AirSync: Enabling Distributed Multiuser MIMO with Full Spatial Multiplexing

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Abstract—The enormous success of advanced wireless devices is pushing the demand for higher wireless data rates. Denser spectrum reuse through the deployment of more Access Points (APs) per square mile has the potential to successfully meet the increasing demand for more bandwidth. In principle, distributed multiuser MIMO (MU-MIMO) provides the best approach to infrastructure density increase, since several access points are connected to a central server and operate as a large distributed multi-antenna access point. This ensures that all transmitted signal power serves the purpose of data transmission, rather than creating interference. In practice, however, a number of implementation difficulties must be addressed, the most significant of which is aligning the phases of all jointly coordinated APs.

In this paper we propose AirSync, a novel scheme which provides timing and phase synchronization accurate enough to enable distributed MU-MIMO. AirSync detects the slot boundary such that all APs are time-synchronous within a cyclic prefix (CP) of the OFDM modulation, and predicts the instantaneous carrier phase correction along the transmit slot such that all transmitters maintain their coherence, which is necessary for multiuser beamforming. We have implemented AirSync as a digital circuit in the FPGA of the WARP radio platform. Our experimental testbed, comprised of four APs and four clients, shows that AirSync is able to achieve timing synchronization within the OFDM CP and carrier phase coherence (after the correction) within a few degrees. For the purpose of demonstration, we have implemented two MU-MIMO precoding schemes, Zero-Forcing Beamforming (ZFBF) and Tomlinson-Harashima Precoding (THP). In both cases our system approaches the theoretical optimal multiplexing gains. We also discuss aspects related to the MAC and multiuser scheduling design, in relation to the distributed MU-MIMO architecture. To the best of our knowledge, AirSync offers the first (to the best of our knowledge) real-world implementation of the full distributed MU-MIMO multiplexing gain, namely the ability to increase the number of active wireless clients per time-frequency slot linearly with the number of jointly coordinated APs, without reducing the per client rate.

I. INTRODUCTION

The enormous success of advanced wireless devices such as tablets and smartphones is pushing the demand for higher and higher wireless data rates and is causing significant stress to existing networks. While new standards (e.g., 802.11n/ac and 4G) are developed almost every couple of years, novel and more radical approaches to this problem are yet to be tested. The fundamental bottleneck is that wireless bandwidth is simply upper bounded by physical laws, in contrast to wired bandwidth, where putting new fiber on the ground has been the de-facto solution for decades. Advances in network protocols, modulation and coding schemes have managed steady but relatively modest spectral efficiency (bit/s/Hz) improvements. 4G-LTE, for instance, offers two to five times better spectral efficiency than 2.5G-EDGE. Denser spectrum reuse, i.e., placing more access points per square mile, has the potential to successfully meet the increasing demand for more wireless bandwidth [1]. On the other hand, in contrast to conventional cellular systems, a very dense infrastructure deployment cannot be carefully planned and managed for reasons pertaining to scale and cost. Therefore, the multiuser interference between different uncoordinated Access Points (APs) represents the main system bottleneck to achieve truly high spectral efficiency.

In theory, the ultimate answer to this problem is distributed multiuser MIMO (also known as “virtual MIMO”), where several (possibly multi-antenna) APs are connected to central server and operate as a large distributed multi-antenna base station. When using joint decoding in the uplink and joint precoding in the downlink, all transmitted signal power is useful, as opposed to conventional random access scenarios (e.g., carrier-sense) which waste power through interference. This approach is particularly suited to the case of an enterprise network (e.g., a WLAN covering a conference center, an airport terminal or a university), or to the case of clusters of closely spaced home networks connected to the Internet infrastructure through the same cable bundle.

Distributed multiuser MIMO (MU-MIMO) is regarded today mostly as a theoretical solution because of some serious implementation hurdles, such as providing accurate timing and carrier phase reference to all jointly coordinated APs and the ability to perform efficient joint precoding at a central server connected to the APs through a wired backhaul of limited capacity.

We consider a typical enterprise network as illustrated in Figure 1. Since in such networks the wired backhaul is fast enough to allow for efficient joint processing at the server (see Section IV), the major obstacle to achieving the full distributed MU-MIMO multiplexing gain is represented by the lack of synchronization between the jointly processed APs. The perceived difficulty of this task has led some researchers to believe that it is practically impossible to achieve full multiplexing gain in the context of distributed MU-MIMO. In this paper, we present the first (to the best of our knowledge) real-world testbed implementation which achieves the theoretical optimal gain by correcting, in real-time, the instantaneous phase offsets between geographically separated access points. We achieve this via AirSync.

In a nutshell, AirSync tracks the instantaneous phase of a
pilot signal broadcasted by a reference AP (the master AP), and predicts the phase correction across the duration of the MU-MIMO precoded slot in order to de-rotate the complex baseband signal samples. This enables APs to maintain phase coherence, which is necessary for MU-MIMO precoding. Notice that each AP transmits the precoded data signal and receives the master AP pilot signal simultaneously. This is accomplished by dedicating one antenna per AP to pilot reception, while the others are used for MU-MIMO transmission. We have implemented AirSync as a digital circuit in the FPGA of the WARP radio platform. We have also implemented Zero-Forcing Beamforming and Tomlinson-Harashima Precoding, two popular MU-MIMO precoding schemes widely investigated, from a theoretical viewpoint, in the literature. Using AirSync in a testbed consisting of eight WARP radios, four acting as access points connected to a central server and four acting as clients, we have shown that the theoretical optimal gain of multiuser MIMO is achievable in practice.

Optimal scheduling of downlink transmissions involves taking into account channel state information. Consequently, we investigate the protocol design for the MAC layer in distributed MU-MIMO systems, including channel-based client selection, downlink scheduling and adaptive coding modulation, either obtained through a family of Modulation and Coding Schemes (MCSs), as in IEEE 802.11n, or through the use of Incremental Redundancy rateless coding at the physical layer.

Several extensions and improvements of this basic layout are possible, and are discussed in the paper. For example, the master pilot signal range can be extended by regenerating and repeating the pilot signal at different frequencies. Also, we discuss the possibility of estimating the downlink channel matrix from training signals in the uplink and exploit the physical channel reciprocity of Time-Division Duplexing (TDD). In particular, this latter issue is discussed in detail in the recent work for a large centralized MU-MIMO system where all transmit antennas are clocked together (both timing and carrier frequency) and therefore are perfectly synchronous. Thanks to AirSync, the same approach for calibration can be used in a distributed MU-MIMO system, although in the present implementation we use a more conventional downlink training and feedback configuration.

While recently there have been a number of very interesting and important works in which some of the gains of multiuser MIMO have been shown (see Section for more details) none of these has managed to achieve both timing and carrier phase synchronization between remote transmitters precise enough to implement MU-MIMO with optimal multiplexing gain in the distributed transmitters scenario. It is also worthwhile to notice that while single-user MIMO and single AP (centralized) MU-MIMO can offer multiplexing gains for a given configuration of transmit and receive antennas, these can be further increased by extending the cooperation to the distributed case, provided that the transmitters (the APs) can be synchronized with sufficient accuracy. Therefore, our approach can potentially provide additional multiplexing gains on top of an existing configuration.

In summary, in this paper we make the following contributions:

- We introduce AirSync, the first (to the best of our knowledge) scheme which achieves phase synchronization in a distributed MU-MIMO setting.
- We implement AirSync as a digital circuit in the FPGA of the WARP platform.
- We showcase in a testbed consisting of eight WARP radios, the theoretical optimal spatial multiplexing gain is achievable in practice.
- We discuss practical implementation aspects of the MAC layer for distributed MU-MIMO, including scheduling, user selection, and adaptive MCS versus incremental redundancy with rateless coding.

We conclude this introduction by providing a brief outline for the rest of the paper. In Section I we discuss in detail related work both on the theoretical side and on the practical side. In Section II we discuss in detail related work both on the theoretical side and on the practical side. In Section III we discuss in detail related work both on the theoretical side and on the practical side. In Section IV we present an overview of the MU-MIMO precoding techniques implemented in our work. In Section V we present the hardware implementation of AirSync in detail. In Section VI we present a number of results obtained using our testbed implementation with four access points and four clients. We show results regarding the synchronization accuracy, the beamforming gain, the zero-forcing precision and the implementation of ZFBF and THP as the MU-MIMO precoding schemes to achieve optimal multiuser multiplexing gains. Section VII discusses the issues related to the MAC design for distributed MU-MIMO. Finally, Section VIII points out some developments under on-going investigation.

II. RELATED WORK

Theoretical Foundations. The pioneering papers by Foschini and Telatar have shown that adding multiple antennas both to the transmitter and to the receiver increases the capacity of a point-to-point communication channel. At practical medium-to-high Signal to Noise Ratios (SNRs), this gain manifests as a multiplicative factor equal to the rank of the matrix representing the transfer function between the transmit and the receive antennas. For sufficiently rich propagation scattering, with probability 1 this factor is equal
to min\{N_t, N_r\}, where \(N_t\) and \(N_r\) denote the number of transmit and receive antennas, respectively. The MIMO capacity gain can be interpreted as the implicit ability to create min\{\(N_t, N_r\)\} “parallel” non-interfering channels corresponding to the channel matrix eigenmodes, and it is referred to in the literature as multiplexing gain, or as the degrees of freedom of the channel. Subsequently, Caire and Shamai \([9]\) have shown that the MIMO broadcast channel, where the transmitter has \(N_t\) antennas and serves \(K\) clients with \(N_r\) antennas each, exhibits an analogous capacity factor increase of min\{\(N_t, K N_r\)\}, suggesting that a transmitter with multiple antennas could transmit simultaneously on the same frequency to independent users. Such multiuser communication has two additional requirements. First, precoding of the transmitted data is needed to prevent the different spatial streams from mutually interfering. Second, the transmitter requires accurate knowledge of the channel matrix (channel state information) in order to realize this precoding.

The idea of precoding has spurred research beyond the scope of this paper. Dirty Paper Coding (DPC) \([10]\) with a Gaussian coding ensemble achieves the capacity of the MIMO broadcast channel \([40]\), but is difficult to implement in practice. The well-known linear Zero-Forcing Beamforming (ZFBF) \([6]\) achieves the same high-SNR capacity factor increase, with some fixed gap from optimal that can be reduced when the number of clients is large and the transmitter can dynamically select the clients to be served depending on their channel state information \([23]\), [42]. Tomlinson-Harashima Precoding is another well-studied, but infrequently implemented technique, which efficiently approximates DPC at high SNRs \([41]\). A number of other precoding strategies (e.g., lattice reduction, regularized vector perturbation) have been studied and the interested reader is referred to \([34]\) and references therein. For the purposes of this paper ZFBF and THP will be the primary methods of interest because of their conceptual simplicity and good complexity/performance tradeoff.

**Practical Implementations.** A number of recent system implementations have made forays into the topics of multiuser MIMO transmission and distributed, slot aligned OFDM transmission. MU-MIMO ZFBF as a precoding scheme in a centralized setting have been examined in \([3]\), for a system consisting of a single AP with multiple antennas hosted on the same radio board. The use of interference alignment and cancellation as a precoding technique, which does not require slot synchronization or phase synchronization of the transmitters, has been illustrated in \([20]\). While this solution achieves a part of the potential spatial multiplexing gain, in order to realize the full spatial multiplexing requires tight phase synchronization between the jointly precoded transmitters \([24], [38]\).

In order to be able to adopt the classical discrete-time symbol-synchronous complex baseband equivalent channel models used in communication and information theory, the fundamental underlying assumption is that transmissions from different nodes align within the cyclic prefix of OFDM (referred to as “slot alignment” in the following). If this is not verified, then inter-block interference arises and the channel does not decompose any longer into a set of discrete-time parallel channels. Slot alignment was used in SourceSync \([29]\) in conjunction with space-time block coding in order to provide a diversity gain in a distributed MIMO downlink system. In Fine-Grained Channel Access \([35]\), a similar technique allows for multiple independent clients to share the frequency band in fine increments, without a need for guard bands, resulting in a flexible OFDMA (OFDM with orthogonal multiple access) uplink implementation.

**III. A Multiuser MIMO Primer**

We consider the OFDM signaling format, as in the last generation of WLANs and cellular systems (e.g., IEEE 802.11a/g/n and 4G-LTE \([26]\)). OFDM is a block precoding scheme. One OFDM symbol corresponds to \(N\) frequency-domain information-bearing symbols. By inverse FFT (IFFT), an OFDM symbol is converted into a block of \(N\) time-domain samples. This block is augmented by the cyclic prefix (CP), i.e., by repeating the \(L \leq N\) last samples at the beginning of the block. The OFDM symbol length \(N\) and the CP length \(L\) are design parameters. With CP length \(L\), any frequency selective channel with impulse response of length \(\ell \leq L + 1\) samples is turned into a cyclic convolution channel, such that, by applying an FFT at the receiver, it is exactly decomposed into a set of \(N\) parallel frequency-flat discrete-time channels in the frequency domain. Typical CP length is between 16 to 64 time-domain samples. For example, for a 20 MHz signal, as in IEEE 802.11g, the time-domain sampling interval is 50 ns, so that a typical CP length ranges between 0.7 and 3.2 \(\mu\)s.

In a multiuser environment OFDM has also a significant side advantage: as long as the different users’ signals align in time with an offset not larger than \(L - \ell\), where \(L\) denotes the CP and \(\ell\) is the maximum length of any channel impulse response in the system, their symbols after OFDM demodulation remain perfectly aligned in time and frequency. In other words, the timing misalignment problem between user signals, which in single-carrier systems creates significant complications for joint processing of overlapping signals (e.g., multiuser detection \([19]\), successive interference cancellation \([37]\), Zig-Zag decoding \([19]\)), completely disappears in the case of OFDM, provided that all users achieve timing alignment within the CP.

In a point-to-point MIMO link with \(N_r\) receive and \(N_t\) transmit antennas, the time-domain channel is represented by an \(N_r \times N_t\) matrix of channel impulse responses. Thanks to OFDM, the channel in the frequency domain is described by a set of channel matrices of dimension \(N_r \times N_t\), one for each of the \(N\) OFDM subcarriers. An intuitive explanation of the MIMO multiplexing gain can be given as follows: in the high-SNR regime, the receiver observes \(N_r\) (noisy) equations with \(N_t\) unknown coded modulation symbols on each time-frequency dimension, each of which carries ~

\[\text{1A time-frequency dimension corresponds to one symbol in the frequency domain, spanning one OFDM subcarrier over one OFDM symbol duration, and spans (approximately) 1 s \times \text{Hz}.}\]
log(snr) + O(1) bits, where O(1) indicates constants that depend on the channel matrix coefficients but are independent of SNR. For sufficiently rich scattering, the rank of the channel matrix is equal to \( \min\{N_r, N_t\} \) with probability 1. Therefore, using appropriate coding in order to eliminate the effect of the noise, up to \( \min\{N_r, N_t\} \) symbols per channel time-frequency dimension can be recovered with arbitrarily high probability, thus yielding the high-SNR capacity scaling

\[
C(\text{snr}) = \min\{N_r, N_t\} \log \text{snr} + O(1) \text{ bits/s/Hz}.
\]

**Zero-Forcing Beamforming.** In contrast to point-to-point MIMO, in a MU-MIMO system with one \( M \)-antennas sender and \( K \) single antenna receivers, \(^\text{2}\) it is not generally possible to jointly decode all the receivers observations, since the receivers are spatially separated and not generally able to communicate with each other. In this case, joint precoding from the transmit antennas must be arranged in order to invert, in some sense, the channel matrix and control the multiuser interference. One of the techniques to achieve this is linear Zero-Forcing Beamforming (ZFBF).

In ZFBF, the transmitter multiplies the outgoing symbols by beamforming vectors such that the receivers see only their intended signals. For instance, let the received signal on a given OFDM subcarrier at user \( k \) be given by

\[
y_k = h_{k,1}x_1 + h_{k,2}x_2 + \cdots + h_{k,N_t}x_{N_t} + z_k
\]

where \( h_{k,j} \) is the channel coefficient from transmit antenna \( j \) to user \( k \) and \( z_k \) is additive white Gaussian noise. Then, the vector of all received signals can be written in matrix form as

\[
y = H^H x + z
\]

where \( H \) has dimension \( M \times K \). Assuming \( K \leq M \), we wish to find a matrix \( V \) such that \( H^H V \) is zero for all elements except the main diagonal, that is \( H^H V = \Lambda^{1/2} = \text{diag}(\sqrt{\lambda_1}, \ldots, \sqrt{\lambda_M}) \). Letting \( x = Vu \), where \( u \) is the vector of coded-modulation symbols to be transmitted to the clients, we have

\[
y = H^H Vu + z = \Lambda^{1/2} u + z,
\]

so that each receiver \( k \) sees the interference-free Gaussian channel \( y_k = \sqrt{\lambda_k} u_k + z_k \).

When \( H \) has rank \( K \) (which is true with probability 1 for sufficiently rich propagation scattering environments typical of WLANs and for \( K \leq M \)) a column-normalized version of the Moore-Penrose pseudo-inverse generally yields the ZFBF matrix. This takes on the form

\[
V = H(H^H H)^{-1} \Lambda^{1/2},
\]

where \( \Lambda \) is chosen in order to ensure that the norm of each column of \( V \) is equal to 1, thus setting the total transmit power equal to \( \text{tr}(\text{Cov}(V u)) = E[\|u\|^2] \), i.e., equal to the power of the data vector \( u \).

As far as the achievable rate is concerned, since ZFBF converts the MU-MIMO channel into a set of independent Gaussian channels for each user, subject to the sum-power constraint \( \mathbb{E}[\|u\|^2] \leq \text{snr} \), we have immediately that the maximum sum rate of ZFBF is given by

\[
P_{\text{sum}}^\text{zfbf}(\text{snr}) = \sum_{k=1}^K \log (1 + \lambda_k q_k),
\]

where \( q_k \) denotes the power of the \( k \)-th data symbol in \( u \). The above expression can be maximized over the power allocation \( \{q_k\} \), subject to the constraint \( \sum_{k=1}^K q_k \leq \text{snr} \), resulting in the classical water filling power allocation of parallel Gaussian channels \(^\text{11}\).

**Tomlinson-Harashima Precoding.** In Tomlinson-Harashima Precoding (THP), the mapping from the data symbol vector \( u \) to the transmitted symbol vector \( x \) is non-linear. Consider again the channel model (2). THP imposes a given precoding ordering, and it pre-cancels sequentially the interference of already precoded signals. Without loss of generality, consider the natural precoding ordering to be from 1 to \( K \). Let \( H = QR \) be the QR factorization of \( H \), such that \( R \) is \( K \times K \) upper triangular with real non-negative diagonal coefficients, and \( Q \) is \( M \times K \) full unitary, such that \( Q^H Q = I \). THP precoding is formed by the concatenation of a linear mapping, defined by the unitary matrix \( Q \), with a non-linear mapping that does the interference pre-cancellation. Let \( \hat{u} = \text{THP}(u) \) denote the non-linear mapping of the data vector \( u \) into an intermediate vector \( \hat{u} \), that will be defined later. The linear mapping component of THP is then given by

\[
x = Q \hat{u},
\]

where \( \text{Cov}(\hat{u}) = \Sigma = \text{diag}(q_1, \ldots, q_K) \) and, as before, \( q_k \) denotes the power allocated to the \( k \)-th data symbol. It follows that the channel reduces to

\[
y = H^H \hat{x} + z = R^H Q^H Q \hat{u} + z = L \hat{u} + \tilde{z},
\]

where \( L = R^H \) is lower triangular. The signal seen at client \( k \) receiver is given by

\[
y_k = [L]_{k,k} \hat{u}_k + \sum_{j < k} [L]_{k,j} \hat{u}_j + z_k.
\]

Next, we look at the non-linear mapping \( u \mapsto \hat{u} \). The goal is to pre-cancel the term indicated by “interference” in (7). Notice that this term depends only on symbols \( \hat{u}_j \) with \( j < k \). Therefore, the elements \( \hat{u}_1, \ldots, \hat{u}_K \) can be calculated sequentially. A simple pre-subtraction of the interference term at each step would increase the effective transmit power and would result in a suboptimal version of the linear ZFBF treated before.

The key idea of THP is to introduce a modulo operation that limits the transmit power of each precoded stream \( \hat{u}_k \). This is defined as follows. Assume that the data symbols \( u_k \) are points from a QAM constellation uniformly spaced in the squared region of the complex plane bounded by the interval \([-\tau/2, \tau/2]\) on both the real axis and the imaginary axis. Then, for a complex number \( s \), let \( s \) modulo \( \tau = \]

\(^\text{1}\) We assume single-antenna receivers for simplicity of exposition. The extension to \( 1 \leq N_r \leq M \) antenna receivers is immediate.
s − Q_\tau(s)$, where $Q_\tau(s)$ is the point $(n + jm)\tau$ with integers $n, m$ closest to $s$. In short, $Q_\tau(s)$ is the quantization of $s$ with respect to a square grid with minimum distance $\tau$ on the complex plane, and $[s]_{\text{mod} \ \tau}$ is the quantization error. We let

$$\hat{u}_k = \sqrt{q_k} u_k - \sum_{k < k'} \frac{|L_{k,k'}| q_k q_{k'}}{|L|_{k,k'} \sqrt{q_k}} \mod \tau.$$  (8)

In this way, the symbol $\hat{u}_k$ is necessarily bounded into the squared region of side $\tau \sqrt{q_k}$, and its variance (assuming a uniform distribution over the squared region, which is approximately true when we use a QAM constellation inscribed in the square) is given by $E[|\hat{u}_k|^2] = \tau^2 / 6 q_k$. Letting $\tau = \sqrt{6}$ we have that the precoded symbols have the desired power $q_k$.

Let’s focus now on receiver $k$ and see how the modulo precoding can be undone. The receiver scales the received symbol $\hat{y}_k$ by $|L|_{k,k} \sqrt{q_k}$ and applies again the same the modulo $\tau$ non-linear mapping. Simple algebra then shows that

$$\hat{y}_k = u_k + \frac{z_k}{|L|_{k,k} \sqrt{q_k}} \mod \tau.$$  (9)

It follows that the interference term is perfectly removed, but we have introduced a distortion in the noise term. Namely, while $u_k$ is unchanged by the modulo operation, since by construction it is a point inside the square, the noise term $\frac{z_k}{|L|_{k,k} \sqrt{q_k}}$ is “folded” by the modulo operation, i.e., the tails of the Gaussian noise distribution are folded on the squared region. Noise folding is a well-known effect of THP [16].

As far as the achievable rate is concerned, it is possible to show (see [4], [14]) that this is given by

$$R_{\text{sum}}^{\text{THP}}(\text{snr}) = \sum_{k=1}^{K} \left[ \log(1 + |L|_{k,k}^2 q_k) - \log(\pi e / 6) \right]_+,$$

where $[.]_+$ indicates the positive part. Again, this sum rate can be optimized with respect to the power allocation $\{q_k\}$, subject to the sum power constraint $\sum_{k=1}^{K} q_k \leq \text{snr}$. The rate penalty term $\log(\pi e / 6)$ is the shaping loss, due to the fact that THP produces a signal which is uniformly distributed in the square region (therefore, a codeword of $n$ signal components is uniformly distributed in an $n$-dimensional complex hypercube).

IV. SYNCHRONIZATION IN DISTRIBUTED MIMO SYSTEMS

In a distributed MU-MIMO setting, timing and carrier phase synchronization across the jointly precoded APs are needed in order for ZFBF and THP precoding to work. As discussed above, timing synchronization requires only that all nodes align their slots within the length of the OFDM CP. This is relatively easy to achieve, and it has already been implemented in software radio testbeds as in [29], [35]. Carrier phase synchronization, however, is much more challenging. While a centralized MU-MIMO transmitter has a common clock source for all its RF chains [32], in a distributed setting each AP has an individual clock. The relative time-varying instantaneous phase offset between the different transmitters may cause a phase rotation of the transmitter signals across a downlink slot such that, even though at the beginning of the slot we have ideal precoding (e.g., ZFBF or THP), the interference nulling effect is completely destroyed towards the end of the slot.

It is important to remark here that, while synchronizing a receiver with a transmitter for the purpose of coherent detection is a well-known problem for which robust and efficient solutions exist and are currently implemented in any coherent digital receiver [28], here we are faced with a different and significantly harder problem, which consists of synchronizing the instantaneous carrier phase of different transmitters. This requires that APs must track an RF carrier reference and compensate for the relative (time-varying) phase rotation while they are transmitting the downlink slot. Simultaneous transmission of the data signal and reception of the carrier reference signal cannot be implemented by standard off-the-shelf terminals. Instead, we have devised a system architecture to accomplish this goal.

Why is distributed MU-MIMO challenging? For simplicity of exposition, consider a distributed MU-MIMO system with two clients and two access points, each one with a single antenna and using ZFBF (the following considerations apply immediately to more general scenarios). For nomadic users, as in typical WLAN setting, the physical propagation channel changes quite slowly with time, so that we may assume that the channel impulse response is locally time-invariant. In order to use ZFBF, the channel matrix coefficients at each OFDM subcarrier must be estimated and known to the transmitter central server. Various methods for learning the downlink channel matrix at the transmitter side have been proposed, including closed-loop feedback schemes (see [5] and the references therein) or open-loop schemes that exploit the uplink/downlink channel reciprocity of TDD systems [22]. For simplicity of exposition, we will assume here that the channel estimates correspond perfectly to the actual channel.

The central server computes the precoding matrix as seen in in Section III for each subcarrier $n = 1, \ldots, N$. Let

$$H(n) = \begin{bmatrix} H_{11}(n) & H_{12}(n) \\ H_{21}(n) & H_{22}(n) \end{bmatrix}$$

(11)

denote the $2 \times 2$ downlink channel matrix between the two clients and the two access point antennas on subcarrier $n$, and let $V(n)$ denote the corresponding precoding matrix such that $H^H(n) V(n) = \Lambda^{1/2}(n)$ is diagonal. If the timing and carrier phase reference remain unchanged from the slot over which the channel is estimated and the slot over which the precoded signal is transmitted, we obtain perfect zero-forcing of the multiuser interference.

Suppose now that the timing reference and carrier phase reference between the estimation and transmission slots of the two APs is not ideal. With perfect timing, the downlink channel from AP $i$ to client $j$ would have impulse response $h_{ij}(\tau)$. Instead, due to lack of synchronization, the impulse response is $h_{ij}(\tau - \tau_i - \delta_j) e^{j(\phi_i(t) + \theta_j(t))}$ where $\tau_i, \delta_j$ denote the timing misalignment of AP $i$ and client $j$, respectively, and $\phi_i(t), \theta_j(t)$ denote the instantaneous phase differences (with respect to $t$ to the nominal RF carrier reference) of AP $i$.
and client $j$, respectively. For simplicity, we assume here that the sampling clock at all nodes is precise enough such that we may assume that the sampling frequency is the same and does not change significantly in time over the duration of a slot. Hence, $\tau_i$ and $\delta_j$ can be considered as unknown constants. Furthermore, we assume that they are multiples of the sampling interval $T_s$ (i.e., the duration of the time-domain samples) otherwise the derivation is more complicated, involving the folded spectrum of the channel frequency response, but the end result is equivalent to what derived here. Instead, we model the instantaneous phases of the RF carrier oscillators as

$$
\phi_i(t) = \phi_i(0) + 2\pi\Delta_i t + w_i(t)
$$

$$
\theta_j(t) = \theta_j(0) + 2\pi\delta_j t + w_j(t)
$$

(12)

where $\phi_i(0), \theta_j(0)$ are unknown constants, $\Delta_i = (f_{c,i} - f_c)(N + L)T_s$ is the normalized frequency offset of node $i$ with respect to the nominal carrier frequency $f_c$, and $w_i(t)$ is a zero-mean stationary phase noise process, whose statistics depends on the hardware implementation. In the above expression, the time index $t$ ticks at the OFDM symbol rate, i.e., at intervals of duration $(N + L)T_s$.

From the well-known rules of linearity and time-shift of the discrete Fourier transform, we arrive at the expression for the effective channel matrix in (13). The diagonal matrix of phasors $\Phi(n; t)$ and $\Theta(n; t)$ depend, in general, on both the subcarrier and OFDM symbol indices $n$ and $t$. The multiplication of the nominal channel matrix $H(n)$ from the right (receiver side, according to the channel model (2)) poses no problems, since these phase shifts can be recovered individually by each client as in standard coherent communication (28). In contrast, the diagonal matrix $\Phi(n)$ multiplying from the left (transmitter side) poses a significant problem: since the server computes the MIMO precoding matrix $V(n)$ based on its estimate $\hat{H}(n)$, it follows that when applied to the effective channel $\hat{H}(n)$ in (13) the matrix multiplication $\hat{H}(n)V(n)$ is in general far from diagonal. To stress the importance of this aspect, we would like to make clear that the resulting signal mixing takes place over the actual transmission channel, making it impossible for the receivers to eliminate it.

**Why Synchronization Is Possible.** Any discussion on phase synchronization of distributed wireless transmitters must necessarily start with the mechanisms through which phase errors occur. Digital wireless transmission systems are constructed using a number of clock sources, among which the two most important ones are the sampling clock and the carrier clock. In a typical system, signals are created in a digital form in baseband at a sampling rate on the order of tens of MHz, then passed through a digital-to-analog converter (DAC). Through the use of interpolators and filters, the DAC creates a smooth analog waveform signal which is then multiplied by a sinusoidal carrier produced by the carrier clock. The result is a passband signal at a frequency of a few GHz which is then sent over the antenna.

Wireless receivers, in turn, use a chain of signal multiplications and filters to create a baseband version of the passband signal received over the antenna. Some designs, such as the common superheterodyne receivers, use multiple high frequency clocks and convert a signal first to an intermediate frequency before bringing it back to baseband. Other designs simply use a carrier clock operating at the same nominal frequency as the carrier clock of the transmitter and perform the passage from passband to baseband in a single step. We will be focusing on such designs in the ensuing discussion. After baseband conversion, the signal is sampled and the resulting digital waveform is decoded.

There are four clocks in the signal path: the transmitter’s sampling clock and RF carrier clock and the receiver’s RF carrier clock and sampling clock. All four clocks manifest phase “drift” (i.e., a linear time-varying term) and “jitter” (i.e., a random fluctuation term). We have assumed that the sampling clocks have no significant drift and jitter, and the only effect of timing misalignment (within the length of a CP) is captured by the constants $\tau_i$ and $\delta_j$ in (13). In contrast, the carrier clocks are affected both by drift and jitter (see (12)). Furthermore, the phase noise term $w_i(t)$ may have some slow dynamics that can be linearized locally, over the duration of a slot, and add up to the linear phase term, such that the slope of the phase drift is constant over a single slot, but it is not constant over longer time intervals, in general.

We have verified experimentally the validity of our model, by letting a transmitter send several tone signals, i.e., simple unmodulated sine waves, corresponding to different subcarriers of the OFDM modulation, and using a receiver to sample, demodulate and extract the instantaneous phase trajectory of the received tones. In the absence of phase offset these signals would exhibit a constant phase when measured over a sequence of several OFDM symbols. Instead, the measured instantaneous phase is time-varying and closely approximate parallel straight lines, as shown in Figure 2. The common slope of these straight lines is given by the carrier frequency offset $\Delta_i$ between transmitter and receiver. The spacing between the lines is given by constant phase terms $\frac{2\pi}{N+L}\tau_i n$ for different subcarrier index $n$, and depends on the time misalignment $\tau_i$ between the AP and the nominal slot initial time. The small fluctuations around the linear behavior of the instantaneous phase is due to the phase noise, which is quite small for the WARP hardware used in our system, as it can be observed qualitatively from plots as in Figure 2.

It follows that by estimating the spacing between the phase trajectories (intercepts with the horizontal axis) and their common slope, we can track and predict across the slot the phase de-rotation coefficients to be applied at each AP in order.

\[
\hat{H}(n; t) = \begin{bmatrix}
H_{11}(n) & H_{12}(n) \\
H_{21}(n) & H_{22}(n)
\end{bmatrix}
\begin{bmatrix}
e^{j\left(\frac{2\pi}{N+L}\tau_1 n + \phi_1(t)\right)} & 0 \\
0 & e^{j\left(\frac{2\pi}{N+L}\tau_2 n + \phi_2(t)\right)}
\end{bmatrix}
\begin{bmatrix}
\Phi(n; t) \\
\Theta(n; t)
\end{bmatrix}
\]

(13)
to “undo” the effect of the matrix \( \Phi(n; t) \). Notice that the de-rotation factor must be predicted a few OFDM symbols ahead, in order to include the delay of the hardware implementation between when an OFDM symbol is produced by the baseband processor (FPGA) to when it is actually transmitted.

V. AIRSYNC

The fact that the common phase drift of all subcarriers can be predicted by observing only a few of them prompts the following approach to achieving phase synchronization between access points: a main access point (master) is chosen to transmit a reference signal consisting of several pilot tones placed outside the data transmission band, in a reserved portion of the available bandwidth. An initial channel probing header, transmitted by the master access point, is used by the other transmitters in order to get an initial phase estimate for each carrier. After this initial estimate is obtained, the phase estimates will be updated using the phase drift measured by tracking the pilot signals. The estimate is used to calculate the difference between the carrier phase of each secondary transmitter and the phase of the master transmitter. This difference depends on the timing offset between the starting points of their frames and the frequency offset between the carrier frequency of the master AP (denoted by \( f_{c,1} \)) and the carrier frequency of each secondary AP (denoted by \( f_{c,i} \), for \( i > 1 \)). After obtaining the channel estimate, the secondary transmitters are able to undo effect of the instantaneous phase difference by derotating the transmitted frequency-domain symbols by the phase difference term along the whole transmission slot, thus eliminating the presence of the time-varying diagonal matrix \( \Theta(n; t) \) in front of the estimated channel matrix and therefore achieving the desired MU-MIMO precoding along the whole transmission slot.

More specifically, at time \( t = 0 \), the \( n \)-th subcarrier signal generated by the master AP has the phase \( \frac{2\pi}{N_T} \tau_1 n + \phi_1(0) \), while the carrier generated by AP \( i \) has the phase \( \frac{2\pi}{N_T} - \tau_1 n + \phi_i(0) \). The phase of the instantaneous phase difference obtained from the master pilot tones is, ignoring the phase noise terms, \( \frac{2\pi}{N_T} - (\tau_1 - \tau_i)n + \phi_1(0) - \phi_i(0) + \angle H_i(n) \), where \( \angle H_i(n) \) is the phase of the channel coefficient between the master AP and AP \( i \). If this phase estimate is added to the phase of the generated \( n \)-th subcarrier at AP \( i \), the resulting phase becomes \( \frac{2\pi}{N_T} \tau_1 n + \phi_1(0) + \angle H_i(n) \), that is the phase of AP \( i \) is the phase of the master AP plus an offset \( \angle H_i(n) \). To keep this offset constant over the duration of a transmission slot, the estimate must be adjusted by adding, for all \( t \) ranging over the transmission slot, the linear relative phase drift term \( 2\pi(\Delta_1 - \Delta_i)t \). In this way, after the phase compensation, all APs transmit at the actual frequency \( f_{c,i} \) of the master AP.

The drift \( 2\pi(\Delta_1 - \Delta_i)t \) is estimated based on the out-of-band pilots using a sliding window smoothing filter over four samples to compute an updated value of the “slope” \( \Delta_1 - \Delta_i \). The secondary AP predicts, based on the current estimate, the instantaneous phase with a few OFDM symbols of look-ahead. The need for look-ahead prediction arises from the fact that the AP must align its phase to the reference at the moment of the actual transmission, not at the moment that the estimate has been recorded. Thus the look-ahead time of \( d \) OFDM symbols corresponds to the synchronization circuit delay. The prediction is obtained by simple linear extrapolation, by letting the correction term at time \( t + d \) be given by \( 2\pi(\Delta_1 - \Delta_i)(t + d) \), where \( \Delta_1 - \Delta_i \) is the estimated slope at time \( t \). The constant offset \( \angle H_i(n) \) becomes a part of the downlink channel estimates and poses no further problems with regard to synchronization both when using downlink and uplink channel estimation schemes.

In our current implementation, for simplicity, we obtain an individual phase estimate of the form \( \frac{2\pi}{N_T} (\tau_1 - \tau_i)n + \phi_1(0) - \phi_i(0) + \angle H_i(n) \) for every subcarrier and use it independently of the estimates for other subcarriers in correcting the subcarrier phase. The form of the phase estimate suggests that it is possible to obtain a better estimate by breaking the estimation process into two distinct parts: obtaining an initial, high quality estimate of the constant \( \angle H_i(n) \) during a system calibration step and then estimating just the two factors \( \tau_1 - \tau_i \) and \( \phi_1(0) - \phi_i(0) \) in subsequent packet transmissions. The constant estimate in this case is needed since undoing the angle \( \angle H_i(n) \) amounts to equalizing the channel between the master AP and the \( i \)-th AP. After equalizing the channel, the resulting phases can be unwrapped along the carrier index \( n \).

It results that, after compensating for the angle \( \angle H_i(n) \), the phase of the estimate is \( \frac{2\pi}{N_T} (\tau_1 - \tau_i)n + \phi_1(0) - \phi_i(0) \), linear in the carrier index plus a constant term. A linear MMSE fitting can be applied in order to find the two factors mentioned, which are in fact the slope of the line (the carrier phase with regard to the subcarrier index) and its intercept.

Software Radio Implementation. We have implemented AirSync as a digital circuit in the FPGA of the WARP radio platform [30]. The WARP radio is a modular software radio platform composed of a central motherboard hosting an FPGA and several daughterboards containing radio frequency (RF) front-ends. The entire timing of the platform is derived from only two reference oscillators, hosted on a separate clock board: a 20 MHz oscillator serving as a source for all sampling signals and a 40 MHz oscillator which feeds the carrier clock inputs of the transceivers present on the RF front-ends. The shared clocks assure that all signals sent and received using the different front-ends are phase synchronous. Phase synchronicity for all sent signals or for all received signals
is a common characteristic of MIMO systems. However, the fact that the design of the WARP ensures phase synchronicity among the sent and received signals, as opposed to using separate oscillators for modulation and demodulation, greatly simplifies the synchronization task. The system’s data bandwidth is 5 MHz. We place the synchronization tones outside the data bandwidth, at about 7.5 MHz above and below the carrier frequency.

The slave APs have to track the out of band pilots (i.e., receive these signals) and transmit the data signal at the same time, in an FDD manner. We have dedicated one antenna of each secondary AP to receiving and tracking the reference signal, while the other antennas are used for transmitting phase-synchronous signals. The system design must mitigate self-interference between the transmit and receive paths.

In FDD transmission schemes in which the front-ends sample the entire system bandwidth, the dynamic ranges of the ADC and DAC circuitry plays an important limiting role. As opposed to a complete full-duplex system, in which self-interference cancellation is the main challenge to be solved, in bandwidth sharing systems the main challenge is accommodating both the incoming signal, i.e., the signal from the master AP, and the secondary AP’s data signal within the limited dynamic range of the secondary’s receiver front-end. A second challenge is shaping this data signal in order to prevent any significant power leakage outside the data band, mitigating the need for large guard bands between the data and the pilots.

The dynamic range needed can be computed as follows: assume that the secondary AP’s signal and the master AP’s pilots are broadcasted at the same power level. If the secondary’s receiver antenna is $\alpha$ times closer to the secondary’s transmitter antenna than to the one of the master’s, assuming a free space propagation model in which the power decays as $\frac{1}{r^2}$, it results that the data signal is received at $10 \cdot \log_{10}(\alpha^2)$ dB above the pilot signals. For $\alpha$ in the 32 to 128 range, this amounts to 30 dB to 42 dB. For comparison, the WARP’s 14-bit ADCs offer a dynamic range of 84 dB.$^5$

For the second problem, the design of WiFi-NC [8], offers a clear indication of what can be achieved in a software radio using the same components as the WARP. To limit the size of required guard bands in a bandwidth-sharing system, in which different APs divide the data band into slices and can transmit in duplex over separate slices, the authors construct an OFDM transmitter with a sharp spectral footprint. By employing digital filters in the FPGA, they achieve a 60 dB power decay with guard bands that total 4% of the data bandwidth, as proved by spectrum analyzer plots. Their filter response time is well within the cyclic prefix. This approach allows for decreasing the over-the-channel power leakage into the pilot band through sender-side filtering. In our system, we achieve a similar effect by using the baseband sender filter present in the transmit signal path of the WARP’s transceiver. In general, self-interference can be avoided using a number of other techniques such as antenna placement [9], digital compensation [13], or simply relying on the OFDMA-like property of a symbol aligned system [35] and preventing the secondary APs from using the pilot subcarriers.

We have implemented a complete system-on-chip design in the FPGA, taking advantage of the presence of hard-coded ASIC cores such as a PowerPC processor, a memory controller capable of supporting transfers through direct memory access over wide data buses and a gigabit Ethernet controller. Atop this system-on-chip architecture we have ported the NetBSD operating system and created drivers for all the hardware components hosted on the platform, capable of setting all system and radio board configuration parameters. The operating system runs locally but mounts a remote root filesystem through NFS. In the same system-on-chip architecture we integrated a signal processing component created in Simulink which provides interfaces for fast direct memory access. This latter component is responsible for all the waveform processing and for the synthesis of a phase synchronous signal and interfaces directly with the digital ports of the radio front-ends. We interfaced the Ethernet controller and the signal processing component using an operating system kernel extension responsible for performing zero-copy, direct memory access data transfers between the two, with the purpose of passing back and forth waveform data at high rates between a host machine and the WARP platform. The large data rates needed (160 Mbps for a 5MHz wireless signal sampled at the 16 bit precision of the WARP DACs for both the real (I) and imaginary (Q) parts of the corresponding baseband signal) required optimizing the packet transfers into and out of the WARP. For example, consider the direct memory access ring associated with the receive end of the Ethernet controller on the board, which is shared between packets destined to the signal processing component and packets destined to the upper layers of the operating system stack. We do not release and reallocate the memory buffers occupied by packets destined to the signal processing component. Instead, we use a lazy garbage collection algorithm in order to reclaim these buffers when they are consumed in a timely manner or reallocate them at a later point if they are not consumed before the memory ring runs low on available memory buffers. The rationale for this particular optimization is that the overhead of managing

$^5$This requirement could be further relaxed through the use of an analog rejection filter over the data band, before sampling, during the tracking period, thus decreasing the needed dynamic range through receive-side filtering.
the virtual-memory based reallocation of memory buffers of tens of thousands of packets every second would bring the processor of the software radio platform to a halt.

All transmitting WARP radios are connected to a central processing server through individual Ethernet connections operating at gigabit speeds. Most of the signal synthesis for the packet transmission is done offline, using Matlab code. We produce precoded packets in the form of frequency domain soft symbols. However, the synchronization step and the subsequent signal generation is left to the FPGA. The server, a fast machine with 32 processor cores and 64GB of RAM, encodes the transmitted packets and streams the resulting waveforms to the radios. Figure 4 illustrates the process of creating a phase synchronous signal at the secondary AP.

Centralized joint encoding. By transmitting phase synchronous signals from multiple APs, we have created a virtual single MU-MIMO transmitter, for which standard MU-MIMO precoding strategies can be used. However, the use of distributed APs complicates the design of the transmitter system. In order to eliminate multiuser interference, the data streams to different clients must be jointly precoded, as we have seen in Section III. For systems with a very large number of jointly processed antennas and targeting mobile cellular communications (e.g., see [32]), the centralized computation of the precoding matrix, of the precoded based band signals, and distribution of these signals to all the antennas would require a large delay, which is incompatible with the short channel coherence time due to user mobility. In contrast, in our enterprise network or residential network scenario, the channel coherence time is much longer (typical users are nomadic, and move at most at walking speed). Therefore, computing the precoding matrix does not represent a significant problem, and it is in fact better to perform centralized precoding and distribution of the baseband precoded signals. For example, using the conjugate beamforming scheme of [32], it is possible to compute the precoded signals in a decentralized way, since each AP $i$ needs just to combine the clients’ data streams with the complex conjugates of its own estimated channel coefficients, i.e., with the elements of the $i$-th row of the channel matrix. In the notation of Section III this corresponds to letting $x = cHu$, for some power normalizing constant $c$, such that the precoded channel becomes $y = cH^iHu + z$. Unless $M \gg K$, the resulting matrix $H^iH$ is far from diagonal, and the system is interference limited, i.e., by increasing the transmit power, the system sum rate saturates to some constant value (the system multiplexing gain in this case is 1, corresponding to serving only one client on each time-frequency dimension, as in standard FDMA/TDMA). Hence, while conjugate beamforming is an attractive scheme for very large $M$, relatively high client mobility and limited power (as in a cellular system), it turns out that in the WLAN setting with not so large $M$, low client mobility and large operating SNR (due to communication range of at most a few tens of meters) this is not a competitive choice.

As a matter of fact, centralized ZFBBF or THP precoding is much better in our setting. It should also be noticed that by centralized precoding we need only to send the I and Q components of the frequency-domain OFDM baseband (precoded) symbols to the APs. This requires roughly $2b \times W$ bit/s, for signal bandwidth $W$ Hz and $b$ quantization bits per real sample. Instead, decentralized processing requires to send all client data streams to all APs. Assume for example that $K$ clients are receiving at 4 bit/s/Hz (corresponding to 20 Mbps over a $W = 5$ MHz bandwidth). This requires $20 \times K$ Mbps to be sent to all APs, while in the case of centralized processing, with $b = 16$ bits of quantization, we need only $32 \times 5$ Mbps. Here, for $K > 5$, centralized processing is convenient also in terms of the backhaul data rate. For sufficiently large $K$, centralized processing is eventually less demanding than decentralized processing in terms of the backhaul data rates.

Our central server has an individual gigabit Ethernet connection to each of the WARP radios serving as APs. We divide the downlink time into slots and in each slot schedule for transmission a number of packets destined to various clients, according to an algorithm that will be presented in Section VII. For each AP, the server computes the I and Q components of the precoded baseband frequency domain waveform to be transmitted in the next downlink slot. However, it does not perform any phase correction at this point. The only information used in the precoding is the data to be transmitted and the channel state information between APs and clients. The server transmits their corresponding waveforms to all secondary APs, and finishes by feeding the master AP, so that the master AP starts transmitting right away and the secondary AP can immediately synchronize and follow.

At the moment we obtain CSI using a downlink estimation procedure, similar to the one presented in 802.11ac. In a future refinement of our system, we would like to reduce the overhead of obtaining CSI by using an uplink estimation scheme that takes advantage of channel reciprocity, thereby reducing considerably the length of the channel estimation procedure.

VI. PERFORMANCE EVALUATION

Our system setup is presented in Figure 3. It consists of a primary transmitter, three secondary transmitters and four receivers. The main sender uses a single RF front-end configured...
in transmit mode, placing an 18 MHz shaping filter around the transmitted signal. The secondary senders use an RF front-end in receive mode and a second RF front-end in transmit mode, with a 12 MHz shaping filter. As mentioned previously, the pilots used in phase tracking are outside the secondary’s transmission band, therefore the secondary transmitter will not interfere with the pilot signals from the main transmitter. The series of experiments is intended to test the accuracy of the synchronization, the efficiency of channel separation and the extent to which we achieve the theoretical gains that multiuser MIMO promises in our setup.

A. Synchronization Accuracy

In this particular experiment we have placed two transmitters and two receivers at random locations. We placed a third RF front-end on the secondary sender and configured it in receive mode. The secondary transmitter samples its own synthesized signal over a wired feedback loop and compares it with the main transmitter’s signal. The synchronization circuit measures and records the phase differences between these two signals. Since we use the primary transmission as a reference, in this experiment we do not broadcast the signal synthesized by the secondary transmitter in order to protect the primary transmission from unintended interference.

We have modified the synchronization circuit to produce a signal that is not only phase synchronous with that of the primary transmitter but has the exact same phase when observed from the secondary transmitter. To achieve this, the circuit estimates the phase rotation that is induced between the DAC of the secondary transmitter and the ADC through which the synthesized signal is resampled. It then compensates for this rotation by subtracting this value from the initial phase estimate. It is worth noting that this rotation corresponds to the propagation delay through the feedback circuit and is constant for different packet transmissions, as determined through measurements. The result was a synthesized signal that closely follows the phase of the signal broadcast by the master transmitter, as illustrated in Figure 5. The figure illustrates the initial phase acquisition process, the initial phase estimation, the tracking and estimation of the phase drift, as well as the synthesis of the new signal. The phase discontinuities appearing in the main transmitter’s signal are due to the presence of the PN sequence along with a temporary disturbance needed in order to tune the feedback circuit.

Figure 6 illustrates the CDF of the synchronization error between the secondary transmitter and the primary transmitter.
The error is measured on a frame-to-frame basis using the feedback circuit. In decimal degree values, the standard deviation is 2.37 degrees. The 95th percentile of the synchronization error is at most 4.5 degrees.

The radios were placed in a typical office environment. We have measured the SNR value of the synchronization pilots in the signal received by the secondary transmitter to be around 28.5 dB above the noise floor. This is easily achievable between typically placed access points.

### B. Beamforming gain

Our second experiment was done using two transmitters and a receiver with the secondary transmitter broadcasting a secondary signal over the air. We measured the channel coefficients between the two transmitters and the receivers using standard downlink channel estimation techniques and arranged the amplitudes and the phases of the transmitted signals such that at one of the receivers the amplitudes of the two transmitted signals would be equal while the phases would align. The maximal theoretic power gain over transmitting the two signals independently is 3.01dB. We compared the average power of the individual transmissions from the two senders to the average power of a beamformed joint transmission. Our measurements show an average gain of 2.98 dB, which is consistent with the precision of the synchronization determined in the previous experiment.

This result shows that for all practical purposes we are able to achieve the full beamforming gain in our testbed.

### C. Zero-Forcing Accuracy

![Graph showing the percentage of experiments vs zero forcing leaked power.](image)

Fig. 7. The Power Leakage of Zero-Forcing. The leaked power is significantly smaller than the total transmitted power, transforming each receiver’s channel into a high SINR channel.

The following experiment measures the amount of power which is inadvertently leaked when using Zero-Forcing to non-targeted receivers due to synchronization errors. Again we have placed two transmitters and a receiver at random locations in our testbed. We have estimated the channel coefficients and arranged for two equal amplitude tones from the two transmitters to sum as closely as possible to zero. The residual power is the leaked power due to angle mismatching. Figure 7 illustrates the CDF of this residual power for different measurements. The average power leaked is -24.46 dB of the total transmitted power. This establishes that Zero-Forcing is capable of almost completely eliminating interference at non-targeted receiver locations.

### D. Zero-Forcing Beamforming Data Transmission

![Scattering plots for two independent data streams transmitted concurrently using ZFBF.](image)

Fig. 8. The scattering diagram for two independent data streams transmitted concurrently using ZFBF demonstrates that AirSync achieves complete separation of the user channels.

This experiment transmits data from two transmitters to two receivers using ZFBF. We have used symbols chosen independently from a QAM-16 constellation at similar power levels. The scattering plots in Figure 8 illustrate the received signals at the two receivers. From the figure it is clear that we have created two separate channels, achieving thus a multiplexing factor of two over point-to-point transmissions. The actual rates achieved will depend on the quality of the two channels.

We would like to compare the performance of the multiuser MIMO system to a current standard. In current enterprise WiFi networks transmissions within a small area occur from single access points to single clients and are separated in time using TDMA. We use the best achievable point-to-point rate as an upper limit for the rates that the TDMA approach can achieve and compare the rates achieved by our system.

The SINR values at the two receivers are 29 dB and 26 dB respectively. In the same experiment, we measured the best point-to-point link to have a 32 dB SNR value. Using Shannon’s formula, these values translate to maximally achievable rates of 9.96 bits/second/Hz (bps/Hz) for the point-to-point channel and 18.27 bps/Hz for the compound MIMO channel. Thus, when using ideal codes, we achieve a multiplexing rate gain of 1.83, which is close to the theoretical value of 2.

At all the mentioned SNR levels 802.11g (a point-to-point standard) uses the same 64-QAM modulation, resulting in a rate of 6 bps/Hz (ignoring the error correcting code overhead, which is identical for all three SNR levels). Thus, we can say that both of the channels obtained through zero-forcing support WiFi operation at the highest commonly used rates and therefore equivalent to independent WiFi channels. We conclude that, using practical modulations, the experimental multiplexing gain equals the theoretical value of 2.

### E. Tomlinson-Harashima precoding

The final experiment uses four transmitters and four receivers. We employ Tomlinson-Harashima precoding. The results are illustrated in Figure 9, which presents the four distinct wireless channels created for the four users. Thus, we...
have achieved a multiplexing factor of 4. As before, the actual rate gains will depend on the quality of the channels.

We measured the SINR values of the four channels to be 16.8 dB, 19.2 dB, 21.4 dB and 20.8 dB. The lower SINR values are caused by increased levels of power leakage due to the presence of more transmissions to other receivers (see Figure [7] for the distribution of leaked power from a single interfering transmission). Again, the Shannon rate formula predicts achievable channel rates of 5.6 bps/Hz, 6.4 bps/Hz, 7.11 bps/Hz and 6.91 bps/Hz. The sum rate is 26 bps/Hz. As mentioned before, the best point-to-point channel in our setup has a quality level of 32 dB, allowing for 9.96 bps/Hz. Therefore the rate gain is about 2.6 when using four degrees of freedom and ideal codes.

More practically, we can compare the performance of our system when employing an extended 16-QAM constellation on every channel with the performance of 802.11g using a typical modulation. At 32 dB SNR, 802.11g would use a 64-QAM constellation and achieve (ignoring the error correcting code overhead) a spectral efficiency of 6 bps/Hz. In the MIMO case, we can achieve a sum rate of 16 bps/Hz using four 16-QAM constellations, leading to a multiplexing gain of 2.66 under practical modulations, while the theoretical value is 4. In a commercial implementation, we expect the leakage to be further reduced and we expect to be able to come closer of a rate gain of 4. In general, nearing the theoretical rate gains through spatial multiplexing requires precise channel state information and tight synchronization, as evidenced by our experiments.

VII. MEDIUM ACCESS CONTROL

Given that we have achieved the necessary synchronization accuracy between access points and realized the full multiplexing gain, we turn to the large body of work on optimal scheduling for centralized multiuser MIMO systems (see for example [12], [23]). Inspired by this work, we propose a MAC layer that significantly departs from the classic networking layered architectural model and adopts a cross-layer “PHY/MAC” design strategy.

A. High level description

Time Division Duplexing. First, we consider the issue of allocating air time and frequency spectrum between the uplink and the downlink. We can choose between two natural strategies for separating the uplink from the downlink: time division duplex (TDD) and frequency division duplex (FDD). TDD has the following two advantages. First, with TDD one can exploit channel reciprocity and measure the uplink channel, using pilots from the users to infer the downlink channel.

In the case of FDD, an explicit closed-loop channel estimation (from the downlink pilots sent by the access points) and feedback (from the clients to the server) needs to be implemented, with a protocol overhead that increases linearly with the number of jointly precoded access point antennas [21]. Second, TDD is ideally suited for the transport of asymmetric traffic, as is typical in an enterprise WiFi scenario, whereas an FDD system provides less flexibility for managing different traffic patterns. Specifically, with TDD, the downlink channel estimation procedure and the downlink time reservation proposed in the 802.11ac standard [2] can be applied to our distributed MIMO system as well.

We shall consider the scheduling of users in the uplink and downlink periods separately. In the uplink, clients compete for bandwidth using regular CSMA/CA. Thus, in the rest of this section we focus on the downlink. We note here that in order for our system to be backward compatible with legacy 802.11 clients and access points, protection mechanisms and modes of operation have to be implemented. Such mechanisms are described in the 802.11n/ac standards [2] where, using RTS/CTS, CTS-to-self frames and legacy format preambles, nearby devices can sense that the channel is in use and avoid collisions.

Downlink scheduling. The central server keeps track of packet queue sizes and other readily available QoS information, e.g. the time since these queues have been served last. It then selects a subset of users to transmit to at each downlink time slot. In the following we discuss in detail how the server selects these users at each time slot when ZFBF is the precoding scheme of choice. A similar approach can be applied to THP precoding with minimal changes.

The user selection and power allocation problem for linear Zero-Forcing precoding has a rich literature (for example [12], [18], [42]). Conceptually, this optimization problem can be solved by exhaustively searching over all feasible subsets of users, optimizing a weighted rate function under some general power constraints. In practice, greedy algorithms have proven to provide excellent results at moderate complexity [12], [23].

We begin by evaluating the achievable sum-rate using such a greedy policy. Firstly, the use of coding rates equal to the
corresponding Gaussian channel capacity $\log(1 + \text{SINR})$ is overly idealized; by mapping the SINRs into a discrete set of modulation and coding schemes (MCSs), we can model a more realistic scenario. For the sake of simplicity we assume that we can choose the best scheme based on the received SINR. Table 1 provides one such mapping that corresponds to the 9 mandatory MCSs of 802.11ac [2], keeping in mind that mappings vary by vendor or may be dynamically chosen in practical scenarios. In Figure 10 the sum rates for the two schemes, greedy ZFBF with ideal rates (ZF-G) and the adaptive coding and modulation (ACM) scenario (ZF-ACM) described above are evaluated for multiple SNRs in the case of 10 clients and 4 total access points antennas. For purposes of reference, the optimal, capacity-achieving Dirty Paper Coding (DPC) [10] precoding technique is shown in the same plot.

The huge gap between ZF-ACM and the ideal ZF motivates us to turn our attention to more flexible ways of allocating the rates in the multiuser MIMO scenario. The current standard, 802.11n, offers many code combinations to fully utilize the capacity of the MIMO channel. Since a multiuser MIMO system serves multiple users in the same time slot, an even larger set of rates and codes would have to be supported for efficiently using capacity. An attractive and innovative approach is the use of rateless codes (e.g., Raptor codes [15], [33] and the recently proposed Spinal codes [27]) at the physical layer, in a so-called Incremental Redundancy (IR) configuration (see [7], [25], [31]), as already exemplified by Strider, to decrease the signaling and retransmission overhead.

In an ideal rateless coding adaptation scenario, we would achieve the coded modulation capacity of a fixed large QAM constellation. In Figure 10 the performance of greedy zero-forcing with such an ideal rateless code (ZF-IR) is also depicted for an ideal family of random rateless codes based on a 256-QAM constellation. It is immediately obvious that the gains of using this IR configuration are tremendous in comparison with classic ACM.

Our protocol design follows the lines of 802.11ac [2]: before a downlink transmission period the access points broadcast a request for a number of clients to estimate their channels based on a channel probing message broadcasted shortly after. The access points then transmit requests for feedback in succession to each targeted client and wait for the corresponding feedback. Once all the information has been collected, the downlink period can begin. We note that the use of a STBC for control frames can improve their robustness, given that from a client perspective the phases of the access points are essentially random during this phase.

The downlink data packet starts with a transmission from the main sender containing a pseudo-noise sequence used to achieve frame alignment by the transmitters and for block boundary detection by the receivers. The master access point then transmits the first set of channel estimation pilots which are used by the other access points through uplink pilots (based on TDD reciprocity) or at the receivers using a standard downlink estimation procedure as described in 802.11ac. The central server uses the estimates to select a set of clients for the following transmission slots, according to the scheduling algorithms introduced earlier.

The choice between uplink and downlink estimation has an important impact on the design of the synchronization system. When using downlink pilots, the system must guarantee that the effective channel matrix before the receiver, that is the product of the two right-most matrices in (13), is constant between packet transmissions. This calls for actively aligning the phases of the transmitters (the right-most matrix) for every packet. In contrast, when using uplink estimation, the phase shifts induced by frame misalignment on the uplink and the downlink path cancel each other, allowing the access points to skip the phase alignment step.

Our protocol design follows the lines of 802.11ac [2]: before a downlink transmission period the access points broadcast a request for a number of clients to estimate their channels based on a channel probing message broadcasted shortly after. The access points then transmit requests for feedback in succession to each targeted client and wait for the corresponding feedback. Once all the information has been collected, the downlink period can begin. We note that the use of a STBC for control frames can improve their robustness, given that from a client perspective the phases of the access points are essentially random during this phase.

The downlink data packet starts with a transmission from the main sender containing a pseudo-noise sequence used to achieve frame alignment by the transmitters and for block boundary detection by the receivers. The master access point then transmits the first set of channel estimation pilots which are used by the other access points to determine the initial phases of the subcarrier tones, as described in Section V. After this point, all access points take part in the downlink transmission.

We tested each component of the downlink and uplink protocol slots. However, since our radios do not switch from receive to transmit in a timely manner, we could not perform complete real-time MAC experiments.

**Overhead.** A note on the overhead of the above MAC is in order. The overhead of our MAC is not more than that
of 802.11n. The additional signaling overhead comes from requiring a few frames to predict the initial phase, and a few frames to dictate the MAC addresses of the nodes from which we wish to request channel state information for the next time slot. Even with very conservative estimates this will be less than a 20% increase in header time duration over that of a traditional 802.11 system. Note, however, that we get a bandwidth increase that grows almost linearly in the number of clients. This means that our overhead, normalized such that we consider the total control bits over the total data bits transmitted during a fixed airtime slot, is much less than in a traditional 802.11 system.

VIII. DISCUSSION AND FUTURE WORK

Our future work is concerned with improving the robustness of the parameter estimators used in the synchronization system while reducing the synchronization overhead. We plan to complete a MAC layer implementation which relies on uplink pilots estimates (based on channel reciprocity) for obtaining low-overhead estimates of the channel matrix. Another research topic is making Airsync scalable through semi-decentralized precoding and the use of a hierarchical structure. Finally, we would like to extend our system by implementing a joint PHY/MAC layer based on an actual family of rateless codes and an incremental redundancy-based MAC layer.

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