Abstract—We present an experimental broadband wireless 4x4 multiple-input multiple-output (MIMO) space-time-block-coded (STBC) OFDM system based on an extension of the IEEE 802.11a/g physical layer (PHY) to multiple-antenna scenarios. A general description of the hardware architecture of the demonstrator is shown. The FPGA and DSP-based algorithms implemented at the transmit and receive sides are defined and compared to traditional single-antenna designs. Real measurement results over different spatial diversity schemes are finally presented to illustrate the effective gains obtained from the use of multiple transmit and receive antennas in a multicarrier system.

I. INTRODUCTION

The growing demand for broadband wireless data communications has motivated many research efforts in the last decades. Spectral efficiency, robustness and implementation complexity are the most important issues that should be taken into account for the design of new physical layer technologies.

Orthogonal frequency division multiplexing (OFDM) systems have been shown to be robust to multipath fading channels. Moreover, they can be efficiently implemented via simple discrete Fourier transforms at the transmit and receive sides, which makes it a suitable choice for future high-data-rate wireless systems. Nevertheless, it is widely known that OFDM systems suffer from severe degradation in the presence of phase-noise and frequency offsets. On the other hand, the use of multiple transmit and receive antennas has been recently shown to be an efficient way to increase system reliability or capacity. However, the combined use of multicarrier modulations and spatial diversity schemes has a variety of challenges that still need to be validated experimentally [1].

In this paper, we present a real implementation of a 4x4 MIMO-OFDM demonstrator based on an extension of the IEEE 802.11a/g PHY layer to multi-antenna schemes. The presented work was developed within the framework of the Medea+ (A1111) project Multi-Antenna TRansceivers for QoS, Ubiquitous and Improved wireless Systems (MARQUIS), where the partners participating in this paper collaborated in the implementation of a common multi-antenna demonstrator. The results obtained in this project show the effective gains achieved through the use of multiple transmit and receive antennas in a multicarrier system.

In Figure 1, the architecture of the presented demonstrator is illustrated. As far as the design of the transmitter is concerned, the main activities developed by Agilent were focused on the design of an FPGA-based real-time multichannel baseband signal source generating four output streams according to the PHY parameters described in Table I. Part of the specifications are compliant to the IEEE 802.11a/g standard, whereas the modifications made for the multiple antenna case were focused on the cyclic delays applied to the preamble, the rotation of the pilots and the selection of different STBC schemes. A description of these modifications is exposed in Section II. At the RF level, a four-channel 802.11g-compliant RF front-end and a four-antenna uniform linear array (ULA) were developed by OMP. Both the baseband and the RF parameters of the transmitter are controlled from a dedicated control node through PCI and RS232 interfaces. The activities developed by CTTC were focused on the design of the receive part of the demonstrator. The hardware of the receiver is integrated in a VME backplane. As far as the RF section is concerned, a dual-band (2.4GHz and 5.2GHz) multi-antenna radio integrated in a single-slot VME board and a four-antenna ULA were implemented. At the output of the RF front-end, a multi-DSP based platform is used to implement all the required synchronization and decoding algorithms, which are detailed in Section III. The synchronization strategy for a single-antenna system was revisited and new algorithms were proposed and implemented to solve the synchronization problems caused by the presence of multiple elements at the transmit side. The receiver can be reconfigured through the network from the debug and control node. All the parameters estimated in the receiver, as well as the decoded data, are sent to the visualization and logging node, where the measurement results presented in Section IV are obtained.

II. TRANSMITTER

Figure 1 shows a high level overview of the current version of the transmitter. It features two Agilent N6030A dual-channel boards, a multichannel RF front-end and a four-antenna array. The baseband signal generation replaces the normal ARB waveform generation functionality present on the Virtex 2 Pro FPGA on each N6030A board. The data source consists of a flexible random bit generator. Before the mapping to in-phase and quadrature (I/Q) samples, the output bit stream is cut in frames and a customized signal field is inserted (see Figure 2). No coding or interleaving is applied. The
Fig. 1. Global Architecture of the Demonstrator

modulation mapper fully supports the modes defined in the 802.11a standard [2]. The transmitter supports multiple MIMO schemes, such as Alamouti and Tarokh [3][4], as well as some feed-through and hybrid modes. Pilot tones and preamble are inserted in the encoded signal in the frequency domain. To ease tracking of each space stream signal at the receiver, pilot tones are rotated following the same method as in the TGnSync proposal for the 802.11n standard [5]. The preamble design matches the 802.11a specification. The signal is translated into time domain by taking an Inverse Fast Fourier Transform (IFFT) of the signal. The mapping of subcarriers and pilots prior to the IFFT is compliant with the 802.11a standard. Before the insertion of the cyclic prefix, some parts of the time domain signal (short training sequence, long training sequence and signal field) are cyclically shifted, the amount of shift depending on the spatial stream (see Table I). The cyclic prefix has a fixed length of a fourth of an OFDM symbol duration. Each spatial stream is upconverted to an intermediate frequency (IF) of 140MHz and finally converted to the analog domain in the two high speed DACs present on each module. Each board being limited to two output channels, the space-time encoding process for three and four output streams is split over the two boards. To ease development, baseband blocks prior to the spatial encoding block are duplicated on each board. The different parameters, like for example the modulation or the spatial processing scheme, are real-time reconfigurable thanks to a flexible dataflow mechanism and the use of trigger signals. This reconfiguration mechanism enables the reuse of the same image file for each FPGA. A careful synchronization process between the boards is thus necessary to ensure that the signals at the outputs are well aligned. A precision of 500ps has been achieved.

The outputs of the DACs are externally connected to the IF inputs of the RF front-end through high-quality RF cables. The RF front-end has been designed for the demonstrator purpose. It consists of five modules embodied in a chassis. The first module contains a local oscillator common to the four transmitter RF chains (phase noise $\leq$-85dBm/Hz@10KHz), a 10MHz reference (OCXO) and a serial interface. Two 10MHz reference outputs and an external reference input are also available. Each transmitter chain has a superheterodyne architecture with an IF input of 140MHz. The operating frequency range goes from 2.4GHz to 2.4835GHz and the output power can be adjusted from -20dBm to 11dBm with a resolution of 1dB for an input level of -10dBm. The RF outputs are connected to a four-antenna ULA with adjustable distance between the antennas. They are designed to operate in the 2.4GHz ISM, 5.1GHz and 5.8GHz bands. The type of design approach is a dipole with integrated balun. The external dimensions (excluding connector) are 50×50×1.6mm.

To control and configure the transmitter, a Graphical User Interface (GUI) has been developed. Besides basic operations, like setting up the board and loading the FPGA, it can be used to automate tests by scripting the configuration of real-time parameters. Hence, it is possible without any extra computation or loading to send different frame formats with different rates, different spatial encoding schemes and even different cyclic delays one after each other. The GUI integrates also the control of the RF front-end. Each RF module parameter, like attenuation or frequency, can be accessed and tuned. Automated tests with a variation in time of the attenuation in each module can be generated.

III. RECEIVER

1) Hardware Architecture: the hardware architecture of the receiver is illustrated in Figure 1. Four key building blocks can be identified: the four-antenna ULA, the four-channel dual-band RF front-end, the dual-channel ADC modules and the multi-DSP and FPGA processing units. All the boards in the receiver cabinet, except the ADC modules, are connected together through a VME backplane. The VME bus is mainly used for power distribution, debugging and reconfigurability purposes. The DSP and FPGA processing units provide the interface for the remote configuration of the receiver. The ADC modules are attached to the DSP or FPGA motherboards.
through a high-performance mezzanine bus that delivers high-speed data transfers to the digital processing units. The ADC modules provide also a general purpose input/output interface that is externally connected to the RF front-end. This is the control bus used for the reconfiguration of the MIMO front-end from the digital section of the receiver. In the RF section, the four-channel dual-band RF front-end will downconvert signals from RF to IF. The multiband capabilities of the front-end are focused on two well-known frequency bands specified in the IEEE 802.11a/g standards: 5.2GHz and 2.4GHz. The IF output interfaces of the front-end are connected to the ADC modules through external connections. The front-end generates also a digitally adjustable sampling frequency output which is synthesized from the same reference as the internal local oscillator. This output can be externally split if more than one ADC module needs to be fed with an external sampling clock. The RF input interfaces are externally connected to a 4-antenna ULA through high-quality semi-rigid RF cables. The mechanical assembly of the antenna array allows the distance between elements to be adjusted to up to 2λ at 5.2GHz, and the antenna used is a standard monopole.

2) Algorithm Description: in the DSP section, two processing units, multi-DSP and multi-FPGA, are available for the implementation of complex signal processing algorithms. For the development of the presented demonstrator, all the algorithms were integrated in a Pentek 4292 quad-DSP VME Board. Each of the four TI TMS320C6203 fixed-point DSPs, operating at 300MHz, is in charge of implementing the synchronization tasks required for each of the four receive antennas. After parameter estimation and correction, a joint space-time decoding is implemented in a single DSP, as shown in Figure 3.

It is well-known that in SISO WLAN systems, the short preamble has a known periodicity that makes it unique from the synchronization point of view. Due to the cyclic delays applied at the transmitter and depending on the channel response, long training sequences (LTS) may also be endowed with the same periodic characteristics. This is the main source of false alarm in the start of frame (SoF) detection algorithm when a traditional autocorrelation approach is used [6]. The algorithm implemented in this receiver for the frame detection emulates a traditional received signal strength indicator (RSSI) detector in the digital domain. This approach has been shown to be robust to any number of transmit antennas. Once the SoF has been detected, the coarse carrier frequency offset (CFO) is estimated during the short training sequence (STS). The algorithm implemented is based on a classic autocorrelation of the received samples [6] and yields the Maximum Likelihood (ML) estimate of the frequency offset. In the fine CFO estimation, the autocorrelations are applied to the LTS, thereby giving a more accurate estimate.

The main goal of the coarse timing estimation algorithm is to detect the starting sample of the first LTS. Based on the a-priori knowledge of the training sequence, the algorithm correlates the received signal with a known part of the preamble. In order to find a global maximum, the received signal is cross-correlated with the cyclically-shifted sequences from all the active transmit antennas. The accuracy of the resulting timing estimation is of ±0.5 samples. Timing errors in the time domain are transformed into a linear phase increment after an FFT. The fine timing estimation is performed over the two LTS, where an average timing and constant phase error is obtained in the frequency domain. Sampling frequency errors may create significant timing offsets over long data frames. By considering that both the sampling and the local oscillator frequencies are generated from the same reference, it can be shown that the estimation of the CFO is proportional to the sampling frequency error. After the estimation, sampling and timing offset errors are compensated using a third-order polynomial interpolator over the oversampled-by-2 input buffer.

Channel estimation is performed after the estimation and compensation of all the synchronization parameters. In a \(N_t \times N_r\) MIMO-OFDM scenario, the receiver must determine the channel weights from the \(n_t\)th to the \(n_r\)th element for each of the \(K\) subcarriers. Let us define the symbols received at the \(n_r\)th element after the FFT as

\[
r_{nr}(n) = \Delta \cdot (I_{N_t} \otimes F_{K_{0}}^{R}) h_{nr}(n) + w_{nr}(n),
\]

where \(n\) represents the OFDM symbol index, \(\Delta = [\text{diag}(t_1) \ldots \text{diag}(t_{N_t})]\), \(t_{n_t}\) is a column vector with the \(K\) cyclically-shifted training symbols transmitted at the \(n_t\)th element, \(K_0\) is the maximum allowed length for any channel impulse response, \(h_{nr}(n) = [h_{t_{n_t}n_r}(n) \ldots h_{F_{K_{0}}^{R}t_{n_t}n_r}(n)]^T\), \(h_{n_tn_r}(n)\) is a \(K_0\)-length column vector representing the channel impulse response from the \(n_t\)th to the \(n_r\)th antenna, \(F_{K_{0}}^{K_{0}}\) is the \(K \times K_0\) Fourier matrix and \(w_{nr}(n)\) models additive white Gaussian noise (AWGN). Channel estimation is performed in the time domain. At the receive antenna, \(K - K_0\) valid observations corresponding to the non-zero subcarriers are available, and \(K_0N_t\) parameters are estimated using a least-squares approach [7]. Since the least-squares model is not dependent on the OFDM symbol index \(n\), this matrix is precomputed off-line in order to optimize the computational cost of the algorithm. The selection of the cyclic delays applied to the LTS is directly related to the maximum number of taps that are estimated per channel. In the presented solution, the duration of one OFDM symbol was divided by the maximum \(N_t\) (see Table I). Practical results show that, from the maximum theoretical number of taps
The performance improvement due to MIMO has been mostly contributed to a rich scattering environment. Transmit and receive antenna positions were fixed with no direct line-of-sight (NLOS) for any of the paths between them. The room was classical: desks, shelves and metallic cupboards of-sight (NLOS) for any of the paths between them. The room was classical: desks, shelves and metallic cupboards. In order to compare different MIMO schemes, it is important to have control over the propagation channel to have the same channel conditions for each scheme. The performance improvement due to MIMO has been mostly studied in a simulated fast-fading channel, where MIMO systems outperform SISO in the presence of fast channel variation in time [3][4]. Because of some practical limitations, controlling that type of channel in the field was very difficult and a fair comparison could not be done. Consequently, we chose for our tests a channel very close to a static channel, with as much correlation in time as possible. The tests were done in an office environment with no movement. The configuration of the room was classical: desks, shelves and metallic cupboards contributed to a rich scattering environment. Transmit and receive antenna positions were fixed with no direct line-of-sight (NLOS) for any of the paths between them. The antenna array setups at both transmitter and receiver were classical ULA with all antenna elements linearly polarized (see Figures 4(a) and 4(c)). The antennas were 8cm spaced which corresponds to 0.64λ. The distance between transmitter and receiver was about 8m. During the measurements, real-time monitoring of some key parameters of the transmission was possible thanks to the feedback provided by the receiver to the visualization node (see Figure 4(b)). The node consists of a PC running real-time visualization software connected via an Ethernet link to the receiver. The software plots parameters like channel estimation per path, constellation after STB decoding, bit error rate (BER) and estimations of CFO and SFO. Besides visual monitoring, a complete logging of the data was done.

2) MIMO channel: before discussing performance results, we first characterized the channel. To verify that the channel is indeed close to time invariant, it is important to check that we have a good channel correlation in time (large coherence time). Figure 5(a) shows the power delay profile (PDP) variation over time during a complete measurement. On the plot, the multipath profile of the channel can be observed. To determine the coherence time of the channel, correlation coefficients have been calculated over the length of one measurement. The result shows that the magnitude of the correlation stays very high (≥0.9) during a complete measurement. The coherence time is thus large enough to conclude that the channel is time invariant. As a consequence, performance curves should not reflect any improvement due to diversity gain.

3) Performance results: the performance results that are provided in this subsection have been measured with identical OFDM transmission parameters. The uncoded random bit stream is mapped using 16QAM constellation and the frame length is 25 OFDM symbols long. The BER measurements have been done on about 10^6 bits. Many measurements have been done, we display and comment here the most important of them. Figure 5(b) shows the BER as a function of the signal-to-noise ratio (SNR) for different receive schemes. No encoding was applied. The system was tested with one (SISO), two (SIMO2) and four (SIMO4) receive antennas. To get different values of SNR, the output power of the transmitter was changed, the channel remaining constant. It can be seen that a significant increase of performance is obtained when the number of receive antennas is increased. For the same BER, the SNR improvement is in the order of 3dB when doubling the number of antennas. This is the benefit of array gain. It can be noted that the slopes of the curves are almost equal due to a lack of diversity gain of the time-invariant channel. Another BER measurement has been done for each separate SISO link (Tx1 to Rx1 and Tx2 to Rx1) and the
same applying the Alamouti scheme (Tx1 + Tx2 to Rx1). It can be shown that the resulting curves are on top of each other. This is again due to the static channel: diversity gain is not present. Adding more transmit antennas does not give an improvement on the BER versus SNR plot. The gain obtained with multiple transmit antennas scheme can thus not be proven this way. For this reason, we defined a new metric for our measurements, namely the Total Transmitted Power (TTP). It is defined as the sum of the powers of the signals at each transmit antenna. Figure 5(c) shows the BER as a function of this new metric for the last case, SISO vs Alamouti (a similar graph comparing SIMO performance against SISO could be shown). It is now possible to compare the Alamouti scheme performance against that of the two SISO. The measurement shows that the best BER performance is achieved when using only SISO1 while the SISO2 seems more attenuated. It is also observed that Alamouti performance is close to the performance of the SISO1 and compared to the arithmetic mean of both SISO, we see a clear improvement. This last point can be interpreted as follows: we know that during the measurements, the environment was kept time invariant. If we would not have done so, the SISO performance would have been affected by several external factors, such as movements in the room or slight move of the antennas, and in average the performance would have been closer to the mean curve. On the other side, the Alamouti performance would theoretically have been much less affected and a constant performance would have been observed.

Some first conclusions can be drawn from our measurements. 1) Adding receive antennas increases the performance thanks to the array gain 2) Adding transmit antennas keeping the TTP constant increases robustness of the link by exploiting the spatial diversity with even no need of channel knowledge at the transmitter. This last observation is of importance if we consider streaming applications, for example, where the stability of QoS is a key factor. In such a scenario, adding multiple transmit antennas would contribute to maintain the QoS constant while adding multiple receive antennas would improve the quality of the link, enabling a greater range without disconnections.

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