ends: this process is called calibration. The calibration problem is generally addressed through two different approaches denoted as absolute and relative calibration [10]. The first one uses a third-party equipment, used as reference, in order to estimate and compensate the analog Tx/Rx circuitry impairments [11] offline. In the latter approach, UL and DL channel estimates obtained at each side of the communication link are exchanged at a low-rate from which calibration factors are deduced. A

more recent algorithm for relative channel calibration has been proposed in [12].

Another important ingredient in the TDD approach is the complete synchronization of the secondary system to the primary: primary and secondary uplink and downlink slots should coincide.

## II. CONTRIBUTIONS

At the beginning of this paper we provide an extension of the CR paradigms, introduced in [2], to the multi antenna case. We propose some definitions for the *Spatial Underlay*, *Overlay and Interweave* paradigms. The main part of the paper is devoted to the joint optimization of the transmit-receive filters in a spatial interweave cognitive radio channel. We describe the entire communication protocol required to acquire the necessary information, in particular channel state information, at primary and secondary users in a spatial IW CR setting. Since CSI acquisition, at the secondary users, is based on semi-blind channel estimation techniques, it is affected by estimation error. This implies that the ideal zero interference constraint at the primary receiver cannot be satisfied in this more practical scenario. We show, on the other hand, that with proper channel estimation the estimation error influences the primary rate only with a signal to noise ratio (SNR) offset. Hence the secondary transmission does not influence the number of primary transmitted streams (also called degrees of freedom (DoFs)). For this reason we can still include the proposed setting in the spatial interweave scenario. The secondary communication can take place only if the number of transmitted streams, at the primary system, is not deteriorated by the secondary transmission.

The system model studied in this work is not novel. What really differentiates this work with previously proposed solutions, for example [5] and [6], is that no work has studied, up to now, how UL DL channel calibration, required to get channel reciprocity in real TDD transmission, influences transmit and receiver filter design at primary and secondary devices. An even more important result that comes out of our analysis is that even though the opportunistic Tx needs to know (learn) the noise subspace at the primary Rx, the determination of the calibration parameters at the secondary Tx required for its BF design with zero forcing to the primary Rx outputs does not require any cooperation from the primary units, as per the spatial interweave paradigm.

In this paper we use relative calibration to compensate for Tx/Rx electronics [10]. In particular we introduce a new calibration algorithm based on a Total Least Squares (TLS) MIMO technique. In situations where RF crosstalk effects can not be neglected, the proposed algorithm outperforms previously proposed solutions [10],[12].

In addition we extend the results provided for the simple setting with one primary and one secondary pair to the case where the primary network is represented as a  $K$ -user interference channel. In this scenario we assume that the primary network designs the transmit and receive filters according to IA [7]. Then, thanks to IA duality, the secondary pair can blindly estimate the DL receive subspace at all primary receivers from the transmitted signal subspace in the UL

primary communication. Also for this case it is shown how calibration influences the beamformer design at both primary and secondary network and also for the case of multiple primary users calibration between non cooperative users is not required. This concept also applies to the different pairs of primary users: for IA design calibration is required in TDD communications but at the same time each user has to know only its own calibration filter. This information can be acquired by performing calibration between the two sides of each primary user pair.

Finally we show how it is possible to implement in practice the proposed CR setting using a system based on an LTE frame structure.

### III. SPATIAL COGNITIVE RADIO PARADIGMS

In this section we extend the cognitive radio paradigms from [2] to the multi-antenna case. This extension is not as straightforward and unambiguous as it may seem at first.

#### A. Spatial Overlay: MISO/MIMO Interference Channel

In the overlay paradigm, primary and secondary users collaborate. This collaboration could be interpreted at multiple levels, at the level of an exchange of transmit signals (as in network (NW) MIMO), or just at the level of transmit covariances, which in the single antenna case translates to coordinated power control. In [2], the overlay paradigm includes the case of the secondary Tx helping the primary link (by acting as a relay), which is a form of NW MIMO. However, it is also observed in [2] that with the signal exchange being performed over the air, the resulting net DoF are not increased in the single antenna case. In the case of multiple antennas, if we limit cooperation to Tx covariances, this would lead to the exploitation of the multiple antennas for coordinated beamforming to achieve parallel interference-free channels. Coordinated beamforming applies to multiple antennas at the transmit side (MISO Interference Channel). In the case of multiple antennas at the receivers, we can have coordinated receivers. The case of the coordination of the multiple antennas on both sides corresponds to the (noisy) MIMO Interference Channel. The recent Authorized Shared Access (ASA) proposal by Qualcomm and Nokia fits in the realm of overlay cognitive radio.

#### B. Spatial Underlay:

In the traditional underlay paradigm, interference caused by a secondary transmitter to a primary receiver is acceptable as long as it remains under a maximum tolerance level. One possible definition of spatial underlay then would be that the primary receiver, equipped with multiple antennas, allows secondary interference as long as it has enough antennas to handle (suppress) it. Hence the primary receiver needs to be active. So, the primary receiver allows an interference subspace of maximum dimension equal to the excess of its number of antennas over the number of primary streams it needs to receive. The primary system is secondary-aware. Of course, potentially multiple secondary transmitters need to align the interference caused to primaries in subspaces of limited dimension.

#### C. Spatial Interweave:

In the interweave paradigm, the primary system should not be disturbed at all, and is not required to exhibit any cooperation with the secondary systems. So in a spatial interweave version, with multiple primary receive antennas also, the secondary systems need to zero-force to all primary receive antennas individually. In this case there is still room for secondary transmission if the secondary transmitters have more antennas than the combined primary receivers. This setting can be extended to the case, considered here, in which the secondary transmitter design its beamformer such that it zero-forces the interference that is causes at the output of the primary receiver. The advantage of this solution is that the number of degrees of freedom available for secondary transmission,  $d_2 \leq N_2 - d_1$  (where  $N_i$  and  $d_i$  are the number of antennas and DoF at the  $i$ -th user) is higher compared to the setting where the interference is zero-forced to all the primary antennas,  $d_2 \leq N_2 - N_1$ . In the case of a single user MIMO primary link (and spatially white noise), the Tx and Rx filter subspaces on a give primary side (e.g. the mobile terminal) are identical, not only at high SNR but also at finite SNR. The main difference between the spatial underlay and interweave paradigms is that in the former the primary receiver is aware of the secondary transmission and adjusts its Rx filter accordingly. The spatial interweave paradigm requires significant CSIT and can be reciprocity based in TDD, or location based in the case of Line of Sight (LoS) secondary-primary cross channels. In the LoS case, the number of primary receive antennas becomes irrelevant (assuming they are in the far field from the secondary). In a non-LOS case, the secondary transmitter needs to have more antennas than the combined number of propagation paths to all primary receivers.

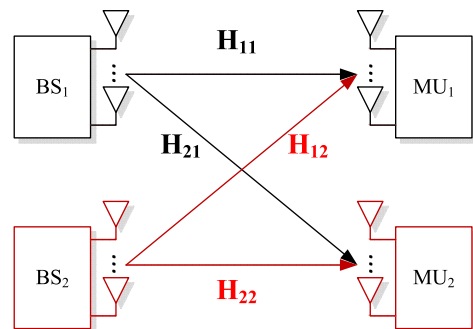


Fig. 1: Downlink Spatial Interweave Channel

### IV. SYSTEM MODEL

We focus on the MIMO interference channel where two point-to-point bidirectional links communicate using a TDD transmission scheme. Even if our work can be applied to a more general system, to simplify the notation we will refer to a primary link composed of a licensee Base Station ( $BS_1$ ) that communicates with the respective Mobile User ( $MU_1$ ) ignoring completely the presence of a secondary transmission in its vicinity. At the same time a cognitive Base Station ( $BS_2$ ) tries to opportunistically communicate with a cognitive Mobile User ( $MU_2$ ) without degrading the licensee's communication.

The key assumption in this work is the lack of cooperation among the two systems, primary and secondary. We assume that all the information that the secondary system needs, such as synchronization and primary communication parameters, to design its communication strategy, is acquired listening the over-the-air communication between primary BS and MU. The knowledge of the communication standard used in the legacy system helps the cognitive users to get useful pieces of information, for example: pilot placement, transmission bandwidth, primary cell ID, frame timing, can be acquired listening to the public control channels of the primary system.

In this system model we assume that  $BS_1$  and  $MU_1$  are both equipped with  $N_1$  antennas while  $BS_2$  and  $MU_2$  have both  $N_2$  antennas. The cognitive radio setting described so far can be used to easily depict the coexistence between a macro-cell and a cognitive femto-cell. The results that we present in this paper can be easily generalized for the case of terminals with an arbitrary number of Tx and Rx antennas,  $M_i$ ,  $N_j$ ,  $i, j = \{1, 2\}$  respectively. We focus on the case where the opportunistic users have a number of antennas greater than or equal to the primary users  $N_2 \geq N_1$ . We denote with  $(\cdot)$  the quantities in the UL transmission, then matrices  $\mathbf{H}_{ij}$  and  $\bar{\mathbf{H}}_{ij} \in \mathbb{C}^{N_i \times N_j}$  are, respectively, the DL and UL channel matrices from transmitter  $j$  to receiver  $i$ , where  $i, j \in \{1, 2\}$ . The entries of these matrices are *i.i.d.* complex Gaussian random variables  $\mathcal{N}(0, 1)$ . We assume that all channels follow a block-fading model having a coherence time of  $T$  symbol intervals without variations. This corresponds to assuming that the channel remains constant for a sufficient number of TDD slots. We also assume that the transmitted signals, in both systems, are drawn from a Gaussian distribution with zero mean and unit variance. Under this assumption we can consider that both useful and interference signals (generated at both systems) have a Gaussian distribution. This assumption, despite being not realistic, gives us the possibility to simplify some mathematical analysis but, at the same time, provides some useful bounds on the real quantities. In a TDD transmission scheme, assuming perfect Tx/Rx calibration, the UL channel is the transpose of the relative downlink channel [10] thanks to channel reciprocity:

$$\bar{\mathbf{H}}_{ij} = \mathbf{H}_{ji}^H \quad (1)$$

where the complex conjugate (obtained by conjugating all signals) in the Hermitian transpose is added for notational convenience. Thus an UL channel estimate can be used for designing the transmit beamformer in the DL communication.

## V. TRANSMISSION TECHNIQUES AND CHANNEL ESTIMATION

In the Interweave cognitive scenario, licensee (primary) systems are not aware of the presence of secondary systems which should ideally cause no interference. The primary Tx is therefore assumed to be a Single User MIMO link (SU-MIMO). In this system the transmitter and receiver filters are designed in order to maximize the transmission rate. The capacity-achieving solution is based on a Beamforming matrix obtained from the singular value decomposition (SVD)

of the channel matrix combined with Water-Filling power allocation [13]. Assuming low-rank Tx, the primary link can be decomposed into a signal and a complementary (noise) subspace,

$$\mathbf{H} = \mathbf{U}\mathbf{\Delta}\mathbf{V}^H = [\mathbf{U}_s \mathbf{U}_n] \begin{bmatrix} \mathbf{\Delta}_s & \\ & \mathbf{\Delta}_n \end{bmatrix} \begin{bmatrix} \mathbf{V}_s^H \\ \mathbf{V}_n^H \end{bmatrix} \quad (2)$$

where subscripts  $s$  or  $n$  refer to signal subspace and noise subspace respectively. The matrices  $\mathbf{U}$  and  $\mathbf{V}$  are unitary matrices and  $\mathbf{\Delta}$  is a diagonal matrix that contains the singular values of the channel matrix. In order to waterfill in UL and DL, both  $BS_1$  and  $MU_1$  must have complete knowledge of the primary channel and Rx noise variances. This information can be obtained partially through TDD reciprocity (pilots for channel estimation) and partially through (unavoidable) feedback.

In the interweave scenario unlicensed users must transmit without disturbing the licensed transmission. At low to medium SNR the primary transmitters are expected to exploit a limited number of channel modes, the opportunistic transmitter can beamform its signal in the noise subspace of the licensed communication. This has been labeled as interference alignment technique in [8]. To adapt its communication, the secondary Tx has to know the signal subspace of the primary Rx. As discussed in the following this subspace can be learnt using channel reciprocity and with an opportunistic exploitation of the signaling exchanged between primary system terminals.

All TDD frames in both UL and DL are composed of two time segments, one comprising possibly multiple data streams and the second includes pilots for channel estimation. In the primary link only the data part of the frame is beamformed but not the pilots. This implies that they span the entire channel space. On the other hand in the cognitive transmission, pilots are also beamformed, thus ensuring that they do not interfere with the primary transmission. We assume that the secondary TDD slots are aligned with the primary using classical spectrum sensing and synchronization techniques. Further synchronization can be achieved exploiting the information transmitted in the broadcast channel in the primary communication.

### A. First TDD Slot

In this first slot all devices in the system should start to acquire the channel state information they need. In particular the licensed BS transmits without knowledge of the downlink channel and therefore cannot beamform.  $MU_1$  can estimate the DL channel matrix  $\mathbf{H}_{11}$  using pilots sequence of length:  $T_T^P \geq N_1$ .

During this phase, cognitive users in particular  $MU_2$ , can use the pilot symbols of the primary communication, and the knowledge about the primary communication standard, to opportunistically estimate the cross channel  $\mathbf{H}_{21}$ .

### B. Second TDD Slot

$MU_1$  now knows the downlink channel matrix and hence it can design the beamforming subspace  $\mathbf{T}_{MU_1} \in \mathbb{C}^{N_1 \times d_1}$  using channel reciprocity in equation (1). With  $d_1$  we represent

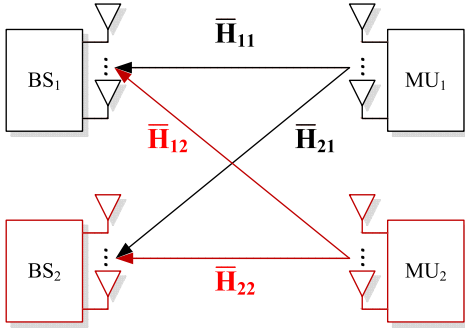


Fig. 2: Uplink Channel

the number of transmitted streams, obtained using WF, and is equal to the signal subspace dimensions. In the same UL frame  $BS_1$  can estimate the UL channel, as done for the DL channel in the previous slot, exploring pilot symbols incorporated in each time segment.

$MU_1$  calculates its Tx beamformer with an SVD of the UL channel,  $\bar{\mathbf{H}}_{11} = \mathbf{H}_{11}^H = \mathbf{V}_1 \mathbf{\Delta}_{11} \mathbf{U}_1^H$ , derived using channel reciprocity. The Tx beamformer matrix  $\mathbf{T}_{MU_1} = \mathbf{U}_{1,s}$  is obtained taking  $d_1$  columns of  $\mathbf{U}_1$  according to the WF solution.  $BS_1$  design its Rx filter as  $\mathbf{R}_{BS_1} = \mathbf{V}_{1,s}^H \in \mathbb{C}^{d_1 \times N_1}$  from the SVD of the UL channel. The signal at the output of the receiver filter at  $BS_1$  is written as

$$\begin{aligned} \bar{\mathbf{r}}_1 &= \mathbf{R}_{BS_1} \bar{\mathbf{H}}_{11} \mathbf{T}_{MU_1} \bar{\mathbf{s}}_1 + \mathbf{R}_{BS_1} \bar{\mathbf{n}}_1 \\ &= \mathbf{V}_{1,s}^H \mathbf{H}_{11}^H \mathbf{U}_{1,s} \bar{\mathbf{s}}_1 + \mathbf{V}_{1,s}^H \bar{\mathbf{n}}_1 = \mathbf{\Delta}_{11,s} \bar{\mathbf{s}}_1 + \bar{\mathbf{n}}_1' \end{aligned} \quad (3)$$

where  $\bar{\mathbf{s}}_1 \in \mathbb{C}^{d_1 \times 1}$  is the transmitted signal vector and  $\mathbf{\Delta}_{11,s}$  is the diagonal matrix containing singular values of  $\mathbf{H}_{11}^H$  corresponding to the signal subspace. Vector  $\bar{\mathbf{n}}_1'$  is the post-processed noise vector that, thanks to the unitary propriety of the Rx filter, preserves the original Gaussian distribution with zero means and variance  $\sigma_1^2$ .

At  $BS_2$  the  $N_2 \times 1$  Rx signal is given by

$$\bar{\mathbf{y}}_2 = \mathbf{H}_{12}^H \mathbf{T}_{MU_1} \bar{\mathbf{s}}_1 + \bar{\mathbf{n}}_2 = \mathbf{H}_{12}^H \mathbf{U}_{1,s} \bar{\mathbf{s}}_1 + \bar{\mathbf{n}}_2. \quad (4)$$

Assuming sufficient data samples at  $BS_2$ , we can obtain a consistent estimate of the primary Tx signal subspace from the autocorrelation matrix of the Rx signal  $\mathbf{R}_{\bar{\mathbf{y}}_2 \bar{\mathbf{y}}_2} = E\{\bar{\mathbf{y}}_2 \bar{\mathbf{y}}_2^H\}$ . In practice in the blind subspace estimation procedure we use the sample covariance matrix

$$\hat{\mathbf{R}}_{\bar{\mathbf{y}}_2 \bar{\mathbf{y}}_2} = \frac{1}{T_E} \sum_{t=1}^{T_E} \bar{\mathbf{y}}_2[t] \bar{\mathbf{y}}_2^H[t]. \quad (5)$$

where  $T_E$  represents the number of primary data samples used for subspace estimation. From the eigenvalue decomposition of (5),  $\hat{\mathbf{R}}_{\bar{\mathbf{y}}_2 \bar{\mathbf{y}}_2} = \hat{\mathbf{U}} \hat{\mathbf{\Lambda}} \hat{\mathbf{U}}^H$ , we can estimate the signal space dimension  $\hat{d}_1$  using the information theoretic criteria described in [14]. Here we assume that the estimate of the primary signal dimension  $\hat{d}_1$  is done without errors to simplify the analysis. Then we can determine the primary noise subspace estimate  $\hat{\mathbf{U}}_n$  from the last  $N_2 - \hat{d}_1$  eigenvector of  $\hat{\mathbf{U}}$ . This approach gives a consistent estimate for sufficient data samples  $T_E$  that is affected by an estimation error [15] that can be characterized as

$$\tilde{\mathbf{U}}_n = \mathbf{U}_n - \hat{\mathbf{U}}_n = \mathbf{K}_2^{H\dagger} \mathbf{N}^H \mathbf{U}_n \quad (6)$$

where  $\mathbf{N} = [\bar{\mathbf{n}}_2[1], \dots, \bar{\mathbf{n}}_2[T_E]]$  and

$$\mathbf{K}_2 = \mathbf{H}_{12}^H \mathbf{T}_{MU_1} \underbrace{[\bar{\mathbf{s}}_1[1], \dots, \bar{\mathbf{s}}_1[T_E]]}_{\tilde{\mathbf{K}}_2}. \quad (7)$$

$\mathbf{A}^\dagger$  denotes the pseudo-inverse of  $\mathbf{A}$ . In a similar way it is possible to determine the estimation error of the primary signal subspace  $\mathbf{U}_s$  [16]. Due to channel reciprocity the Tx and Rx signal subspace in the primary link are the same, hence the knowledge of the primary Tx subspace is sufficient to determine the Rx signal subspace at  $MU_1$ .

Knowing  $\mathbf{U}_s$ ,  $BS_2$  can send at most  $d_2 \leq N_2 - \hat{d}_1$  streams while ensuring its signal lies in the noise subspace at the primary Rx. This implies that

$$\mathbf{R}_{MU_1} \mathbf{H}_{12} \mathbf{T}_{BS_2} = \mathbf{0} \implies \mathbf{T}_{BS_2} \subseteq (\mathbf{R}_{MU_1} \mathbf{H}_{12})^\perp. \quad (8)$$

The equation above says that  $\mathbf{T}_{BS_2}$  belongs to the subspace spanned by  $\text{span}(\mathbf{A}^\perp)$ , where  $\mathbf{A}^\perp$  represents the orthogonal complement of the row space of the matrix  $\mathbf{A}$ . In our case a possible choice for the secondary transmit filter is  $\mathbf{T}_{BS_2} = \hat{\mathbf{U}}_n$ .

Including the receiver at  $MU_1$  in the definition of  $\mathbf{T}_{BS_2}$  has the advantage that in low to medium SNR regimes of the primary link, where the primary Tx sends only  $d_1 < N_1$  of the total available signaling dimension  $N_1$ , the secondary Tx can (opportunistically) transmit at most  $d_2 \leq N_2 - d_1$  streams. On the other hand in the high SNR region, when the primary link uses up its entire DoF for spatial multiplexing, the secondary can always transmit  $d_2 \leq N_2 - N_1$  streams.

### C. Third TDD Slot

From this TDD time slot onwards starts the steady state of the system. This means that also the cognitive BS starts to transmit to  $MU_2$ .  $BS_1$  constructs its beamforming subspace using SVD of the DL channel matrix  $\mathbf{H}_{11}$ . Then the transmit beamformer is  $\mathbf{T}_{BS_1} = \mathbf{V}_{1,s}$ , and  $MU_1$  uses as receiver the matrix  $\mathbf{R}_{MU_1} = \mathbf{U}_{1,s}^H$ . In this slot also the opportunistic BS starts to transmit its data symbols, hence the received signal at primary MU is

$$\mathbf{y}_1 = \mathbf{H}_{11} \mathbf{T}_{BS_1} \mathbf{s}_1 + \mathbf{H}_{12} \mathbf{T}_{BS_2} \mathbf{s}_2 + \mathbf{n}_1 \quad (9)$$

where  $\mathbf{s}_2$  is the  $d_2 \times 1$  DL transmitted signal from secondary BS. After the application of the Rx filter at  $MU_1$  the interference contribution, due to secondary transmission, disappears since, by construction in (8), it lies in the primary receiver noise subspace.

On the other hand  $MU_2$  receives signals from both  $BS_1$  and  $BS_2$ :

$$\mathbf{y}_2 = \mathbf{H}_{22} \mathbf{T}_{BS_2} \mathbf{s}_2 + \mathbf{H}_{21} \mathbf{T}_{BS_1} \mathbf{s}_1 + \mathbf{n}_2. \quad (10)$$

$MU_2$  needs to estimate the noise and signal subspaces of the primary communication to design its beamformer. This can be done using semi-blind estimation procedure. The definition semi-blind comes from the fact that part of the information is obtained using usual training and the remaining information comes from blind subspace estimation.

Using the beamformed pilots incorporated into the secondary

data frame, the secondary receiver can estimate the cascade of secondary direct channel and beamformer  $\mathbf{H}_{22}\mathbf{T}_{BS_2}$  that has dimensions  $N_2 \times d_2$ . The training length should now satisfy:  $T_T^S \geq d_2$ . Once the secondary direct link has been estimated,  $MU_2$  has to estimate the signal and noise subspaces of the primary DL transmission. To accomplish this task the cognitive device can reconstruct the transmitted signal from  $BS_2$  during the secondary pilot transmission and then subtract it from the Rx signal vector:

$$\mathbf{y}'_2 = \mathbf{y}_2 - \widehat{\mathbf{H}_{22}\mathbf{T}_{BS_2}}\mathbf{s}_2 = \mathbf{H}_{21}\mathbf{T}_{BS_1}\mathbf{s}_1 + \mathbf{n}_2. \quad (11)$$

In (11) we assume the estimate  $\widehat{\mathbf{H}_{22}\mathbf{T}_{BS_2}}$  is obtained without error to simplify the analysis.

Using the reconstructed signal  $\mathbf{y}'_2$   $MU_2$  determines the signal and noise subspaces, denoted as  $\mathbf{V}_s$  and  $\mathbf{V}_n$  respectively, of the primary downlink signal using second-order statistics (SOS). This estimation procedure can follow the same steps as the one proposed in section V-B. The estimated noise and signal subspace will be also affected by similar error contribution of (6).

Finally  $MU_2$  designs its beamformer subspace such that it creates zero interference at the primary BS:

$$\mathbf{T}_{MU_2} \subseteq (\mathbf{R}_{BS_1}\mathbf{H}_{21}^H)^\perp \quad (12)$$

a possible choice is  $\mathbf{T}_{MU_2} = \mathbf{V}_n$ .

#### D. Fourth TDD slot

In this slot all nodes have the required knowledge to transmit to corresponding receivers.

Even though  $BS_1$  receives no interference from the secondary transmission,  $BS_2$ , nevertheless receives interference from the primary communication. To suppress primary interference a standard linear MIMO receiver can be used.

#### E. Secondary Link Optimization

The secondary link beamformer subspace, designed to cause zero interference at the primary receivers, is invariant to a multiplication by a square  $d_2 \times d_2$  matrix  $\mathbf{Q}_{BS_2}$ :  $\mathbf{T}_{BS_2}\mathbf{Q}_{BS_2} \in \text{span}(\mathbf{T}_{BS_2})$  hence  $\mathbf{R}_{MU_1}\mathbf{H}_{12}\mathbf{T}_{BS_2}\mathbf{Q}_{BS_2} = \mathbf{0}$ . The remaining degrees of freedom in  $\mathbf{Q}_{BS_2}$  can be used for the optimization of the secondary link communication.

The received signal at  $MU_2$  is given in (10). One possible optimization criterion for matrix  $\mathbf{Q}_{BS_2}$  is the maximization of the secondary user's rate. This means that we need to solve the following optimization problem:

$$\begin{aligned} \max_{\mathbf{Q}_{BS_2}} \log & \left| \mathbf{I} + \mathbf{Q}_{BS_2}^H \underbrace{\mathbf{T}_{BS_2}^H \mathbf{H}_{2,2}^H \mathbf{R}_{int}^{-1} \mathbf{H}_{2,2} \mathbf{T}_{BS_2}}_{\mathbf{K}} \mathbf{Q}_{BS_2} \right| \\ \text{s.t.} \quad & \text{Tr}(\mathbf{T}_{BS_2} \mathbf{Q}_{BS_2} \mathbf{Q}_{BS_2}^H \mathbf{T}_{BS_2}^H) \leq P_2 \end{aligned} \quad (13)$$

where  $P_2$  represents the transmit power constraint at the secondary link and  $\mathbf{R}_{int} = \mathbf{H}_{2,1}\mathbf{T}_{BS_1}\mathbf{S}_1\mathbf{T}_{BS_1}^H\mathbf{H}_{2,1}^H + \sigma_n^2\mathbf{I}$  is the interference plus noise covariance matrix with  $\mathbf{S}_1 = \mathbb{E}\{\mathbf{s}_1\mathbf{s}_1^H\}$ . We use the common notation  $|\mathbf{A}| = \det(\mathbf{A})$ . The solution of this problem corresponds to the traditional water-filling solution in colored noise since in the noise covariance matrix  $\mathbf{R}_{int}$  we accounted also the interference due to the primary communication.

## VI. RATE LOSS DUE TO (BLIND) PRIMARY TRANSMIT SUBSPACE ESTIMATION

As described in section V-B the blind estimation of the signal and noise subspace of the primary transmission is affected by estimation error (6). This implies that when the secondary transmitter sends data using a BF based on the estimated noise subspace some interference leaks in the signal subspace at the primary receiver creating interference. Naturally this interference determines some loss in terms of primary achievable rate.

The received signal at primary MU (9), after the Rx filter  $\mathbf{R}_{MU_1}$ , is:

$$\mathbf{r}_1 = \mathbf{R}_{MU_1}\mathbf{H}_{11}\mathbf{T}_{BS_1}\mathbf{s}_1 + \mathbf{R}_{MU_1}\mathbf{H}_{12}\tilde{\mathbf{U}}_n\mathbf{s}_2 + \mathbf{R}_{MU_1}\mathbf{n}_1.$$

Denoting with  $\mathbf{R}_{\bar{1}}$  the interference plus noise covariance matrix:

$$\mathbf{R}_{\bar{1}} = \underbrace{\mathbf{R}_{MU_1}\mathbf{H}_{12}\tilde{\mathbf{U}}_n\mathbf{Q}_{BS_2}^2\tilde{\mathbf{U}}_n^H\mathbf{H}_{12}^H\mathbf{R}_{MU_1}^H}_{\mathbb{I}_1} + \underbrace{\sigma_1^2\mathbf{I}}_{\mathbf{R}_{n_1}}$$

where  $\mathbf{Q}_{BS_2}^2 = \mathbf{Q}_{BS_2}\mathbf{Q}_{BS_2}^H$  denotes the transmit covariance matrix at the secondary transmitter. The rate at primary MU becomes:

$$\tilde{\mathcal{R}}_{MU_1} = \log \left| \mathbf{I} + \underbrace{\mathbf{R}_{MU_1}\mathbf{H}_{11}\mathbf{T}_{BS_1}\mathbf{S}_1\mathbf{T}_{BS_1}^H\mathbf{H}_{11}^H\mathbf{R}_{MU_1}^H}_{\mathbf{D}_{MU_1}} \mathbf{R}_{\bar{1}}^{-1} \right| \quad (14)$$

The average rate loss due to the estimation error can be determined simply as:

$$\begin{aligned} \Delta\mathcal{R} &= \mathbb{E}\mathcal{R}_{MU_1} - \mathbb{E}\tilde{\mathcal{R}}_{MU_1} \\ &= \mathbb{E} \log \left| \mathbf{I} + \mathbf{D}_{MU_1} \mathbf{R}_{n_1}^{-1} \right| - \mathbb{E} \log \left| \mathbf{I} + \mathbf{D}_{MU_1} \mathbf{R}_{\bar{1}}^{-1} \right| \\ &= \mathbb{E} \log \left| \mathbf{I} + \mathbf{D}_{MU_1} \mathbf{R}_{n_1}^{-1} \right| \\ &\quad - \mathbb{E} \log \left| \mathbf{I} + (\mathbf{D}_{MU_1} + \mathbb{I}_1) \mathbf{R}_{n_1}^{-1} \right| + \mathbb{E} \log \left| \mathbf{I} + \mathbb{I}_1 \mathbf{R}_{n_1}^{-1} \right| \\ &\stackrel{(a)}{\leq} \mathbb{E} \log \left| \mathbf{I} + \mathbb{I}_1 \mathbf{R}_{n_1}^{-1} \right| \stackrel{(b)}{\leq} \log \left| \mathbf{I} + \mathbb{E}\{\mathbb{I}_1\} \mathbf{R}_{n_1}^{-1} \right| \end{aligned} \quad (15)$$

where (a) is due to the fact that  $|\mathbf{I} + (\mathbf{D}_{MU_1} + \mathbb{I}_1)\mathbf{R}_{n_1}^{-1}|$  dominates<sup>1</sup>  $|\mathbf{I} + \mathbf{D}_{MU_1}\mathbf{R}_{n_1}^{-1}|$ . Finally (b) comes from the Jensen inequality. To determine theoretically the value of the rate loss we need to compute the expectation  $\mathbb{E}\{\mathbb{I}_1\}$ :

$$\begin{aligned} \mathbb{E}\{\mathbb{I}_1\} &= \mathbb{E}\{\mathbf{R}_{MU_1}\mathbf{H}_{12}\tilde{\mathbf{U}}_n\mathbf{Q}_{BS_2}^2\tilde{\mathbf{U}}_n^H\mathbf{H}_{12}^H\mathbf{R}_{MU_1}^H\} \\ &= \mathbb{E}\{\mathbf{R}_{MU_1}\mathbf{H}_{12}\mathbf{K}_2^H\mathbf{N}^H\mathbf{U}_n\mathbf{Q}_{BS_2}^2\mathbf{U}_n^H\mathbf{N}\mathbf{K}_2\mathbf{H}_{12}^H\mathbf{R}_{MU_1}^H\} \end{aligned} \quad (16)$$

the noise samples in  $\mathbf{N}$  are iid then  $\mathbb{E}\{\mathbf{N}^H\mathbf{U}_n\mathbf{Q}_{BS_2}^2\mathbf{U}_n^H\mathbf{N}\} = \sigma_2^2\text{Tr}\{\mathbf{Q}_{BS_2}^2\} = \sigma_2^2P_2$ . Finally equation (16) becomes

$$\begin{aligned} \mathbb{E}\{\mathbb{I}_1\} &= \sigma_2^2P_2\mathbb{E}\left\{ \mathbf{U}_{1,s}^H\mathbf{H}_{12}\left[ \mathbf{H}_{12}^H\mathbf{U}_{1,s}(\mathbf{U}_{1,s}^H\mathbf{H}_{12}\mathbf{H}_{12}^H\mathbf{U}_{1,s})^{-1} \times \right. \right. \\ &\quad \left. \left. \left( \tilde{\mathbf{K}}_2\tilde{\mathbf{K}}_2^H \right)^{-1} \left( \tilde{\mathbf{K}}_2\tilde{\mathbf{K}}_2^H \right) \left( \tilde{\mathbf{K}}_2\tilde{\mathbf{K}}_2^H \right)^{-1} \times \right. \right. \\ &\quad \left. \left. \left( \mathbf{U}_{1,s}^H\mathbf{H}_{12}\mathbf{H}_{12}^H\mathbf{U}_{1,s} \right)^{-1} \mathbf{U}_{1,s}^H\mathbf{H}_{12} \right] \mathbf{H}_{12}^H\mathbf{U}_{1,s} \right\} \\ &= \sigma_2^2P_2\mathbb{E}\left\{ \left( \tilde{\mathbf{K}}_2\tilde{\mathbf{K}}_2^H \right)^{-1} \right\}. \end{aligned} \quad (17)$$

<sup>1</sup>This can be proved using the Mikowski inequality that for positive semi-definite matrices  $\mathbf{A}$  and  $\mathbf{B}$ :  $|\mathbf{A} + \mathbf{B}|^{\frac{1}{n}} \geq |\mathbf{A}|^{\frac{1}{n}} + |\mathbf{B}|^{\frac{1}{n}}$  then  $|\mathbf{A} + \mathbf{B}| \geq |\mathbf{A}| + |\mathbf{B}|$ . Introducing  $\log()$  function we get for semi-definite matrices:  $\log |\mathbf{A} + \mathbf{B}| \geq \log(|\mathbf{A}| + |\mathbf{B}|) \geq \log |\mathbf{A}|$ . Denoting with  $\mathbf{A} = \mathbf{I} + \mathbf{D}_{MU_1}\mathbf{R}_{n_1}^{-1}$  and  $\mathbf{B} = \mathbb{I}_1\mathbf{R}_{n_1}^{-1}$  we complete the proof.

In the equation above we used the property of pseudo-inverse that if  $\mathbf{K}_2$  can be parametrized as  $\mathbf{K}_2 = \mathbf{C}\mathbf{D}$  then  $\mathbf{K}_2^\dagger = \mathbf{D}^H(\mathbf{D}\mathbf{D}^H)^{-1}(\mathbf{C}^H\mathbf{C})^{-1}\mathbf{C}^H$ . According to the definition of  $\tilde{\mathbf{K}}_2$  in (7),  $(\tilde{\mathbf{K}}_2\tilde{\mathbf{K}}_2^H)^{-1}$  is distributed as a complex inverse Wishart matrix:  $\mathcal{W}_{d_1}^{C^{-1}}(T_E, \mathbf{S}_1^{-1})$  [17]. Then

$$\mathbb{E}\{\mathbb{I}_1\} = \frac{\sigma_2^2 P_2}{T_E - d_1} \mathbf{S}_1^{-1}$$

we can finally conclude that the upper bound of rate loss is :

$$\Delta\mathcal{R} \leq \log \left| \mathbf{I} + \frac{\sigma_2^2 P_2}{\sigma_1^2 (T_E - d_1)} \mathbf{S}_1^{-1} \right| \quad (18)$$

Assuming that  $\mathbf{S}_1$  is roughly proportional to the primary transmit power  $P_1$ , from (18) we can see that if both powers,  $P_1$  and  $P_2$ , grow at the same rate,  $\Delta\mathcal{R}$  stays constant. This means that at high SNR the estimation error in the noise subspace at secondary devices determines only a loss in terms of SNR offset. On the other hand the multiplexing gain (or degrees of freedom) achievable by the primary users remains constant. The observation above allows us to still label this transmission setting as spatial interweave, since the used spatial dimensions at the primary users are not influenced by the secondary communication. This will be more clear in the simulation results of Section IX.

## VII. BEAMFORMING DESIGN WITH CHANNEL CALIBRATION

Up to this point we have considered UL and DL channel to be perfectly reciprocal. In practice this is true only after calibration. In this section we describe the basic principle of relative calibration. Then we introduce the new calibration algorithm for MIMO systems and, finally, we study how calibration filters influence the beamformer design problem.

### A. Uplink Downlink Calibration

In Fig. 3, we model the overall UL and DL channels  $\mathbf{U}_{ii}$ ,  $\mathbf{D}_{ii}$  [10]. These overall channel responses, which are estimated in the digital (sampled) domain, go from before the Digital to Analog Converter (DAC) at the transmitter to after the Analog to Digital Converter (ADC) in the receiver. They can

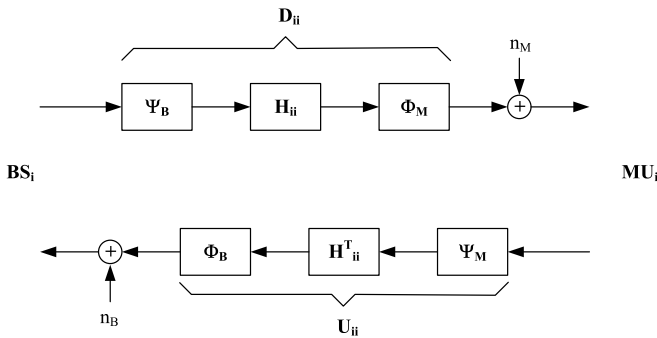


Fig. 3: Reciprocity Model.

be modeled as the following cascades

$$\mathbf{U}_{ii} = \Phi_B \mathbf{H}_{ii}^T \Psi_M; \quad \mathbf{D}_{ii} = \Phi_M \mathbf{H}_{ii} \Psi_B \quad (19)$$

where all matrices are  $N_i \times N_i$ . In the downlink,  $\mathbf{H}_{ii}$  is the actual propagation channel, going from the BS Tx/Rx switches (where Tx and Rx RF chains connect to the same antenna) to the MU Tx/Rx switches (for each antenna),  $\Psi_B$  represents the response from BS DACs to BS Tx/Rx switches (for all antennas), and  $\Phi_M$  represents the response from the MU Tx/Rx switches to the MU ADCs. We have similar components in the uplink where now the MIMO propagation channel response is  $\mathbf{H}_{ii}^T$  due to reciprocity. All representations are in baseband and we assume a frequency flat channel or to work at a specific subcarrier in OFDM. The matrices  $\Psi$ ,  $\Phi$  represent absolute calibration factors between the digital and EM domains. Now the DL channel can be expressed as function of the UL channel as:

$$\mathbf{D}_{ii} = \underbrace{\Phi_M \Psi_M^{-T}}_{\mathbf{P}_{MU_i}} \mathbf{U}_{ii}^T \underbrace{\Phi_B^{-T} \Psi_B}_{\mathbf{P}_{BS_i}} \quad (20)$$

The relative calibration matrices  $\mathbf{P}_{MU_i}$  and  $\mathbf{P}_{BS_i}$  only depend on the RF chains at the respective sides. The objective of relative calibration is to find these matrices using estimates of the UL and DL channels obtained through classical training and channel feedback operation [10]. Calibration requires an UL to DL and another DL to UL training phase between users. Many techniques in MIMO CR assume directly the reciprocity without a calibration process. Although the full calibration model of (20) was proposed in [10], a number of simplifications have been introduced in actual calibration factor estimation algorithms introduced in [10] and [12]. One simplification is to assume no antenna coupling in the RF chains, in which case the matrices  $\mathbf{P}_{MU_i}$  and  $\mathbf{P}_{BS_i}$  can be assumed to be diagonal. Other simplifications involve the separate estimation of elements of these matrices or of each matrix in an iterative process, the convergence of which has not been shown. In the sequel, we will describe a new technique to find simultaneously both MIMO calibration matrices.

### B. MIMO Calibration Procedure

We describe the calibration procedure in each BS-MU pair which is executed after a training phase. For calibration by a BS, the BS requires an UL channel estimate, and a DL channel estimate, which is obtained by the corresponding MU in a DL training operation, followed by feedback of the DL channel estimate from MU to BS. So apart from the usual training operations in each direction, also feedback of channel estimates is required.

As previously mentioned, the calibration matrices depend essentially on the Tx and Rx electronic components (see Fig. 4). Subsequently, these component characteristics (stability, impedance, isolation, etc) vary according to the frequency range or the temperature, but they are otherwise time-invariant as mentioned in [10], [18]. Hence, in the sequel we suppose that the calibration matrices ( $\mathbf{P}_{MU_i}$  and  $\mathbf{P}_{BS_i}$ ) vary much more slowly in time than the channel matrices.

To find calibration matrices  $\mathbf{P}_{MU_i}$  and  $\mathbf{P}_{BS_i}$  we propose a MIMO calibration scheme using a total least squares technique (TLS) described in a previous study, where we evaluated the performance and the complexity of reciprocity calibration in



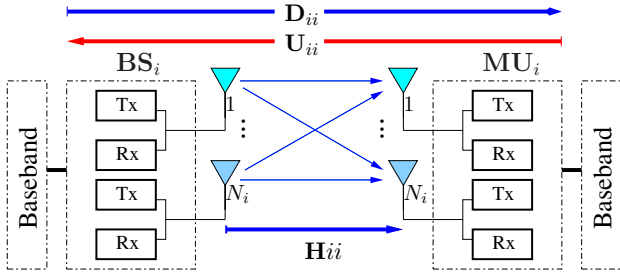


Fig. 4: Overall uplink and downlink channels  $\mathbf{U}_{ii}$  and  $\mathbf{D}_{ii}$ , with the RF front-ends and the electromagnetic channel  $\mathbf{H}_{ii}$ .

a MIMO environment [12], [19]. From this study, we observe that the equation (20) can be rewritten as:

$$\mathbf{P}_{MU_i}^{-1} \mathbf{D}_{ii} - \mathbf{U}_{ii}^T \mathbf{P}_{BS_i} = \mathbf{0}. \quad (21)$$

Then, the calibration problem consists in determining  $\mathbf{P}_{MU_i}$  and  $\mathbf{P}_{BS_i}$  minimizing the following distance:

$$\underset{\{\mathbf{P}_{MU_i}, \mathbf{P}_{BS_i}\}}{\operatorname{argmin}} \quad \|\mathbf{P}_{MU_i}^{-1} \mathbf{D}_{ii} - \mathbf{U}_{ii}^T \mathbf{P}_{BS_i}\|_F^2, \quad (22)$$

which is equivalent to:

$$\underset{\{\mathbf{P}_{MU_i}, \mathbf{P}_{BS_i}\}}{\operatorname{argmin}} \quad \|\operatorname{vec}(\mathbf{P}_{MU_i}^{-1} \mathbf{D}_{ii}) - \operatorname{vec}(\mathbf{U}_{ii}^T \mathbf{P}_{BS_i})\|^2$$

where the operator  $\mathbf{a} = \operatorname{vec}(\mathbf{A})$  transforms the matrix  $\mathbf{A}^{N \times M}$  into a vector  $\mathbf{a}$  of dimensions  $NM \times 1$  by stacking the consecutive columns. Given the following relation  $\operatorname{vec}(A_{M \times N} B_{N \times P}) = (B^T \otimes \mathbf{I}_M) \operatorname{vec}(A) = (\mathbf{I}_P \otimes A) \operatorname{vec}(B)$ , we write:

$$\begin{aligned} & \operatorname{vec}(\mathbf{P}_{MU_i}^{-1} \mathbf{D}_{ii}) - \operatorname{vec}(\mathbf{U}_{ii}^T \mathbf{P}_{BS_i}) \\ &= (\mathbf{D}_{ii}^T \otimes \mathbf{I}_{N_i}) \operatorname{vec}(\mathbf{P}_{MU_i}^{-1}) - (\mathbf{I}_{N_i} \otimes \mathbf{U}_{ii}^T) \operatorname{vec}(\mathbf{P}_{BS_i}). \end{aligned} \quad (23)$$

Hence, a solution to equation (22) is to find  $\mathbf{P}_{MU_i}^{-1}$  and  $\mathbf{P}_{BS_i}$  such that:

$$\begin{aligned} & \underbrace{\begin{bmatrix} (\mathbf{D}_{ii}^T \otimes \mathbf{I}_{N_i}) & -(\mathbf{I}_{N_i} \otimes \mathbf{U}_{ii}^T) \end{bmatrix}}_{\mathbf{Z}} \underbrace{\begin{bmatrix} \operatorname{vec}(\mathbf{P}_{MU_i}^{-1}) \\ \operatorname{vec}(\mathbf{P}_{BS_i}) \end{bmatrix}}_{\mathbf{C}} \\ &= \mathbf{Z}_{N_i^2 \times 2N_i^2} \mathbf{C}_{2N_i^2 \times 1} = \mathbf{0}_{N_i^2 \times 1} \end{aligned} \quad (24)$$

We observe that it is possible to find the calibration parameters in  $\mathbf{C}$  if  $\mathbf{Z}$  is a square or tall matrix. Moreover, in practice, the channel matrices are not perfect, since they are determined with an estimation error. To consider the channel estimation error in further analysis on UL/DL MIMO channels, in (24) we introduce a perturbation model on  $\mathbf{Z}$ , leading to a Total Least-Squares (TLS) formulation. Next, assuming the calibration parameters constant for  $K_{max}$  channel measurements, we can find reliable calibration matrices by over-parameterizing the equation (24) with  $K < K_{max}$  channel measurements over the time ( $\mathbf{Z}_K = [\mathbf{Z}^{1T}, \dots, \mathbf{Z}^{KT}]^T$ ). Finally the concatenation of these measurements yields the relation  $KN_i^2 > 2N_i^2$  between the numbers of rows of  $\mathbf{Z}$  and  $\mathbf{C}$ :

$$\mathbf{Z}_K \{KN_i^2 \times 2N_i^2\} \mathbf{C}_{2N_i^2 \times 1} = \mathbf{0}_{KN_i^2 \times 1}. \quad (25)$$

The over-parametrized TLS problem can now be rewritten as:

$$\underset{\{\mathbf{C}, \Delta \mathbf{Z}_K\}}{\operatorname{argmin}} \quad \|\Delta \mathbf{Z}_K\|_F \text{ s.t. } (\mathbf{Z}_K + \Delta \mathbf{Z}_K) \mathbf{C} = \mathbf{0}_{KN_i^2 \times 1} \quad (26)$$

where  $\Delta \mathbf{Z}_K$  represents the compensation of the error intro-

duced by the noisy MIMO channel matrices ( $\mathbf{D}_{ii}$  and  $\mathbf{U}_{ii}$ ) in the matrix  $\mathbf{Z}_K$ . Given the SVD decomposition of:  $\mathbf{Z}_K = \mathbf{U} \mathbf{D} \mathbf{V}^H$ ,  $\mathbf{V}$  can be read as an orthogonal basis consisting of the right singular vectors of  $\mathbf{Z}_K$ . The kernel of  $\mathbf{Z}$  is spanned by the last column of  $\mathbf{V}$ , then the TLS solution of equation (26) lies in this column and is given by:  $\hat{\mathbf{C}} = -\mathbf{V}_{(2N_i^2)}$ , where  $\mathbf{V}_{(2N_i^2)}$  represents the last column vector of  $\mathbf{V}$ , see [19] (note that  $\mathbf{C}$  needs to be determined only up to a scalar scale factor).

Introducing an SVD decomposition in the solution of the TLS problem has a high computational complexity for large  $\mathbf{Z}$  matrices (e.g. when using a frequency domain approach with multiple subcarriers). Nevertheless, in our case only one singular vector needs to be determined and the MIMO system model typically takes small size channel matrices ( $N_i \leq 8$ , and  $K$  can be kept small). Calibration parameters are finally found in equation (24).

In our study, according to the relative calibration principle the question is: "How to calibrate the cross links in a CR system where communication between primary and secondary systems is not allowed?". As we shall see, despite the stringent secondary beamformer requirement (the interference should lie in the crosslink Rx noise subspace) no calibration is required between crosslink Tx-Rx devices. Indeed, note that in case of a crosslink, the reciprocity model (20) becomes

$$\mathbf{D}_{ki} = \mathbf{P}_{MU_k} \mathbf{U}_{ik}^T \mathbf{P}_{BS_i}. \quad (27)$$

This result, which states that the calibration factor on the side of a BS or MU is independent of which unit on the other side the channel is considered to, is a key element to implement in practice spatial interweave CR systems.

It must be noted that in our CR scenario, the calibration phase of secondary link will interfere a little with the primary link (and vice versa) but considering that the training phase for calibration is infrequent, the interference caused is negligible.

In the following sections we will discuss how the calibration of Tx-Rx RF can be incorporated in the beamformer design at both primary and secondary systems.

### C. Primary Beamformer Design with Channel Calibration

Consider at  $BS_1$  an SVD decomposition of the UL channel  $\mathbf{U}_{11} = \mathbf{Z} \mathbf{D} \mathbf{W}^H$  that  $BS_1$  estimates directly using pilots transmitted by  $MU_1$ . The primary DL channel can be written as function of the UL channel SVD decomposition using the calibration filters as:

$$\mathbf{D}_{11} = \mathbf{P}_{MU_1} \mathbf{U}_{11}^T \mathbf{P}_{BS_1} = \mathbf{P}_{MU_1} \mathbf{W}^* \mathbf{D} \mathbf{Z}^T \mathbf{P}_{BS_1} \quad (28)$$

(where  $*$  denotes complex conjugate). In order to diagonalize the DL channel  $BS_1$  designs its beamformer subspace as  $\mathbf{T}_{BS_1} = \mathbf{P}_{BS_1}^{-1} \mathbf{Z}^*$ , and hence the receiver filter at  $MU_1$  is given by:  $\mathbf{R}_{MU_1} = \mathbf{W}^T \mathbf{P}_{MU_1}^{-1}$ .

During UL transmission it is possible to design the transmit and receive filters using the UL channel as reference. In doing so, calibration filters do not appear in the expression and thus the transmit matrix at  $MU_1$  is  $\mathbf{T}_{MU_1} = \mathbf{W}$  and the receive filter at  $BS_1$  is:  $\mathbf{R}_{BS_1} = \mathbf{Z}^H$ .

#### D. Secondary Beamformer Design with Channel Calibration

The signal received at the secondary BS due to primary and secondary transmissions is expressed as

$$\bar{\mathbf{y}}_2 = \mathbf{U}_{21} \mathbf{T}_{MU_1} \bar{\mathbf{s}}_1 + \mathbf{U}_{22} \mathbf{T}_{MU_2} \bar{\mathbf{s}}_2 + \bar{\mathbf{n}}_2 \quad (29)$$

Knowing  $\mathbf{U}_{22} \mathbf{T}_{MU_2}$ , estimated through  $MU_2$  beamformed pilots,  $BS_2$  can determine the Tx subspace  $\mathbf{U}_{21} \mathbf{W}$  from  $MU_1$  blindly, using second order statistics.

Now let us consider the signal at  $MU_1$ , after the Rx filter, which is given by

$$\mathbf{r}_1 = \underbrace{\mathbf{R}_{MU_1} \mathbf{D}_{11} \mathbf{T}_{BS_1} \mathbf{s}_1}_{\mathbf{r}_{1,s}} + \underbrace{\mathbf{R}_{MU_1} \mathbf{D}_{12} \mathbf{T}_{BS_2} \mathbf{s}_2}_{\mathbf{r}_{1,int}} + \mathbf{n}_1 \quad (30)$$

where  $\mathbf{r}_{1,s}$  represents the useful signal part and  $\mathbf{r}_{1,int}$  contains the interference term. The objective of the secondary user is to transmit without causing any interference to the primary system. So  $BS_2$  must design its beamformer subspace such that  $\mathbf{r}_{1,int} = \mathbf{0}$ . Expressing the DL channel  $\mathbf{D}_{12}$  as function of the UL channel and the calibration filters we can write

$$\mathbf{r}_{1,int} = \mathbf{R}_{MU_1} \mathbf{D}_{12} \mathbf{T}_{BS_2} \mathbf{s}_2 = \mathbf{W}^T \mathbf{U}_{21}^T \mathbf{P}_{BS_2} \mathbf{T}_{BS_2} \mathbf{s}_2. \quad (31)$$

Since  $BS_2$  knows its own calibration filter  $\mathbf{P}_{BS_2}$ , it is possible to parameterize the secondary BF as:  $\mathbf{T}_{BS_2} = \mathbf{P}_{BS_2}^{-1} \hat{\mathbf{T}}_{BS_2}$ . Finally the beamformer subspace  $\hat{\mathbf{T}}_{BS_2}$  can be found as:

$$\hat{\mathbf{T}}_{BS_2} \subseteq (\mathbf{U}_{21}^* \mathbf{W}^*)^\perp \quad (32)$$

So  $\hat{\mathbf{T}}_{BS_2}$  has to lie in the orthogonal complement of the column space of  $\mathbf{U}_{21}^* \mathbf{W}^*$ . A similar treatment applies to the design of the  $MU_2$  beamformer which is not discussed here.

It is important to remark that the secondary Tx can design the beamformer subspace using only its own calibration factor, obtained during the calibration phase only with its intended receiver. Then the UL channel and the receiver subspace at  $MU_1$  are estimated using second order statistics of the received signal. Calibration between non cooperative users is not required.

### VIII. EXTENSION TO MULTIPLE PRIMARY PAIRS

The system model described so far can be easily extended to the situation where a cognitive system wants to coexist with a set of  $K$  primary transmitter and receiver pairs, Fig. 5. This problem formulation depicts the scenario where a femto-cell is deployed at the cell edge of a macro cell, thus the femto communication suffers from the interference received from the surrounding macro-cells. The primary system can be interpreted as a  $K$ -user MIMO interference channel (IFC). To mitigate the interference that each macro user receives from the other macro transmissions we assume that an interference alignment transmission strategy is used at the level of the primary communication. This strategy has been shown to maximize the degrees of freedom for the  $K$ -user MIMO IFC [7]. In this section we generalize the system model assuming that the  $k$ -th primary BS is equipped with  $M_k$  antennas while the corresponding primary MU has  $N_k$  antennas and they want to exchange  $d_k$  streams. We denote the number of antennas at secondary BS and MU as  $M_s, N_s$  respectively. To simplify

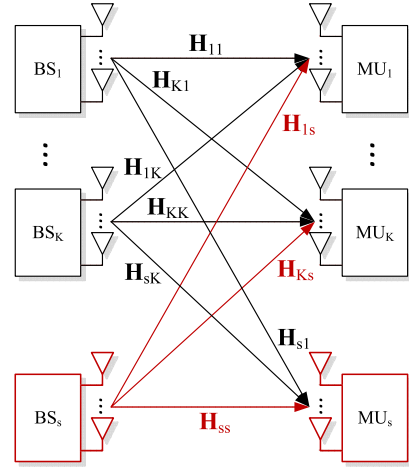


Fig. 5: Setting with multiple primary pairs

the notation we define with  $\mathbf{T}_{BS_l}$  and  $\mathbf{R}_{MU_l}$  the transmit and receive filter matrices at primary BS and MU number  $l$  and with  $\mathbf{T}_{BS_s}$  and  $\mathbf{R}_{MU_s}$  the same quantities at the secondary pair. According to IA, the transmit beamformers are designed such that the interference caused by all transmitters at each non-intended Rx lies in a common interference subspace. Then with a ZF receiver the interference can be completely suppressed. The interference alignment conditions can be simply described as:

$$\mathbf{R}_{MU_k} \mathbf{H}_{kl} \mathbf{T}_{BS_l} = \mathbf{0} \quad \forall l \neq k \quad (33)$$

$$\text{rank}(\mathbf{R}_{MU_k} \mathbf{H}_{kk} \mathbf{T}_{BS_k}) = d_k \quad \forall k \in \{1, 2, \dots, K\} \quad (34)$$

This last rank condition leads to the traditional single user MIMO constraint  $d_k \leq \min(M_k, N_k)$  for  $d_k$  streams to be able to pass over the  $k$ -th link. Since we suppose to use a TDD communication protocol thanks to channel reciprocity IA duality still holds, then :

$$\mathbf{R}_{BS_k} \bar{\mathbf{H}}_{kl} \mathbf{T}_{MU_l} = \mathbf{0} \quad \forall l \neq k \quad (35)$$

$$\text{rank}(\mathbf{R}_{BS_k} \bar{\mathbf{H}}_{kk} \mathbf{T}_{MU_k}) = d_k \quad \forall k \in \{1, 2, \dots, K\} \quad (36)$$

where  $\mathbf{R}_{BS_l} = \mathbf{T}_{BS_l}^H$ ,  $\mathbf{T}_{MU_l} = \mathbf{R}_{MU_l}^H$  are the UL-DL relationship between transmit and receive filters. From this conditions we can see that, as in the previous case with one single primary pair, the transmit signal subspace corresponds to the receive signal subspace.

The signal at the output of the  $k$ -th primary MU receiver can be written as:

$$\begin{aligned} \mathbf{r}_k &= \mathbf{R}_{MU_k} \mathbf{H}_{kk} \mathbf{T}_{BS_k} \mathbf{s}_k + \sum_{l \neq k} \mathbf{R}_{MU_k} \mathbf{H}_{kl} \mathbf{T}_{BS_l} \mathbf{s}_l \\ &+ \mathbf{R}_{MU_k} \mathbf{H}_{ks} \mathbf{T}_{BS_s} \mathbf{s}_s + \mathbf{R}_{MU_k} \mathbf{n}_k \\ &= \mathbf{R}_{MU_k} \mathbf{H}_{kk} \mathbf{T}_{BS_k} \mathbf{s}_k + \mathbf{R}_{MU_k} \mathbf{H}_{ks} \mathbf{T}_{BS_s} \mathbf{s}_s + \mathbf{n}'_k \end{aligned}$$

where we denoted with  $\mathbf{H}_{ks}$  the channel matrix between the secondary BS and the  $k$ -th primary MU and with  $\mathbf{n}'_k$  the noise at the output of the receive filter.

The stated objective of our investigation is to design the transmit filter at the secondary network such that the interference generated at all primary receivers is zero:  $\mathbf{R}_{MU_k} \mathbf{H}_{ks} \mathbf{T}_{BS_s} = \mathbf{0}$ ,  $\forall k$ . The received signal at cognitive BS in the UL

transmission phase can be written as:

$$\bar{\mathbf{y}}_s = \bar{\mathbf{H}}_{ss} \mathbf{T}_{MU_s} \bar{\mathbf{s}}_s + \sum_{l=1}^K \bar{\mathbf{H}}_{sl} \mathbf{T}_{MU_l} \bar{\mathbf{s}}_l + \bar{\mathbf{n}}_s \quad (37)$$

As shown in section V-B and V-C from the received signal at the secondary BS we can estimate the primary signal subspace generated at the secondary BS using semi-blind subspace estimation. Due to channel reciprocity and duality of IA the Tx and Rx signal subspace at each primary device are the same:

$$\begin{aligned} \bar{\mathbf{H}}_I &= [\bar{\mathbf{H}}_{s1} \mathbf{T}_{MU_1}, \dots, \bar{\mathbf{H}}_{sK} \mathbf{T}_{MU_K}] \\ &= \begin{bmatrix} \mathbf{R}_{MU_1} \mathbf{H}_{1s} \\ \vdots \\ \mathbf{R}_{MU_K} \mathbf{H}_{Ks} \end{bmatrix}^H = \mathbf{H}_I^H \end{aligned} \quad (38)$$

$\mathbf{H}_I$  represents the composite secondary to primary channel as seen at the joint outputs of the primary UEs. Then the signal subspace at all primary MUs in the DL communication, spanned by  $\mathbf{H}_I$ , can be estimated from the received signal at the secondary BS (37), using semi-blind techniques. The knowledge of the primary Tx subspace is enough to design the Tx filter at the secondary BS such that the interference that the secondary communication generates at each primary receiver lies in the noise subspace defined by IA. This implies:

$$\mathbf{H}_I \mathbf{T}_{BS_s} = \mathbf{0} \implies \mathbf{T}_{BS_s} \subseteq \text{span}(\mathbf{H}_I^H)^\perp$$

The last relationship says that the BF chosen by the secondary BS should be in the orthogonal complement of the subspace spanned by the matrix  $\mathbf{H}_I$ . In order to have a possibility to design the Tx filter each cognitive device should be equipped with a number of antenna greater than the total number of streams transmitted in the primary network:  $N_s, M_s > \sum_k d_k$ . The system model introduced so far can be read as an asymmetric  $K+1$  MIMO IFC since the primary set of users are not aware of the additional secondary pairs. In this setting the cognitive user has to align its interference to all the  $K$  primary receivers but at the same time it receives interference from all primary transmitters.

#### A. Transmit and receive filter design with calibration filters

In this section we show how to design the IA filters for the primary network and the transmit and receive filters at the cognitive users when calibration needs to be included in the filter design.

Initially we consider the IA design at the primary network.

Primary BSs and MUs, with the estimate of the UL channels, calculate the transmit and receive filters for the UL transmission using one of the iterative algorithms available in literature, for example [20] or [21]. Then the UL IA conditions are satisfied:  $\mathbf{R}_{BS_k} \mathbf{U}_{kl} \mathbf{T}_{MU_l} = \mathbf{0}$ . To apply the UL filters in the DL communication each terminal should pre-compensate for the UL-DL channel mismatch, as done in section VII:

$$\mathbf{T}_{BS_k} = \mathbf{P}_{BS_k}^{-1} \mathbf{R}_{BS_k}^T; \quad \mathbf{R}_{MU_l} = \mathbf{T}_{MU_l}^T \mathbf{P}_{MU_l}^{-1} \quad (39)$$

Applying the IA filter, found above, in the DL transmission

we get:

$$\begin{aligned} \mathbf{R}_{MU_l} \mathbf{D}_{lk} \mathbf{T}_{BS_k} &= \mathbf{R}_{MU_l} \mathbf{P}_{MU_l} \mathbf{U}_{kl}^T \mathbf{P}_{BS_k} \mathbf{T}_{BS_k} \\ &= \mathbf{T}_{MU_l}^T \mathbf{U}_{kl}^T \mathbf{R}_{BS_k}^T = (\mathbf{R}_{BS_k} \mathbf{U}_{kl} \mathbf{T}_{MU_l})^T = \mathbf{0} \end{aligned}$$

then the IA conditions also in the DL are satisfied. This concludes the filter design in the primary network. To understand how the secondary network can find the BF matrices for the concurrent transmission we first study the received signal at the secondary BS in the UL transmission:

$$\bar{\mathbf{y}}_s = \mathbf{U}_{ss} \mathbf{T}_{MU_s} \bar{\mathbf{s}}_s + \sum_{l=1}^K \mathbf{U}_{sl} \mathbf{T}_{MU_l} \bar{\mathbf{s}}_l + \bar{\mathbf{n}}_s \quad (40)$$

from the received signal above, the subspace spanned by the matrix

$\bar{\mathbf{H}}_I = [\mathbf{U}_{s1} \mathbf{T}_{MU_1}, \dots, \mathbf{U}_{sK} \mathbf{T}_{MU_K}]$  can be estimated. The objective of the secondary BF design is to cause zero interference at all the primary receivers, then the interference contribution at, for example, primary  $MU_k$  can be written as:

$$\mathbf{r}_{k,int} = \mathbf{R}_{MU_k} \mathbf{D}_{ks} \mathbf{T}_{BS_s} \mathbf{s}_2 = \mathbf{T}_{MU_k}^T \mathbf{U}_{sk}^T \mathbf{P}_{BS_s} \mathbf{T}_{BS_s} \mathbf{s}_2. \quad (41)$$

We parameterize the secondary BF as  $\mathbf{T}_{BS_s} = \mathbf{P}_{BS_s}^{-1} \hat{\mathbf{T}}_{BS_s}$ .  $BS_s$  knows the calibration filters obtained during the secondary calibration phase, then the secondary BF subspace can be designed such that  $(\mathbf{U}_{sk} \mathbf{T}_{MU_k})^T \mathbf{T}_{BS_s} = \mathbf{0}$ ,  $\forall k$ , then we choose:  $\mathbf{T}_{BS_s} \subseteq \text{span}(\mathbf{H}_I^H)^\perp$ . In this section we have shown that it is possible to extend the results obtained for spatial interweave in the simple scenario of one primary and secondary pair also to the case of multiple primary users when the transmit and receiver filters, of the primary system, are found according to IA. Here we have proved that also in this setting calibration between primary and secondary users is not required for the calculation of the secondary transmitters. A sub-product of this analysis is that also to find an IA solution, in the pure MIMO IFC, using UL-DL duality each device only needs its own calibration coefficient, so calibration between different pairs of primary users is also not required. This is another important remark because if we want to exploit channel reciprocity also for IA design we have shown that calibration is necessary and this must be done only between users belonging to the same pair.

## IX. NUMERICAL RESULTS

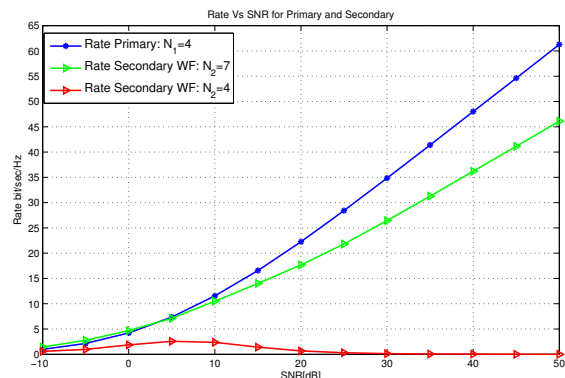


Fig. 6: Rate Comparisons

Fig. 6 depicts the rate curves for the primary and secondary links. We compare the performances of a cognitive radio system where the licensed users have  $N_1 = 4$  transmitting and receiving antennas. We report in the same figure the rate performances of a secondary system that have two possible antenna configurations:  $N_2 = 4$  and  $N_2 = 7$ . The primary communication is not affected by the opportunistic transmission thanks to the proper beamformer design of the secondary devices. On the other hand the rate of the secondary is very dependent on the number of cognitive antennas. The plot shows that if the secondary users have the same number of antennas of the primary the transmission takes place only in the low SNR region because the opportunistic users can only communicate using unused modes of the primary communication. When licensed users use all the possible modes there is no room for secondary transmission and hence the rate curve goes to zero. Different is the situation of an opportunistic user that is equipped with more antennas than the licensed one. In this case the cognitive user can transmit in all SNR regimes. In particular at high SNR the secondary system is able to sustain a significant rate with a maximum DoF equal to  $d_2 = N_1 - N_1$ .

Finally we want to study the effect of estimation error in the blind subspace estimation procedure at the secondary transmitters. In Fig. 7 we report the rate curves achieved by the primary user with and without error in the subspace estimation procedure. We can notice that if an error is present then the beamforming design at the secondary is not perfect and hence a residual interference affects the primary performances. In figure 7 we also compare the rate at the primary user when different number of samples  $T_E$  are used in the subspace estimation procedure. As we were expecting the longest the estimation period is the better the estimate, hence the rate loss decreases. We can also see that the rate loss due to signal subspace estimation affects only the SNR offset and not the multiplexing gain. This can be seen comparing the slopes of the three curves. In Fig. 8 we compare the theoretical

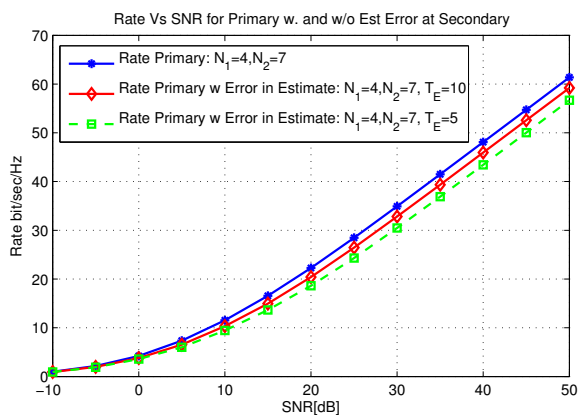


Fig. 7: Primary rate with estimation error at the secondary transmitter

rate loss,  $\Delta \mathcal{R}_{theo}$  upper bound found in section VI, with the experimental one  $\Delta \mathcal{R}_{exp}$ . As we can see the upper bound that we found becomes tighter as the number of samples  $T_E$  used for the estimation procedure increases. The rate loss is not constant over the SNR because it also depends on the number

of transmitted streams  $d_1$ , so it increases with the increase of  $d_1$ . When the primary transmitter uses all the available modes then  $\Delta \mathcal{R}$  remains constant.

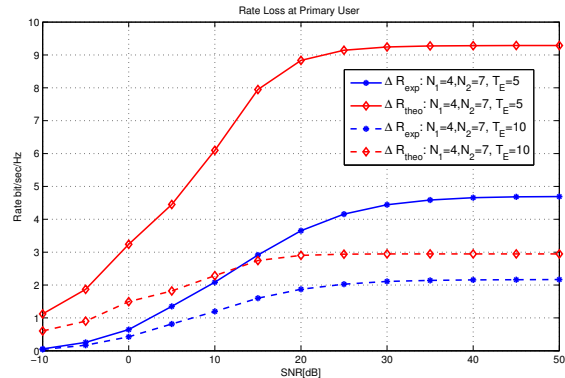


Fig. 8: Rate loss comparisons

#### A. RF perturbations and calibration

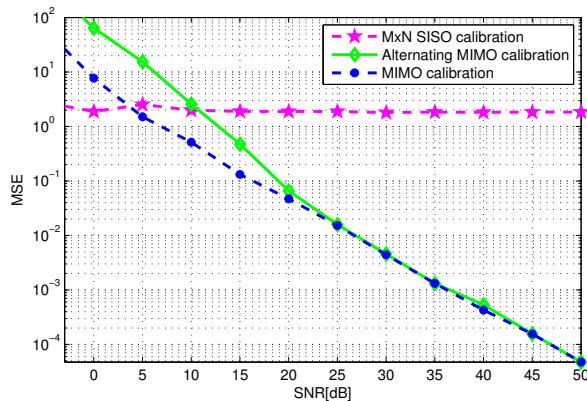


Fig. 9: Comparison of relative calibration techniques

In order to evaluate the performance of our calibration algorithm, we assume a  $2 \times 2$  MIMO system, where the calibration matrices are supposed to be frequency flat. The RF front-ends matrices  $\mathbf{R}_{M,B}, \mathbf{T}_{M,B}$  are randomly generated according to a normal distribution. The simulation are performed by the techniques described in [12], where we observe that it is possible to restore the reciprocity using  $K = 10$  channels estimations.

The Fig. 9 illustrates the performance of the selected calibration method (denoted "MIMO calibration"), in comparison to previous relative calibration techniques mentioned in [12] (denoted "alternating MIMO calibration" and " $M \times N$  SISO calibration"). It describes the reconstruction error of the downlink channel estimated after the calibration phase in one system. According to equation 19, the downlink channel is here reconstructed using the uplink channel estimation and the determined calibration matrices through algorithms. The simulation results show that our calibration technique outperforms the others, even with antenna coupling effects.

We propose now to simulate some RF perturbations related to the frequency flat RF front-ends in our cognitive radio

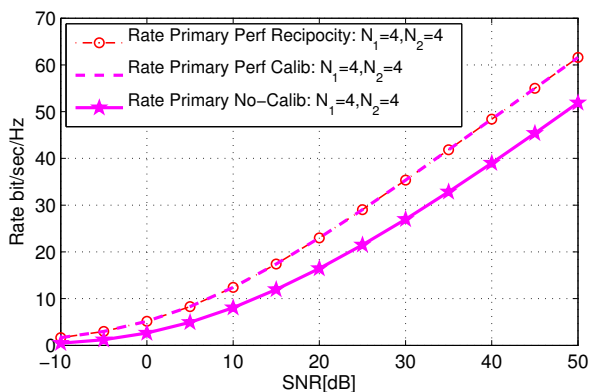


Fig. 10: Rate comparison assuming calibration

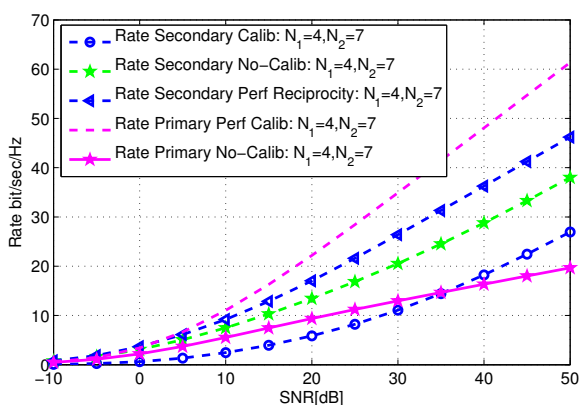


Fig. 11: Rate comparison assuming calibration

scenario. Here we will suppose a slight disturbance due to the RF parameters ( $\mathbf{R}_{MU}$ ,  $\mathbf{T}_{MU_1}$ ,  $\mathbf{R}_{BS}$ ,  $\mathbf{T}_{BS}$ ) and modeled using 4 complex matrices following a normal distribution  $CN(0, \mathbf{I})$ . Then, using the proposed calibration algorithm in Section VII-B, we evaluate the performances when the system is well calibrated ( $K = 10$  channel measurements over the time) and not calibrated which means that the transpose of the uplink channel is directly used to infer the downlink.

Assuming no Water Filling power allocation in the primary and the secondary system with  $N_1 = 4$  and  $N_2 = 4$ , we observe in Fig. 10 that due to the RF front-ends, the rate is reduced in comparison to the perfect reciprocity case described in Fig. 6, even when calibration parameters are well estimated ( $K = 10$ ). The Fig. 11 shows the rate with  $N_2 = 7$  antennas at the secondary system, it is also reduced in comparison to the perfect reciprocity rate (Fig. 6), due to the RF perturbations. Nevertheless when the primary and the secondary systems are not calibrated, the primary system is disturbed by the interference from the secondary. The effect of the RF front-ends without calibration in our simulations perturbs the primary and the secondary system, but the primary is mainly impacted because of the reduced number of antennas and the interference from the secondary system. Therefore, the implementation of the cognitive system described in this paper is conditioned by a proper calibration scheme, the next section addresses the practical consideration based on LTE.

## X. IMPLEMENTATION IN A LTE-TDD SYSTEM

The implementation in a realistic system is a big constraint for many theoretical approaches in CR. This section gives an overview of the main parts to address for an implementation using LTE-TDD specifications. The flexibility and the TDD mode offered by the LTE-TDD represent a promising choice for the implementation of CR methods.

### A. Calibration in LTE

To achieve our scenario, the first step consists in designing a reciprocity calibration protocol, thereby a new signaling is required in the LTE-TDD system in order to activate the calibration procedure. For the calibration at the base station side, we propose to modify the structure of downlink control information (DCI) send by the base station (eNodeB: eNB) in LTE, to start the feedback and the training phase allowing to exchange the  $K$  channel measurements from mobile users (user equipment: UE) to the eNB and vice versa. However, in the LTE implementation the impact of the selected feedback (analog or digital) should be further investigated. It is possible to quantize, encode and retransmit like an uplink data, the channel estimated by the UE.

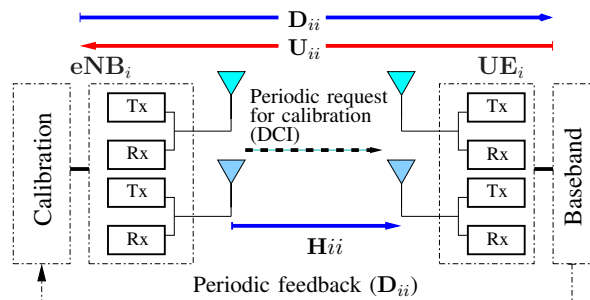


Fig. 12: Description of practical calibration process in LTE.

This data will be decode by the eNB, and finally, calibration parameters are determined in running algorithms at the eNB side using the receive downlink channel and the uplink channel estimation. This procedure is called periodically, according to the variation time of calibration matrices.

### B. LTE-TDD Frame Structure

Note that the beamformers described in the previous sections are designed during the coherence time of the channels. The 10ms periodic LTE-TDD frame is subdivided into 10 UL/DL subframes, each composed of two 0.5ms time slots (see Fig. 13). Consequently, the scenario defined in section V will be recast in LTE frame according to the subframes instead of the time slots. The channel coherence time in LTE is also a

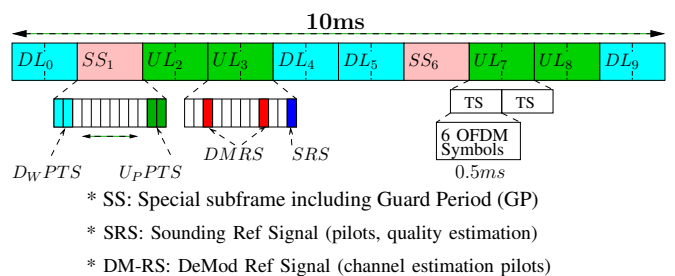


Fig. 13: LTE-TDD Frame configuration number 1

function of the UE speed according to the eNB like described in [22]. In this paper, the channel is assumed constant for at least 1 frame duration (10ms Fig. 13).

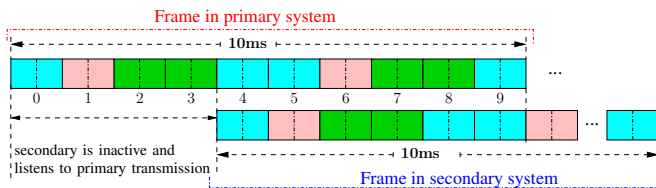


Fig. 14: LTE-TDD Frame specifications for primary and secondary system

Assuming that the secondary transmitter is aware of the primary TDD frame structure, it designs its frame structure in listening the pilots (RS, DRS, SRS) from primary transmitters. This scenario is conditioned by a good timing synchronization between the devices. Like illustrated in Fig. 14, in the primary subframe 0 and 2, the secondary system listens to the primary transmission and applies the processing described in first and second TDD slot (Section. V). After the subframe 3 all nodes transmit to their respective receivers and the processing in the third and fourth TDD slots is applied. The LTE implementation is finally conditioned by an efficient reciprocity calibration scheme, a good timing synchronization, a low speed of the UE, and a faster processing at eNB and UE. Some investigations base on these points are ongoing in order to achieve the LTE implementation.

## XI. CONCLUSIONS

We addressed beamformer design for a secondary communication system in a spatial interweave CR system. The practical problem of opportunistic CSI acquisition was addressed by exploiting primary signal statistics and reciprocity of the underlying TDD channel. Beamformer for secondary Tx is designed so that the secondary signal lies in the noise subspace of the primary signal. It must be noted that spatial interweaving of secondary's signal with the primary's relies on reciprocity of the TDD channel. Tx/Rx calibration is therefore mandatory. In this work we introduced a new calibration algorithm based on TLS MIMO technique. In situations where RF crosstalk effects can not be neglected, the proposed algorithm outperforms solutions introduced elsewhere e.g., [10]. The algorithm proposed is characterized by higher computational complexity. We believe this work paves the way for exploitation of reciprocity in MIMO systems and more generally in CR, thus reducing channel estimation and feedback overhead associated with CSI feedback.

The main contribution of this paper is the discovery that despite the requirement for channel reciprocity between noncooperation users, calibration between crosslinks is not required. We also extended the results to a CR system where a single secondary pair coexists with multiple primary user organized as an interference channel.

We finally proposed a possible practical implementation of the studied CR setting using a system based on the LTE frame structure.

## REFERENCES

- [1] S. Haykin, "Cognitive radio: brain-empowered wireless communications," *Selected Areas in Communications, IEEE Journal on*, vol. 23, no. 2, pp. 201–220, Feb. 2005.
- [2] A. Goldsmith, S. A. Jafar, I. Maric, and S. Srinivasa, "Breaking spectrum gridlock with cognitive radios: An information theoretic perspective," *Proceedings of the IEEE*, vol. 97, no. 5, pp. 894–914, May 2009.
- [3] Rui Zhang and Ying-Chang Liang, "Exploiting multi-antennas for opportunistic spectrum sharing in cognitive radio networks," *Selected Topics in Signal Processing, IEEE Journal of*, vol. 2, no. 1, pp. 88–102, feb. 2008.
- [4] G. Scutari, D. Palomar, and S. Barbarossa, "Cognitive MIMO radio," *Signal Processing Magazine, IEEE*, vol. 25, no. 6, pp. 46–59, november 2008.
- [5] Rui Zhang, Feifei Gao, and Ying-Chang Liang, "Cognitive beamforming made practical: Effective interference channel and learning-throughput tradeoff," *Communications, IEEE Transactions on*, vol. 58, no. 2, pp. 706–718, february 2010.
- [6] Feifei Gao, Rui Zhang, Ying-Chang Liang, and Xiaodong Wang, "Design of learning-based MIMO cognitive radio systems," *Vehicular Technology, IEEE Transactions on*, vol. 59, no. 4, pp. 1707–1720, may 2010.
- [7] V.R. Cadambe and S.A. Jafar, "Interference alignment and degrees of freedom of the K-user interference channel," *IEEE Trans. on Inform. Theory*, vol. 54, no. 8, pp. 3425–3441, Aug. 2008.
- [8] S.M. Perlaza, M. Debbah, S. Lasaulce, and J.-M. Chaufray, "Opportunistic interference alignment in MIMO interference channels," in *Personal, Indoor and Mobile Radio Communications. PIMRC 2008. IEEE 19th International Symposium on*, Sept. 2008, pp. 1–5.
- [9] F. Negro, I. Ghauri, and D.T.M. Slock, "Transmission techniques and channel estimation for spatial interweave TDD cognitive radio systems," in *Signals, Systems and Computers, 2009 Conference Record of the 43rd Asilomar Conference on*, nov. 2009, pp. 523–527.
- [10] M. Guillaud, D. T. M. Slock, and R. Knopp, "A practical method for wireless channel reciprocity exploitation through relative calibration," in *ISSPA 2005, 8th International Symposium on Signal Processing and Its Applications, Sydney, Australia*, Aug 2005.
- [11] J. Liu, G. Vandersteen, J. Craninckx, M. Libois, M. Wouters, F. Petre, and A. Barel, "A novel and low-cost analog front-end mismatch calibration scheme for MIMO-OFDM WLANs," in *Radio and Wireless Symposium, 2006 IEEE*, pp. 219–222.
- [12] B. Kouassi, I. Ghauri, B. Zayen, and L. Deneire, "On the performance of calibration techniques for cognitive radio systems," in *The 14th International Symposium WPMC'11*, Brest, France, oct 2011.
- [13] Wei Yu, Wonjong Rhee, S. Boyd, and J.M. Cioffi, "Iterative water-filling for gaussian vector multiple-access channels," *Information Theory, IEEE Transactions on*, vol. 50, no. 1, pp. 145–152, Jan. 2004.
- [14] M. Wax and T. Kailath, "Detection of signals by information theoretic criteria," *Acoustics, Speech and Signal Processing, IEEE Transactions on*, vol. 33, no. 2, pp. 387–392, apr 1985.
- [15] Huiyue Yi, "Nullspace-based secondary joint transceiver scheme for cognitive radio MIMO networks using second-order statistics," in *Communications (ICC), 2010 IEEE International Conference on*, may 2010, pp. 1–5.
- [16] F. Li, H. Liu, and R.J. Vaccaro, "Performance analysis for DOA estimation algorithms: unification, simplification, and observations," *Aerospace and Electronic Systems, IEEE Transactions on*, vol. 29, no. 4, pp. 1170–1184, oct 1993.
- [17] D. Maiwald and D. Kraus, "Calculation of moments of complex Wishart and complex inverse Wishart distributed matrices," *Radar, Sonar and Navigation, IEEE Proceedings*, vol. 147, no. 4, pp. 162–168, aug 2000.
- [18] Chris Bowick, John Blyler, and Cheryl Ajluni, *RF circuit design*, Newnes, 2008.
- [19] I. Markovsky and S. Van Huffel, "Overview of total least-squares methods," *Signal processing*, vol. 87, no. 10, pp. 2283–2302, 2007.
- [20] S.W. Peters and R.W. Heath, "Interference alignment via alternating minimization," in *Acoustics, Speech and Signal Processing, 2009. ICASSP 2009. IEEE International Conference on*, april 2009, pp. 2445–2448.
- [21] K. Gomadam, V.R. Cadambe, and S.A. Jafar, "Approaching the capacity of wireless networks through distributed interference alignment," in *Proc. IEEE Global Telecommunications Conf. (GLOBECOM)*, Dec 2008.
- [22] S. Sesia, I. Toufik, and M. Baker, *LTE, The UMTS Long Term Evolution: From Theory to Practice*, pp. 574–578, John Wiley & Sons Inc, 2009.