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Research Paper

2X2 STBC MIMO OFDM RECEIVER DESIGN FOR WLAN WITH ESTIMATION OF DIFFERENT TRANSMIT CARRIER FREQUENCY OFFSETS

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The combination of Multiple-Input Multipleoutput (MIMO) and Orthogonal Frequency Division Multiplexing (OFDM) has been viewed as the trend of the new wireless LAN technology. In this paper, the design of an up-to-date 2 x 2 Space-Time Block Coded (STBC) MIMOOFDM baseband receiver is presented with reference to an IEEE 802.11 n proposal. To solve possible different carrier frequency offsets between the two transmit antennas due to the antenna resistance match problem, this work proposes a new carrier frequency offset tracking method at the receiver. The overall design is verified on the Xilinx FPGA with an implementation loss of about 1.5 dB.

Keywords: MIMO-OFDM, Wireless LAN, Space-Time Block Code (STBC), Carrier frequency offset, Transmit antenna, IEEE 802.11n

INTRODUCTION

Orthogonal Frequency Division Multiplexing (OFDM) (U.S Patent No.3, 1970; Weinstein and Ebert, 1971; van Nee and Prasad, 1999; and Terry and Heiskala, 2002) has become the most popular technique for Wireless Local Area Network (WLAN) and Metropolitan Area Network (MAN) applications in the past decade because of its advantage to deal with multipath channels. OFDM is a multi-carrier modulation method with which the available spectrum is occupied by many subcarriers. Each of subcarriers, which number is as many as the number of thousands, is modulated in parallel such that the size of the Fast Fourier Transform (FFT) becomes a critical factor in the implementation of an OFDM system. The OFDM system benefits from high spectral efficiency, resiliency to RF interference, and lower multipath distortion. It is especially useful in wireless dispersive channels.

It is well-known that the OFDM technology has been adopted for many standards, such as DVB, Hiperlan/2 (Esli *et al.*, 2004), IEEE 802.11a/g/n, IEEE 802.16, LTE, etc. The challenge of the next-generation wireless LAN

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system is to increase more throughput rate and better performance than that obtained with the present WLAN standards for video transmissions. The multiple-input multiple-output (MIMO) system is an attractive technique to achieve this goal compared to the Single-Input Single-Output (SISO) system, especially in the rich scattering channel environments (van Nee and Prasad, 1999; and Terry and Heiskala, 2002).

The combination of these two powerful techniques, MIMO and OFDM, is naturally the candidate for the next-generation WLANs, which can provide higher transmission rate and better transmission quality for broadband communications. The Space-Time Block Code (STBC) scheme proposed in Alamouti (1998) has been adopted to implement modern MIMO systems in order to achieve higher channel capacity at the same Signal to Noise Ratio (SNR) (Lambotharan and Toker, 2005). The advantage of the STBC is able to increase the system performance by encoding data before transmission on multiple transmit antennas and decoding data from the outputs of multiple receive antennas, without invoking complicated computation.

In this work, a MIMO-OFDM baseband receiver with a two transmit and two receive antennas system is developed based on the proposal contributed to IEEE 802.11n (Kose and Edwards, 2005; and Tsai *et al.*, 2010). In practical applications, different carrier frequency offsets may exist among different transmit antennas due to resistance mismatch (Schenk and van Zelst, 2003; and Kai, 2012). Hence, a new Carrier Frequency Offset (CFO) estimation method is proposed here. Main modules of the 2 x 2 MIMO OFDM receiver, including frame detection, symbol timing, CFO correction, and channel estimation, are verified through the Xilinx FPGA, and the resultant implementation loss is about 1.5 dB.

THE 2X2 STBC MIMO OFDM SYSTEM MODEL

System Model

Consider a 2 x 2 MIMO OFDM system with two transmit and two receive antennas as depicted in Figure 1, where *N* subcarriers are employed. A simple and effective encoding scheme for two transmit antennas can be obtained by the Alamouti's proposal Alamouti (1998). In the Alamouti STBC encoder, two encoded symbols are transmitted over two antennas in two symbol durations. For the first symbol index, the first antenna transmits symbol 1 S and the second antenna symbol S_2; while for the second symbol index, the first antenna transmits symbol $-S_2^*$ and the second antenna symbol S_1^*. To modulate the transmitted symbols, the Inverse FFT (IFFT) is used to produce the time-domain signals on the channels. To avoid the intersymbol interference (ISI) effect due to the channels, the transmitted signals are added by Cyclic Prefix (CP) at the IFFT output before they are put on the channels, where the maximum delay spread of the channel has to be less than the CP length.

To best accommodate the application of multiple antennas, the spatial distance between transmit antennas is assumed to be larger than half of the wavelength of the radio carrier, and by this condition, the channels induced by transmit and receive antennas can be regarded uncorrelated such that the maximum diversity gain can be expected. For a 2 x 2 MIMO system in the WLAN application, the channel responses are assumed to be almost time-invariant between two consecutive symbols. Denote by h_{ij} with i, j = 1 or 2 the channel impulse response observed from the i^{th} transmit antenna to the j^{th} receive antenna. Then, the time-domain value of h_{ij} can be expressed as [12]

$$h_{ij}(t) = \sum_{k=0}^{L-1} a_{ij}(k) \cdot ¢\left(t - ²_{ij}(k)\right) \quad ...(1)$$

where $a_{ij}(k)$ represents the kth path gain, $²_{ij}(k)$ is the time delay of the k^{th} path, and L is the number of channel paths. Note that $²_{ij}(L-1)$ is the maximum channel delay spread.

The following wireless indoor channel model with exponentially decaying power delay profile is used to establish the 2 x 2 channels:

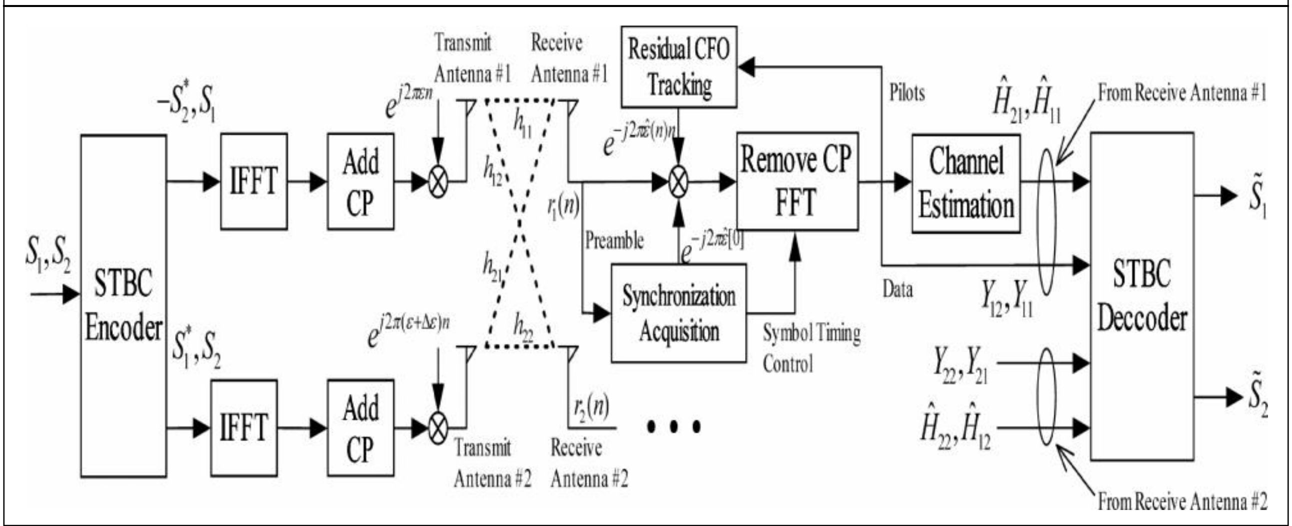
$$a_{ij}(k) = N\left(0, \pm_k^2/2\right) + j \cdot N\left(0, \pm_k^2/2\right) \quad ...(2)$$

$$\pm_k^2 = \pm_0^2 e^{-t_k/T_{RMS}} \quad ...(3)$$

where $\pm_k^2 = \pm_0^2 e^{-t_k/T_{RMS}}$ denotes the zero-mean Gaussian distribution with variance $\pm_k^2/2$, t_k is the delay time of the k^{th} path, T_{RMS} is the Root Mean Squares (RMS) time constant related to the channel power decaying speed, and $\pm_0^2$ is an initial value that satisfies normalized channel power.

Assume that the received signals experience different normalized CFOs: £ for transmit antenna #1, and £ + Δ£ for transmit antenna #2. Here, £ can be viewed as the Common CFO (CCFO). The different CFOs at the transmit side are modeled as shown in Figure 1. At the receiver, an estimated CCFO $\hat{£}$ is used to correct the frequency offset problem. For the WLAN application, a preamble with repeated short symbols is used to estimate the CFO. Besides, the FFT symbol timing is also obtained with the help of the preamble in order to pass the OFDM data payload through the FFT after correctly removing the CP. In addition to the short symbols for synchronization acquisition, a couple of long symbols are next to the short symbols for initial channel estimation. For the STBC decoder, the FFT outputs and the estimated channel values are required to solve

Figure 1: System Model for the 2 x 2 STBC MIMO OFDM Transceiver



the transmitted symbols S_1 and S_2 based on two consecutively encoded OFDM symbols. Moreover, the residual CFO due to inaccurate estimation obtained by the preamble will lead to the accumulation of phase offset at the FFT output such that serious degradation to OFDM demodulation can be induced.

Hence, residual CFO tracking with the help of pilots is required to remain correct constellation of output symbols.

The Alamouti STBC Scheme

Although OFDM systems have good ability of combating multipath channels in wireless communication applications, the performance is restricted by deeply faded subcarriers. Although typical diversity techniques can improve the channel gain, the receiver requires specific schemes for diversity combination. The Alamouti STBC provides a simple technique to solve 2 x 1 and 2 x 2 MIMO systems with enhanced diversity gain.

Suppose the RF impairments are removed and the system synchronization is perfect. Considering the 2 x 2 MIMO channels are time-invariant between two STBC symbols and denoting the DFT of h_{ij} by H_{ij}, the FFT outputs of the two consecutive STBC symbols for antenna 1 are

$$Y_{11} = H_{11}S_1 + H_{21}S_2 + W_{11} \quad ...(4)$$

$$Y_{12} = H_{11}S_2^* + H_{21}S_1^* + W_{12} \quad ...(5)$$

and the FFT outputs for antenna 2 are

$$Y_{21} = H_{12}S_1 + H_{22}S_2 + W_{21} \quad ...(6)$$

$$Y_{22} = H_{12}S_2^* + H_{22}S_1^* + W_{22} \quad ...(7)$$

where the notation * denotes complex conjugate, Y_{ij} denotes the FFT output signal for the j^{th} symbol on antenna i, and W_{ij} is independently complex white Gaussian noise. Once the channel response H_{ij} is known, the following linear combiner can be used to yield two separated signals $\tilde{S}_1$ and $\tilde{S}_2$ through the FFT output Y_{ij} at the STBC decoder:

$$\begin{aligned}\tilde{S}_1 &= H_{11}^*Y_{11} + H_{21}Y_{12}^* + H_{12}^*Y_{21} + H_{22}Y_{22}^* \\ &= \left(|H_{11}|^2 + |H_{21}|^2 + |H_{12}|^2 + |H_{22}|^2\right)S_1 + V_1 \end{aligned} \quad ...(8)$$

$$\begin{aligned}\tilde{S}_2 &= H_{21}^*Y_{11} + H_{11}Y_{12}^* + H_{22}^*Y_{21} + H_{12}Y_{22}^* \\ &= \left(|H_{21}|^2 + |H_{11}|^2 + |H_{22}|^2 + |H_{12}|^2\right)S_2 + V_2 \end{aligned} \quad ...(9)$$

where V_1 and V_2 are treated as new Gaussian noises, and

$$V_1 = H_{11}^*W_{11} + H_{21}W_{12}^* + H_{12}^*W_{21} + H_{22}W_{22}^* \quad ...(10)$$

$$V_2 = H_{21}^*W_{11} + H_{11}W_{12}^* + H_{22}^*W_{21} + H_{12}W_{22}^* \quad ...(11)$$

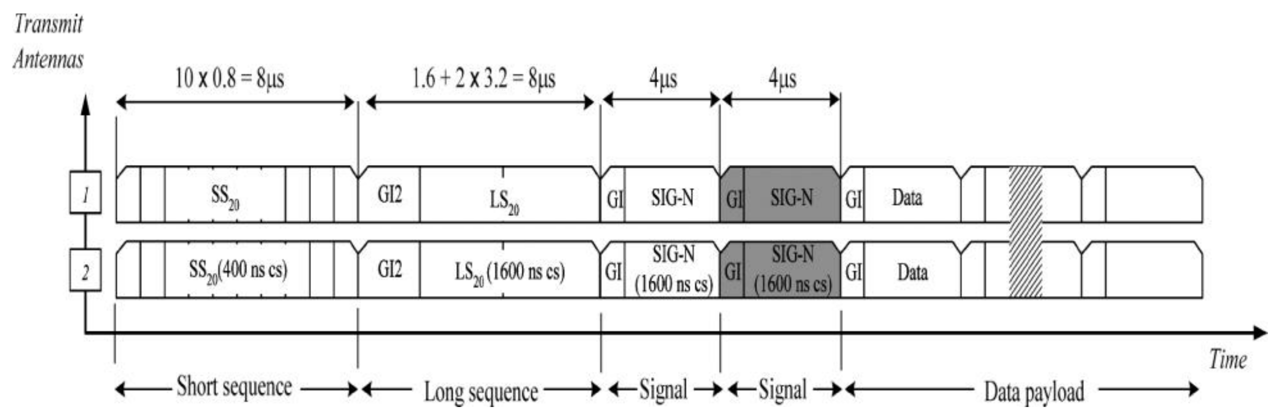
Here, $|H_{11}|^2 + |H_{12}|^2 + |H_{21}|^2 + |H_{12}|^2$ is the diversity gain. It is obvious from (8) and (9) that the combined signals $\tilde{S}_1$ and $\tilde{S}_2$ are obtained from the signals transmitted through the 2 x 2 channels. Suppose the channel responses are mutually uncorrelated, the deep fading occurring at certain subcarriers for one channel will be compensated by other channels with high likelihood. As the number of antennas increases, the diversity gain also increases such that the deep fading effect can be reduced. Finally, the estimate of S_1 and S_2 can be obtained by the Maximum Likelihood (ML) method through (8) and (9).

ALGORITHMS FOR THE MIMO OFDM RECEIVER

Preamble Format

For a WLAN application, the preamble is used for synchronization acquisition and channel

Figure 2: Preamble Structure for the 2 x 2 MIMO OFDM System



estimation. In this paper, the preamble depicted in Figure 2 is referenced to build the MIMO OFDM receiver and the FPGA prototyping design. The preamble format for two transmit antennas was originally presented in the WWiSE and TGn Sync's competitive proposals (Kose and Edwards, 2005; and Tsai *et al.*, 2010) used for the IEEE 802.11n standard. The durations of the short sequence and long sequence are 0.8 us and 3.2 us, respectively. The short sequence contains a sequence of ten repeated signals for Automatic Gain Control (AGC), frame detection, symbol timing, and coarse carrier frequency synchronization. The long sequence preceded by a guard interval of 1.6 us is used for channel estimation. Note that, the preamble in the second antenna is cyclic shift from the first one for cyclic delay diversity.

Synchronization Acquisition

Frame detection and symbol timing are two initial steps for the synchronization task. The frame detection is to detect the packet energy for a packet-based OFDM system. Since each of the repeated signals in the short sequence consists of 16 samples, the frame detecting signal, $C[m]$, can be calculated by

$$C[m] = \sum_{l=0}^{15} r[m-l]q^*[15-l] \quad ...(12)$$

where m is the sample index, $r[m]$ is the received signal, and $q[m]$ is the coefficient of the short preamble signal. The peak value of $C[m]$ occurs when $r[m]$ coincides to $q[m]$. As the short sequence format is detected, $C[n]$ will present repeated peaks by which the symbol timing can be determined as well.

Channel Estimation

From (8) and (9), the STBC data requires the estimated channel response to yield the transmitted data (Lambotharan and Toker, 2005). Since the long training sequence has the cyclic shift property, we can implement channel estimation based on the method proposed in Esli *et al.* (2004). Considering some null subcarriers for preventing interchannel interference, the long training sequence in the frequency domain is actually

$$P = [P_{-Nu}, P_{-Nu-1}, \ldots, P_{Nu-1}, P_{Nu}] \quad ...(13)$$

where the $2N_u + 1$ active subcarrier indices are located at $-N_u \le k \le N_u$. Denote the long training sequence in the time domain by

$$S = [S_0, S_1, \ldots, S_{N-2}, S_{N-1}] \quad ...(14)$$

According to the WWiSE draft, the long training sequence for the first antenna and the second antenna are

$$S_1 = S = [S_0, S_1, \ldots, S_{N-2}, S_{N-1}] \qquad \text{...(15)}$$

and

$$S_2 = CS^{N/2}(S) = [S_{N/2}, \ldots, S_{N-1}, S_0, \ldots, S_{N/2-1}] \qquad \text{...(16)}$$

respectively, where $CS^i(S)$ is the version of the cyclic shift by i of S. The FFT output of the long sequence for receive antenna j becomes

$$R_j[n,k] = \left(H_{1j}[n,k] + H_{2j}[n,k]e^{j®k}\right)\cdot P[k] + V_j[n,k] \qquad \text{...(17)}$$

where n is the OFDM symbol index and k is the subcarrier index. Note that notation j for exponential function $e^{j®k}$ is the imaginary unit, i.e., $j = \sqrt{-1}$. By dividing the subcarriers into even and odd parts, (17) can be re-written as

$$\begin{cases} R_j[n,2\bar{k}] = \left(H_{1j}[n,2\bar{k}] + H_{2j}[n,2\bar{k}]\right)\cdot P[2\bar{k}] + V_j[n,2\bar{k}] \\ R_j[n,2\bar{k}+1] = \left(H_{1j}[n,2\bar{k}+1] - H_{2j}[n,2\bar{k}+1]\right)\cdot P[2\bar{k}+1] + V_j[n,2\bar{k}+1] \end{cases} \qquad \text{...(18)}$$

where $\bar{k}$ is the decimate-by-2 index of k and $\bar{k} = \lfloor k/2 \rfloor$. Neighboring subcarriers can be regarded similar in channel frequency response because the interval between two adjacent subcarriers is quite small for WLAN applications. The complete frequency response can be estimated by the following interpolation technique based on the above assumption. The method has low complexity for hardware implementation.

After simple manipulation for (18) to yield the 2 x 2 channel responses, from antenna 1 we have

$$\hat{H}_{1j}[n,2\bar{k}] = \frac{R_j[n,2\bar{k}-1]}{4\cdot P[2\bar{k}-1]} + \frac{R_j[n,2\bar{k}]}{2\cdot P[2\bar{k}]} + \frac{R_j[n,2\bar{k}+1]}{4\cdot P[2\bar{k}+1]} \qquad \text{...(19a)}$$

$$\hat{H}_{1j}[n,2\bar{k}+1] = \frac{R_j[n,2\bar{k}]}{4\cdot P[2\bar{k}]} + \frac{R_j[n,2\bar{k}+1]}{2\cdot P[2\bar{k}+1]} + \frac{R_j[n,2\bar{k}+2]}{4\cdot P[2\bar{k}+2]} \qquad \text{...(19b)}$$

and from antenna 2 we have

$$\hat{H}_{2j}[n,2\bar{k}] = \frac{R_j[n,2\bar{k}-1]}{4\cdot P[2\bar{k}-1]} + \frac{R_j[n,2\bar{k}]}{2\cdot P[2\bar{k}]} + \frac{R_j[n,2\bar{k}+1]}{4\cdot P[2\bar{k}+1]} \qquad \text{...(20a)}$$

$$\hat{H}_{2j}[n,2\bar{k}+1] = \frac{R_j[n,2\bar{k}]}{4\cdot P[2\bar{k}]} + \frac{R_j[n,2\bar{k}+1]}{2\cdot P[2\bar{k}+1]} + \frac{R_j[n,2\bar{k}+2]}{4\cdot P[2\bar{k}+2]} \qquad \text{...(20b)}$$

By (19) and (20), the 2 x 2 channel frequency responses can be individually obtained from the DFT output of the received preamble long sequence that simultaneously mixes the transmitted signals over the two transmit antennas.

CFO Estimation and Tracking

For simplicity, assuming the CFO is different only between the transmitter and the receiver at first, the typical methods used in a SISO OFDM system can be used to estimate the CFO. Let D be the delay between two consecutive repeated symbols in the short preamble sequence. The initial CCFO £[0] can be estimated by [8] as follows:

$$£[0] = -\frac{1}{2®DT_s}\angle\Lambda \qquad \text{...(21)}$$

where T_s is the sample duration and Λ is obtained as

$$\Lambda = \sum_{j=1}^{2}\sum_{m=0}^{D-1} r_j[m]\, r_j^*[m+D] \qquad \text{...(22)}$$

where $r_j[m]$ is the received time-domain signals for the j^{th} antenna.

More accurate estimation of the CCFO can be performed by pilots in the OFDM data payload. Let N_p be the number of pilots used for CFO tracking. Supposing the intercarrier interference (ICI) effect caused by the residual CFO is small, we can write the output of the DFT for the j^{th} antenna due to the CCFO £ as (van Nee and Prasad, 1999).

$$Y_j[n, k_p] = R_j[n, k_p]e^{j2®£n} \quad ...(23)$$

where k_p is the pilot index and $R_j[n, k_p]$ denotes the ideal output. Then, the output of the next OFDM symbol at $n + 1$ is

$$Y_j[n+1, k_p] = R_j[n+1, k_p]e^{j2®£(n+1)} \quad ...(24)$$

Since the pilots and the estimated channel values can be referenced at this stage, $R_j[n, k_p]$ and $R_j[n+1, k_p]$ are known at the receiver. Based on the following equivalent,

$$Y_j^*[n, k_p]\,Y_p[n+1, k_p]\,R_j[n, k_p]\,R_j^*[n+1, k_p]$$

$$= \left|R_j[n, k_p]\right|^2 \left|R_j[n+1, k_p]\right|^2 e^{j2®£} \quad ...(25)$$

the CCFO can be calculated by

$$£[n] = \frac{1}{2®}\angle\left(\sum_{j=1}^{2}\sum_{k_p=1}^{N_p} Y_j^*[n, k_p]\,Y_j[n+1, k_p]\,R_j[n, k_p]\,R_j^*[n+1, k_p]\right) \quad ...(26)$$

The correction of the CCFO should be executed before the FFT by using a loop filter and a Numerical Control Oscillator (NCO) for improving the tracking capability. The NCO input signal will converge to zero as the residual CCFO is perfectly removed.

As considering a small phase offset $\Delta£(n)$ exists between two transmit antennas, where both antennas have a CCFO £ with respect to the receive antennas, the DFT output signal for the j^{th} receive antenna can be written as:

$$\begin{aligned} &Y_j[n, k_p] \\ &= S_1[n, k_p]H_{1j}[n, k_p]e^{j2®£n} + S_2[n, k_p]H_{2j}[n, k_p]e^{j2®(£+\Delta£)n} \\ &= \left(S_1[n, k_p]H_{1j}[n, k_p] + S_2[n, k_p]H_{2j}[n, k_p]e^{j2®\Delta£ n}\right)e^{j2®£n} \end{aligned} \quad ...(27)$$

where $S_1[n, k_p]$ and $S_2[n, k_p]$ represent the transmitted STBC pilot signals from antenna 1 and antenna 2, respectively. Denote by $\overline{Y}_j[n, k_p]$ the CCFO derotated output, and

$$\overline{Y}_j[n, k_p] = Y_j[n, k_p]e^{-j2®£n} \quad ...(28)$$

The following equation can be established:

$$e^{j2®\Delta£n} = \frac{\overline{Y}_j[n, k_p] - S_1[n, k_p]H_{1j}[n, k_p]}{S_2[n, k_p]H_{2j}[n, k_p]} \quad ...(29)$$

Using the STBC property, $S_1[n+1, k_p] = -S_2^*[n, k_p]$ and $S_2[n+1, k_p] = S_1^*[n, k_p]$, and the quasi-stationary channel property, $H_{1j}[n+1, k_p] = H_{1j}[n, k_p]$ and $H_{2j}[n+1, k_p] = H_{1j}[n, k_p]$, the following equivalent also holds:

$$e^{j2®\Delta£(n+1)} = \frac{\overline{Y}_j[n+1, k_p] - S_2^*[n, k_p]H_{1j}[n, k_p]}{S_1^*[n, k_p]H_{2j}[n, k_p]} \quad ...(30)$$

By (29), (30), and the property $\left|S_1[n, k_p]\right|^2 = \left|S_2[n, k_p]\right|^2 = 1$, we have

$$e^{j2®\Delta£} = \frac{\overline{Y}_j[n+1, k_p] - S_2[n, k_p]H_{1j}[n, k_p]}{\overline{Y}_j[n, k_p]S_1^*[n, k_p] + H_{1j}[n, k_p]} \quad ...(31)$$

Taking into account two receive antennas ... $\Delta£$ is

$$\Delta \pounds [n] = \frac{1}{2\circledR} \angle \sum_{j=1}^{2} \sum_{k_p=1}^{N_p} \frac{\overline{Y}_j[n+1, k_p] S_2[n, k_p] + H_{1j}[n, k_p]}{\overline{Y}_j[n, k_p] S_1^*[n, k_p] + H_{1j}[n, k_p]} \quad ...(32)$$

Note that the estimated transmitter CFO difference $\Delta \pounds$ requires a return path for feedback to the transmitter side in order to correct the mismatched oscillator frequency. The 2 x 2 MIMO OFDM receiver along with the new CFO estimation algorithm is evaluated in the following Section.

SYSTEM SIMULATION AND IMPLEMENTATION

System Parameters and Simulation Results

The 2 x 2 STBC MIMO OFDM system parameters are listed in Table 1. Four independent multipath channel models with exponentially decaying power profile of rms 50 ns are used in simulation. The estimated channel responses by using the even-odd interpolation method in (19) and (20) are shown in Figure 3, where 8 subcarriers are assigned with virtual carriers. Four different CFO cases are set as listed in Table 2 and are compared later by the proposed CFO estimation method. The structure of main functional blocks in the receiver is depicted in Figure 4. The sample timing correction block is designed by an independently digital interpolation module with the Farrow structure (Farrow, 1988; and Lin and Chang, 2007) in front of the frame detection module and is not plotted here. The depicted timing block includes frame detection and symbol timing where the delay correlator can also be used for coarse CFO estimation. The CFO correction module is divided into acquisition mode and tracking mode as shown in Figure 5. In the acquisition mode, the angle of the correlator output in response to the short preamble sequence is calculated by the coordinate rotation digital computer (CORDIC) and Direct Digital Frequency Synthesizer (DDFS) modules (Hsu and Chang, 2005) to reduce ICI due to CFO. The residual CFO and the frequency difference between transmit antennas are estimated in the tracking mode to avoid constellation rotation at the FFT output. Figure 6 shows the CFO tracking curves for case 4 by the proposed scheme. Also as shown in Figure 7, the value of the CFO difference between two transmit antennas is effectively estimated and the feedback for correction is assumed such that the Symbol

Table 1: Mimo OFDM System Parameters

Feature	MIMO-OFDM
Number of spatial streams	1
Number of Tx/Rx Antennas	2 x 2
Bandwidth	20 MHZ
Number of data subcarriers	54
Number of pilot subcarriers	2
Subcarrier frequency spacing	0.3125 MHZ
IFFT/FFT period	3.2 us
Gi duration	0.8 us
Training symbol gurad duration	1.6 us
Symbol interval	4 us
Modulation	16-QAM
Channel model power delay profile	Exponentially decaying

Table 2: CFO Correction Simulation Parameters

Feature (Freq. Hz)	Case 1	Case 2	Case 3	Case 4
Tx Antenna 1	2.4G	2.4G	2.4G	2.4G
Tx Antenna 2	2.4G	2.4G	2.4G	2.4G+1k
Rx Antenna 1	2.4G	2.4G+150k	2.4G+150k	2.4G+150k
Rx Antenna 2	2.4G	2.4G+150k	2.4G+200k	2.4G+200k

Figure 3: The Real Channel Responses and the Estimated Channel Responses in the 2 x 2 MIMO OFDM System

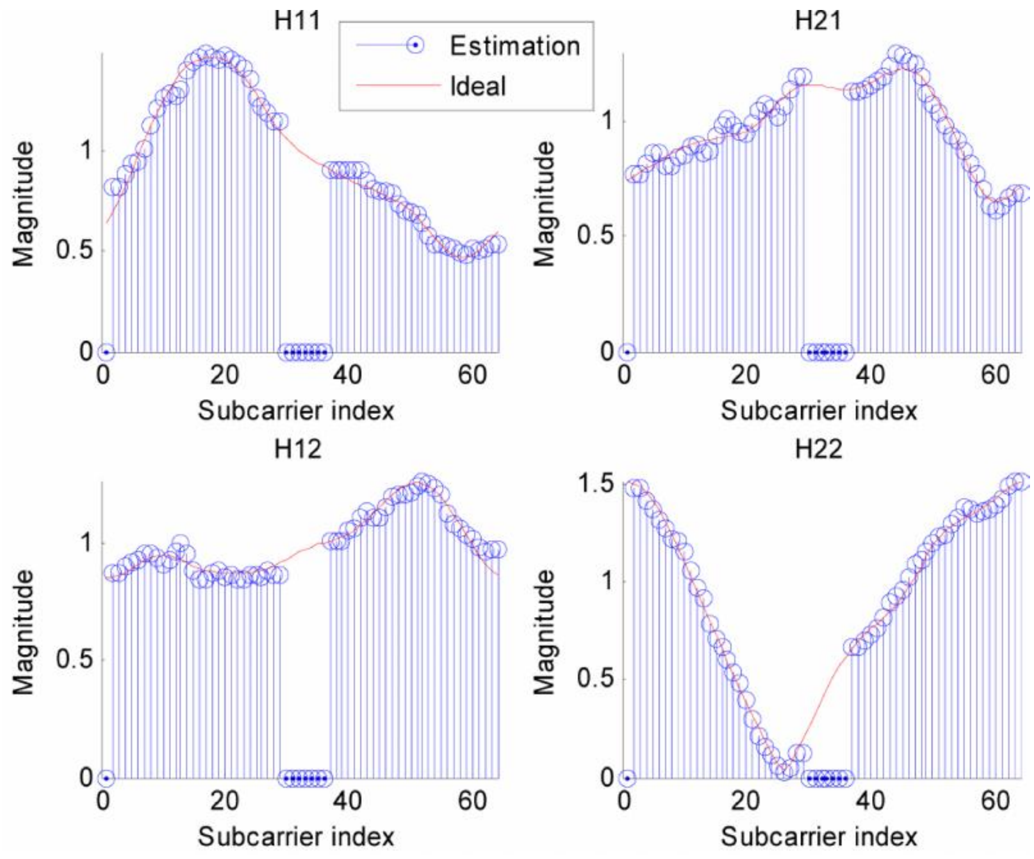


Figure 4: Main Functional Blocks in the MIMO OFDM Receiver

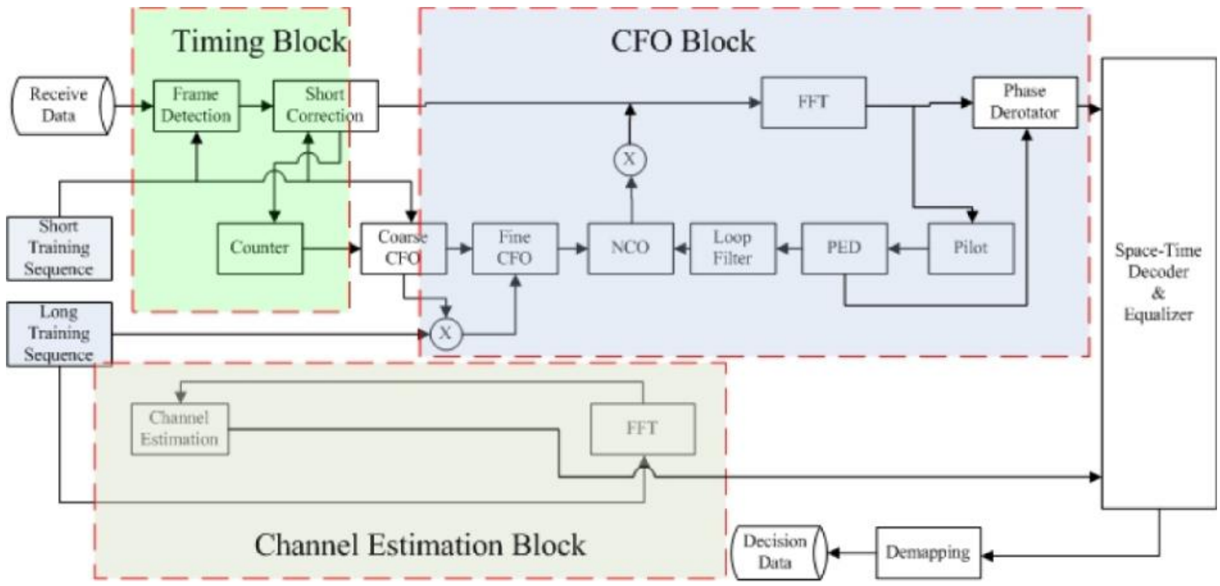


Figure 5: The CFO Correction Module Including Acquisition Mode and Tracking Mode

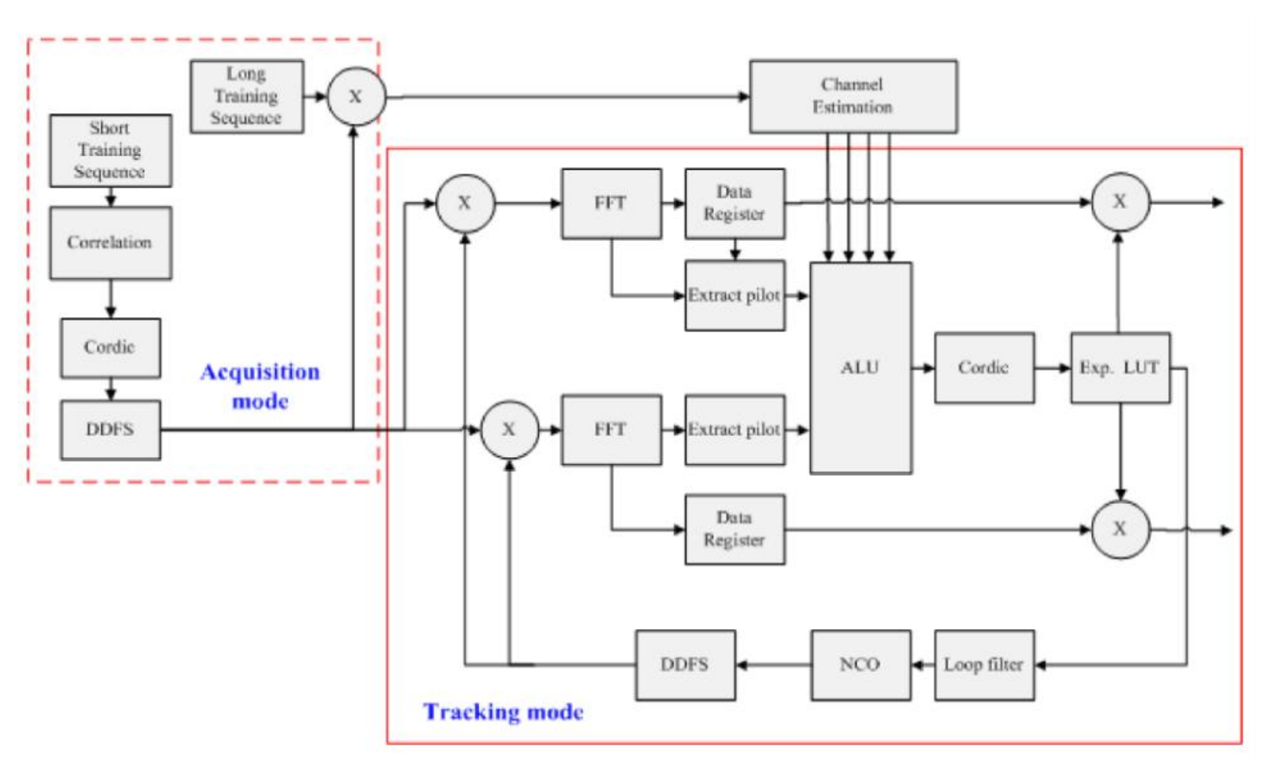


Figure 6: CFO Tracking Curves for Receive Antennas 1 and 2 by the Proposed Scheme

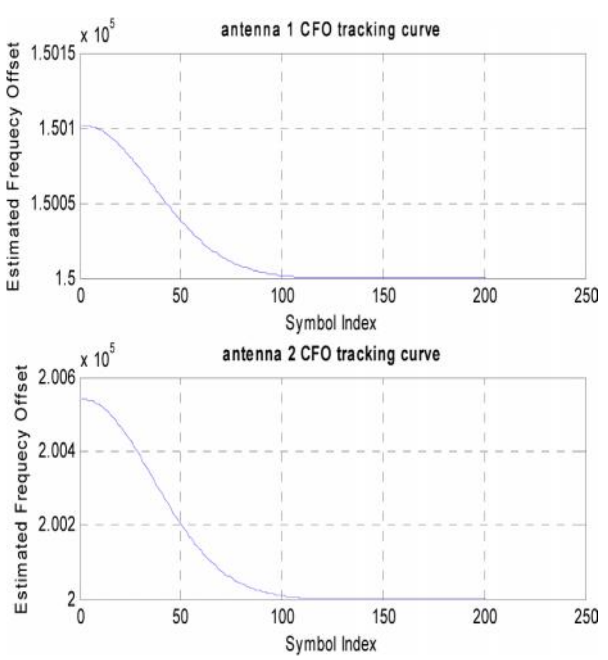
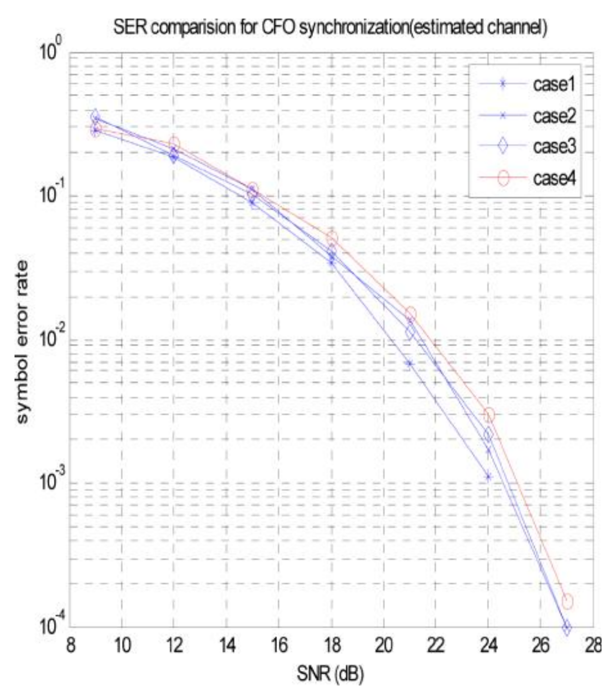


Figure 7: SER Performance Comparison for Different CFO Estimation Cases



Error Rate (SER) performance of the OFDM data payload is quite close to the ideal case.

Implementation and FPGA Verification

The FPGA verification flow for this work uses a Pattern Generator (PG) to create 2 x 2 MIMO OFDM transmitted signals with four independent channel distortions, and then, the Xilinx Virtex-II XC2V3000 FPGA is used to implement the VHDL receiver codes in order to demodulate the 16-QAM constellation. The FPGA resources occupied by individual modules are listed in Table III. Figure 8 shows the SER comparison of the floating point simulation results and the fixed point implementation, and Figure 9 shows the constellation results of demodulated data payload at SNR 25 dB. In this comparison, the overall system function including 40 ppm clock timing frequency offset correction, CFO estimation and correction, and channel estimation are activated. The result shows that the implementation loss is about 1.5 dB with this rapid prototyping FPGA design.

Figure 8: SER Performance Comparison of Floating Point Simulation and Fixed Point Implementation

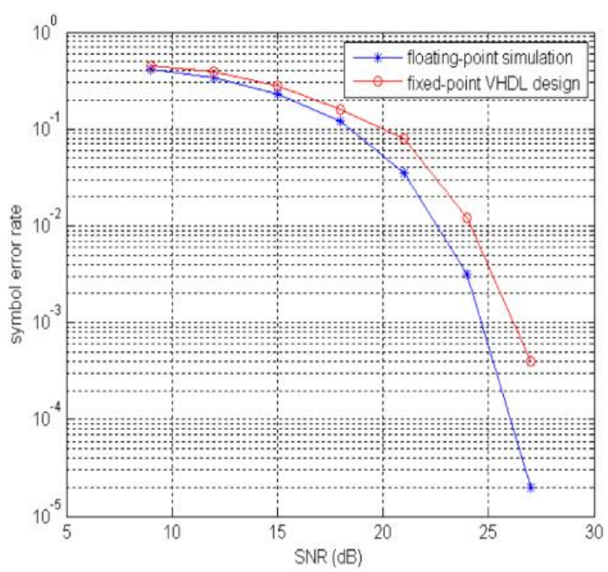


Figure 9: Comparison of Received Signal Constellation at SNR 25 dB

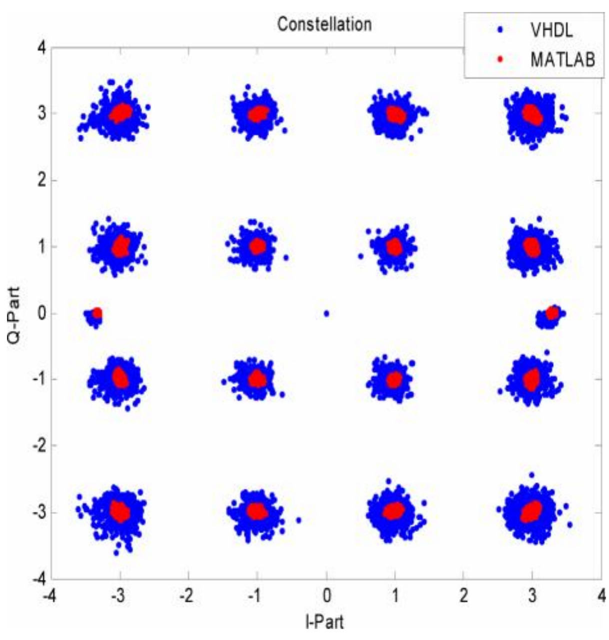


Table 3: The Occupied FPGA Resources

Modules	Slice (%)	Gate Count	Max Rate (MHz)
Timing Synchronization	10%	94,401	33.6
CFO Correction	3%	48,283	71.2
Channel Estimation	3%	9,707	82.4
Channel Buffer	5%	201,439	53.2
Remove CP	1%	21,108	184.9
FFT	17%	198,404	198
Tracking Algorithm	5%	62,985	20.5
Phase Compensation	1%	48,263	103.7
STBC Decoder	4%	208,785	81.5
Equalization	2%	7,025	17.5
Total	**55%**	**1,177,882**	**55.1**

CONCLUSION

A 2 x 2 STBC MIMO OFDM baseband inner receiver is designed based on an IEEE 802.11 n proposal to eliminate the effect of different CFOs caused by mismatched transmit antennas. A proposed CFO tracking algorithm has been shown to be able to combat the problem of different CFO between two transmit antennas. The overall receiver with the new CFO correction method is verified on the Xilinx FPGA with an implementation loss of about 1.5 dB.

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