A SIMPLE QUASI-ORTHOGONAL SPACE-TIME SCHEME FOR USE IN ASYNCHRONOUS VIRTUAL ANTENNA ARRAY ENABLED COOPERATIVE NETWORKS

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ABSTRACT

In this paper we present a novel full-rate and full-diversity quasi-orthogonal space-time block coding scheme for use in asynchronous cooperative relay networks. A variant of orthogonal frequency division multiplexing (OFDM) is implemented at the source node to mitigate the effects of random delays at the relay nodes, which operate in a simple amplify-and-forward mode. Utilizing a feedback channel, a simple phase rotation is applied at the relay nodes to extract full order diversity. A simple feedback quantization approach which reduces the overhead in the feedback channel for practical systems is also proposed. Average bit error rate simulations confirm the utility of the schemes.

***Index Terms*– Virtual antenna arrays (VAAs), space-time block codes (STBCs), closed-loop phase angle feedback, asynchronous communication**

1. INTRODUCTION

The next generation of wireless communication networks will require robust schemes that mitigate the effects of fading and provide synchronization over distributed nodal systems. Receive diversity techniques for combating fading are well established [1]; however in the last decade transmit techniques in the form of space-time block codes (STBCs) pioneered by Alamouti [2] have been gaining popularity. In recent years many schemes have utilized the simple encoding and maximum-likelihood (ML) decoding of STBCs in a distributed framework to implement cooperative diversity, for systems where it may not be practical or even feasible to employ multiple antenna array elements on the nodes.

In this paper we propose increasing the robustness of a simple asynchronous cooperative scheme [3] developed for flat-fading quasi-static channels through the use of a low-rate feedback channel to enable closed-loop space-time coding. By applying feedback in the form of a phase rotation at the transmitter, Toker et al. [4] showed that it was possible to extract full transmission rate and full diversity from a complex quasi-orthogonal design [5] with four transmit antennas; for complex valued orthogonal-STBCs it was shown [6] that full-rate full-diversity designs are not achievable for more than two transmit antennas.

Our contribution, based on the system model represented in Fig.1, mitigates the effects of timing asynchronicity between the relay nodes by implementing orthogonal frequency division multiplexing (OFDM) at the source node. This novel scheme will also provide additional robustness in the form of *fourth-order* diversity when perfect phase feedback is assumed. However, even when there are errors in the feedback channel the scheme is still shown to outperform a previously proposed Alamouti-OFDM [3] based implementation.

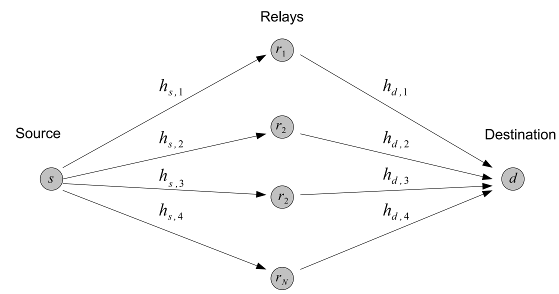

Figure 1: A simple two-stage asynchronous cooperative relay network architecture

The remaining sections of this paper are organized as follows. Section II introduces the system model and description of the two-stage relaying protocol. In this section the source OFDM encoding strategy is illustrated, along with the space-time coding technique at the relay nodes, and the common destination node processing. Section III describes extensions of the phase rotation technique [4] to handle the different delays along the paths through the four relay nodes. Both perfect and imperfect feedback channels are considered. Section IV presents simulation results and discusses the proposed scheme with ideal feedback and with a band-limited feedback channel.

Notation: In this paper $(\cdot)^*$ is used to denote conjugate, $(\cdot)^T$ transpose and $(\cdot)^H$ conjugate transpose. $\mathrm{D}(\mathbf{b})$ denotes a diagonal matrix created from vector $\mathbf{b}$.

2. SYSTEM MODEL

In our model Fig.1, we consider a new asynchronous space-time transmission scheme which assumes one source, four relay and one destination nodes, where each node only utilizes one antenna element. This scheme implements a two-stage transmission protocol.

2.1 First Stage - Source Node Processing

In the first stage the source node broadcasts sequentially four, $i \in \{1,\ldots,4\}$, orthogonal frequency division multiplexed (OFDM) symbols $\mathbf{s}^{(i)} = [s_0^{(i)},\ldots,s_{N-1}^{(i)}]^T$ composed of a set of N-coded quadrature amplitude modulation (QAM) symbols which are converted into time domain samples $\{\mathbf{t}^{(i)}\}$ using DFT and IDFT operations,

$$\mathbf{t}^{(1)} = \sqrt{P_s}F^H\,\mathbf{s}^{(1)} \quad (1)$$

$$\mathbf{t}^{(2)} = \sqrt{P_s}F\,\mathbf{s}^{(2)} \quad (2)$$

$$\mathbf{t}^{(3)} = \sqrt{P_s}F\,\mathbf{s}^{(3)} \quad (3)$$

$$\mathbf{t}^{(4)} = \sqrt{P_s}F^H\,\mathbf{s}^{(4)} \quad (4)$$

where F and F^H denote NxN-point unitary DFT and IDFT matrices respectively and P_s represents the source transmit power. Note that before transmission a cyclic prefix (CP) of length $N_p \le N$ is incorporated into $\{\mathbf{t}^{(i)}\}$. Each relay node j then receives a noisy copy of the transmitted signal perturbed by the channel response,

$$\mathbf{r}_j^{(i)} = H_{s,j}\,\mathbf{t}^{(i)} + \mathbf{v}_j^{(i)} \quad (5)$$

where $H_{s,j}$ denotes a diagonal channel matrix with constant diagonal elements $h_{s,j}$, i.e. a flat-fading quasi-static channel is assumed. Noise perturbations at the various relay nodes are represented by the vectors $\mathbf{v}_j^{(i)}$ with zero mean additive white Gaussian noise (AWGN) distributed elements of variance σ^2.

2.2 Relay Node Processing

In the second stage, relay j re-transmits the received time-domain samples according to the j^{th} column of the block QO-STBC matrix S,

$$S = \beta \begin{bmatrix} \mathbf{r}_1^{(1)} & -\mathbf{r}_2^{(2)*} & -\mathbf{r}_3^{(3)*} & \mathbf{r}_4^{(4)} \\ \zeta(\mathbf{r}_1^{(2)}) & \zeta(\mathbf{r}_2^{(1)*}) & -\zeta(\mathbf{r}_3^{(4)*}) & -\zeta(\mathbf{r}_4^{(3)}) \\ \zeta(\mathbf{r}_1^{(3)}) & -\zeta(\mathbf{r}_2^{(4)*}) & \zeta(\mathbf{r}_3^{(1)*}) & -\zeta(\mathbf{r}_4^{(2)}) \\ \mathbf{r}_1^{(4)} & \mathbf{r}_2^{(3)*} & \mathbf{r}_3^{(2)*} & \mathbf{r}_4^{(1)} \end{bmatrix} \quad (6)$$

$$\beta = \sqrt{\frac{P_r}{P_s+\sigma^2}} \quad (7)$$

where P_r denotes the transmit power at each relay and $\zeta(\cdot)$ is a modulo N time reversal operation illustrated below,

$$\zeta(\mathbf{a}) = \begin{bmatrix} a[N+N_p-1] \\ \vdots \\ a[0] \end{bmatrix} \quad (8)$$

In the absence of channel state information (CSI) at the relay nodes, $P_r = P_s/4$ provides optimal power allocation [1]. This simple relay encoding scheme (6) can be described under the umbrella term amplify-and-forward (AF), where each relay performs simple operations on the received noisy signal in the form of: conjugation, amplification and reordering. Note that using the coding scheme (6) the fourth relay has to wait for three OFDM symbols before commencing re-transmission; however, to shorten the length of CP insertion all relay nodes wait for three symbol durations before re-transmission.

2.3 Destination Node Processing

At the destination node, timing synchronization is easily implemented for the shortest path from source to destination. Without loss of generality, this will be assumed to be the path through relay 1. To process the data, CP removal is performed on the $N+N_p$ received vector using a rectangular window,

$$\varsigma(\mathbf{a}[\mathbf{n}]) = \begin{cases} \mathbf{a}[\mathbf{n}] & \text{for n =}\{N_p, N_p\text{+N-1}\} \\ 0 & \text{otherwise} \end{cases}$$

the position of the windowing operation is determined by the timing information gathered through synchronization with relay 1. This process is the same as in conventional OFDM systems and is illustrated in Fig.2 (time reversal effects have been omitted to simplify the illustration), where τ_j denotes the delay in sample periods of the path through relay j relative to that through relay 1.

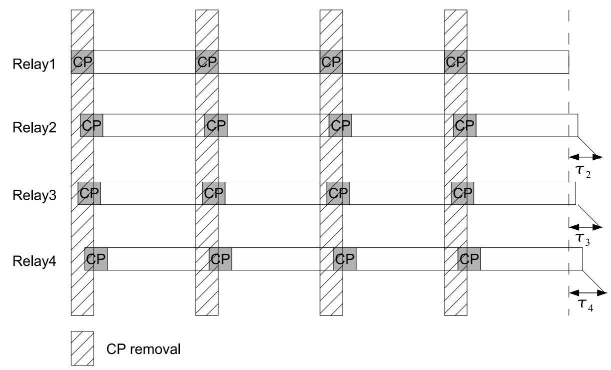

Figure 2: CP removal based on relay 1 synchronization

Post CP removal, reordering needs to be performed on the 2nd and 3rd received frames to correct for the misalignment caused by the time reversal $\zeta(\cdot)$ in (6). This can be performed either in the time or frequency domain as a circular shift

$$\zeta'(\mathbf{a}[n]) = \mathbf{a}[\langle n-(N_p+1)\rangle_N] \quad (9)$$

$$\zeta'(\mathbf{A}[k]) = e^{-j2\pi k(N_p+1)/N}\mathbf{A}[k] \quad (10)$$

where $\mathbf{a}$ and $\mathbf{A}$ denote an arbitrary vector $\mathbf{a}$ in the time and frequency domains respectively with associated sample/sub-carrier indexing parameters n and k. The combined time reversal $\zeta(\cdot)$, windowing $\varsigma(\cdot)$ and misalignment adjustment $\zeta'(\cdot)$ operations, assuming synchronous transmission, results in the overall reordering process,

$$\begin{bmatrix} a[0] \\ \vdots \\ a[N+N_p-1] \end{bmatrix} \rightarrow \begin{bmatrix} a[0] \\ a[N-1] \\ \vdots \\ a[1] \end{bmatrix} \quad (11)$$

which can easily be shown as equivalent to the permutation matrix $P = F^H F^H$. The destination node then computes an N-point DFT of the four consecutive received time-domain OFDM symbols, resulting in the following processed vectors,

$$\mathbf{y}^{(i)} = \begin{cases} \sum_{j=1}^{4} \mathbf{\Phi}_{\tau_j} F \varsigma(H_{d,j}[S]_{i,j}) + F\mathbf{v}_d^{(i)} & i = 1,4 \\ \sum_{j=1}^{4} \mathbf{\Phi}_{\tau_j} F \zeta'(\varsigma(H_{d,j}[S]_{i,j})) + F\mathbf{v}_d^{(i)} & i = 2,3 \end{cases} \tag{12}$$

where $H_{d,j}$ is used to denote the channel matrix between relay j and the destination node and is of the form $\mathrm{D}(\mathbf{h}_{d,j})$. The matrix Φ_{τ_j} reflects the phase offsets in the OFDM sub-carriers arising due to the asynchronous re-transmission delay between relay nodes,

$$\Phi_{\tau_j} = \mathrm{D}(\phi_{\tau_j}) \tag{13}$$

where,

$$\phi_{\tau_j}[k] = \begin{cases} e^{-j2\pi k\tau_j/N} & k = 0,1,\ldots,(N-2)/2 \\ e^{j2\pi(k-N)\tau_j/N} & \text{otherwise} \end{cases} \tag{14}$$

Note: for the purposes of this analysis Φ_{τ_1} can be substituted for the identity matrix $\mathbf{I}_N$ due to the assumed synchronization to relay 1. Observing the phase relationship in (14) a new sub-carrier index k' is introduced which has the following one-to-one mapping with the sub-carrier indexing variable k,

$$\begin{array}{ccc} k & \rightarrow & k' \\ \begin{bmatrix} 0 \\ \vdots \\ (N-2)/2 \\ N/2 \\ \vdots \\ N-1 \end{bmatrix} & \rightarrow & \begin{bmatrix} 0 \\ \vdots \\ (N-2)/2 \\ -N/2 \\ \vdots \\ -1 \end{bmatrix} \end{array}$$

For efficient implementation and a simplified analysis of (12) symbol estimation is considering purely in the frequency domain; in the following expressions the noise terms have been omitted for conciseness,

$$\begin{aligned} \mathbf{y}^{(1)} &= \mathrm{H}_1\mathbf{s}^{(1)} - \mathrm{H}_2\mathbf{s}^{(2)*} - \mathrm{H}_3\mathbf{s}^{(3)*} + \mathrm{H}_4\mathbf{s}^{(4)} & (15) \\ \mathbf{y}^{(2)} &= \mathrm{H}_1\mathbf{s}^{(2)} + \mathrm{H}_2\mathbf{s}^{(1)*} - \mathrm{H}_3\mathbf{s}^{(4)*} - \mathrm{H}_4\mathbf{s}^{(3)} & (16) \\ \mathbf{y}^{(3)} &= \mathrm{H}_1\mathbf{s}^{(3)} - \mathrm{H}_2\mathbf{s}^{(4)*} + \mathrm{H}_3\mathbf{s}^{(1)*} - \mathrm{H}_4\mathbf{s}^{(2)} & (17) \\ \mathbf{y}^{(4)} &= \mathrm{H}_1\mathbf{s}^{(4)} + \mathrm{H}_2\mathbf{s}^{(3)*} + \mathrm{H}_3\mathbf{s}^{(2)*} + \mathrm{H}_4\mathbf{s}^{(1)} & (18) \end{aligned}$$

where,

$$\mathrm{H}_j = \mathrm{D}(\mathrm{h}_j) = \begin{cases} \Phi_{\tau_j} F H_{d,j} H_{s,j} F^H & j = \{1,4\} \\ \Phi_{\tau_j} F H_{d,j} H_{s,j}^{*} F^H & j = \{2,3\} \end{cases}$$

In a synchronous implementation the channel matrices H_j assume a diagonal matrix structure populated with identical terms along the diagonal entries, i.e. $\mathrm{D}(\mathrm{h}_j)$. The introduction of asynchronous delay τ_j at the specific relay node j introduces a proportionally increasing (decreasing) linearly phase rotation in the frequency-domain across positive (negative) frequency sub-carriers; which is reflected in the channel vector $\mathrm{h}_j = [h_{j,0}, \ldots, h_{j,N}]^T$. *Note:* this analysis demonstrates that asynchronous delay does not introduce inter-carrier-interference (ICI) or inter-symbol-interference (ISI) when using a OFDM scheme.

3. RELAY PHASE ROTATION AND MAXIMUM LIKELIHOOD DECODING

3.1 Phase-rotation

Assuming matched filtering is performed at the destination node for each of the $k = 0, 1, \ldots, N-1$ sub-carriers, then because of the quasi-othogonality of the coding scheme each symbol estimate is perturbed by inter-symbol-interference (ISI); as shown in the expression below [4],

$$\begin{bmatrix} \gamma_k & 0 & 0 & \alpha_k \\ 0 & \gamma_k & -\alpha_k & 0 \\ 0 & -\alpha_k & \gamma_k & 0 \\ \alpha_k & 0 & 0 & \gamma_k \end{bmatrix} \begin{bmatrix} \mathbf{s}_k^{(1)} \\ \mathbf{s}_k^{(2)} \\ \mathbf{s}_k^{(3)} \\ \mathbf{s}_k^{(4)} \end{bmatrix} \tag{19}$$

where $\gamma_k = \sum_{j=1}^{4} \|\mathrm{h}_{j,k}\|^2$ and $\alpha_k = 2Re\left\{\mathrm{h}_{1,k}\mathrm{h}_{4,k}^{*} - \mathrm{h}_{2,k}\mathrm{h}_{3,k}^{*}\right\}$. To cancel the α_k terms Toker et al. [4] proposed simple phase rotations, denoted by ϕ_k and θ_k, at the third and fourth transmit antennas, corresponding to relay nodes three and four in our scheme. Although the original implementation of [4] was designed for a single-stage single-carrier implementation, this can be extended to our scheme by applying Toker's algorithm for each OFDM sub-carrier independently using the equivalent frequency-domain source to destination channel coefficients $h_{j,k}$. Applying the phase rotations at the appropriate nodes then enables the cross diagonal α_k terms to be canceled, i.e. $\alpha'_k = 0$,

$$\begin{aligned} \alpha'_{k'} = \; & 2Re\{\mathrm{h}_{1,1}\mathrm{h}_{4,1}^{*} e^{j(\psi k' + \theta_{k'})} \\ & -\mathrm{h}_{2,1}\mathrm{h}_{3,1}^{*} e^{j(\omega k' + \phi_{k'})}\} \end{aligned} \tag{20}$$

where $\psi = 2\pi\tau_4/N$ and $\omega = 2\pi(\tau_3 - \tau_2)/N$ denote the incremental phase offset introduced between sub-carriers by the asynchronous delay of the fourth relay node and the delay difference between the second and third nodes respectively.

However, this simple phase relationship across sub-carriers simplifies the calculation of the phase rotators $\phi_{k'}$ and $\theta_{k'}$ to,

$$\theta_{k'} = \theta_1 - \psi k' \tag{21}$$

$$\phi_{k'} = \phi_1 - \omega k' \tag{22}$$

therefore only requiring the computation of the phase offset variables for the first sub-carrier, i.e. θ_0 and ϕ_0, knowledge about the CSI for the first sub-carrier and the asynchronous offset delays τ_2, τ_3 and τ_4. This then reduces the feedback requirements for our proposed scheme to the quantized parameters ϕ_1, θ_k, ω and ψ enabling the relay nodes to interpolate the necessary phase rotations over each individual sub-carrier.

Assuming ideal error free feedback between the destination and relay nodes three and four and perfect synchronization between the source and relay nodes; the phase rotation can be implemented at each node using an all-pass linear-time-invariant (LTI) filter calculated using the computationally efficient FFT/IFFT algorithm. This is illustrated in the following phase rotated frequency domain channels,

$$\mathrm{H}_3 = \Phi_{\tau_3} F H_{d,3} F^H \mathrm{D}(\phi) F H_{s,3}^* F^H \quad (23)$$

$$\mathrm{H}_4 = \Phi_{\tau_4} F H_{d,4} F^H \mathrm{D}(\theta) F H_{s,4} F^H \quad (24)$$

3.2 Decoding

Assuming that perfect phase feedback is not realizable, the estimation of the transmitted symbols $s^{(1)}$ to $s^{(4)}$ for a particular sub-carrier cannot be performed independently of one another as is normally the case for orthogonal-STBC designs [1]. However the quasi-orthogonal properties of the coding scheme used allow decoupled estimation of $s^{(1)}$ and $s^{(4)}$ from $s^{(2)}$ and $s^{(3)}$ therefore marginally increasing the decoding complexity from an order of $O(lM)$ to $O(2M^{\frac{l}{2}})$; where l denotes the number of symbols transmitted for a given STBC scheme and M denotes the modulation order.

In this paper we adopt the notation used in Jafarkhani's analysis of ML decoding [5] and apply it to our scheme. The ML decision metric for estimating the symbols transmitted for each sub-carrier can be calculated as the sum of two terms $f_{14}(x_1,x_4)$ and $f_{23}(x_2,x_3)$ where omitting the subcarrier indices k it can be shown that,

$$\begin{aligned} f_{14}(x_1,x_4) = & \left(\sum_{i=1}^{4} |\mathrm{h}_i|^2\right)\left(|x_1|^2+|x_4|^2\right) \\ & +2Re\{(-\mathrm{h}_1 r(1)^* - \mathrm{h}_2^* r(2) - \mathrm{h}_3^* r(3) \\ & -\mathrm{h}_4 r(4)^*)x_1 + (-\mathrm{h}_4 r(1)^* + \mathrm{h}_3^* r(2) \\ & +\mathrm{h}_2^* r(3) - \mathrm{h}_1 r(4)^*)x_4 + (\mathrm{h}_1\mathrm{h}_4^* \\ & -\mathrm{h}_2^*\mathrm{h}_3 - \mathrm{h}_2\mathrm{h}_3^* + \mathrm{h}_1^*\mathrm{h}_4)x_1x_4^*\} \end{aligned} \quad (25)$$

$$\begin{aligned} f_{23}(x_2,x_3) = & \left(\sum_{i=1}^{4} |\mathrm{h}_i|^2\right)\left(|x_2|^2+|x_3|^2\right) \\ & +2Re\{(-\mathrm{h}_1 r(2)^* + \mathrm{h}_2^* r(1) - \mathrm{h}_3^* r(4) \\ & +\mathrm{h}_4 r(3)^*)x_2 + (-\mathrm{h}_1 r(3)^* - \mathrm{h}_2^* r(4) \\ & +\mathrm{h}_3^* r(1) + \mathrm{h}_4 r(2)^*)x_3 + (\mathrm{h}_2\mathrm{h}_3^* \\ & -\mathrm{h}_1^*\mathrm{h}_4 - \mathrm{h}_1\mathrm{h}_4^* + \mathrm{h}_2^*\mathrm{h}_3)x_2x_3^*\} \end{aligned} \quad (26)$$

where the ML estimate is assumed to minimize the decision metric when the search of x_i is performed over a finite symbol alphabet. Note that when perfect phase rotation (20) is applied at the relay nodes the coupled terms in (25) and (26) are cancelled.

4. SIMULATION RESULTS

In this section we provide simulation results for our proposed scheme. For all our simulation results we assume perfect CSI and equal signal-to-noise ratio (SNR) at all receiving nodes. For simulation comparisons we utilize OFDM coding

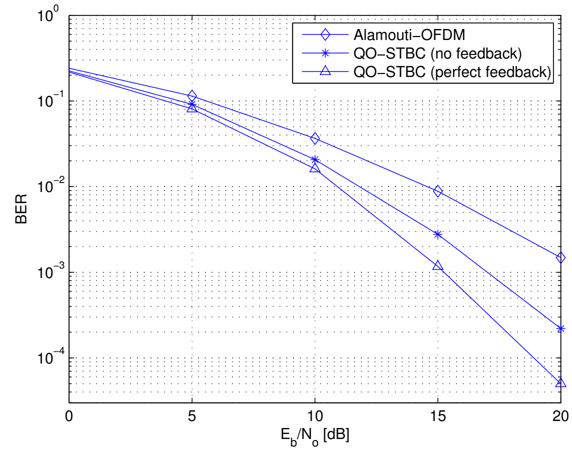

Figure 3: Average BER comparison of the proposed schemes with and without ideal feedback vs. Alamouti-OFDM

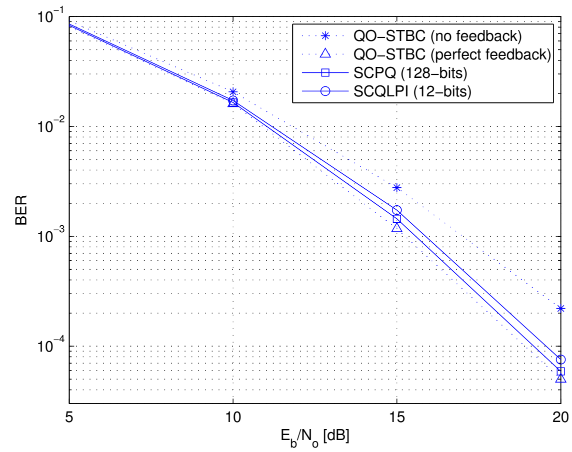

Figure 4: Average BER analysis of the proposed scheme with different feedback quantization approaches

at the source node where; 64 sub-carriers are used (N=64), a CP length of 16 is assumed and the asynchronous delay of each relay node j is uniformly distributed over the interval $\tau_j \in [0,15]$, with the exception of the first relay node where synchronization with the destination node is assumed. To simplify the ML decoding process the symbols are randomly drawn from a constant modulus constellation, i.e. BPSK.

A simple full-rate full-diversity Alamouti coded OFDM (Alamouti-OFDM) scheme using an AF protocol at the relays was proposed in [3]. Fig.3 compares our full-rate quasi-orthogonal scheme with the Alamouti-OFDM implementation and illustrates the performance improvements achieved by phase rotation at the relay nodes, i.e. phase feedback enabled. In addition it is shown that even without feedback our scheme demonstrated second order diversity and improved BER performance for all SNR values over the Alamouti-OFDM scheme; therefore ensuring a more robust transmission scheme in the event of feedback channel failure.

In an implementable system some quantization of the phase rotation to the relay nodes is essential, therefore two simple strategies are adopted to obtain the results in

Fig.4. The first strategy, termed sub-carrier phase quantization (SCPQ), quantizes the phase rotation variables $\phi_k, \theta_k \in \{0, \pm\frac{\pi}{2}, \pi\}$ by only feeding back 2-bits per sub-carrier to relay nodes three and four; therefore creating an overhead of 128 bits per feedback channel. The second strategy, referred to on Fig.4 as sub-carrier quantized linear phase interpolation (SCQLPI), adopts the same encoding approach previously used for the first sub-carrier, $k = 0$. In addition a quantized linear phase offset, exploiting the linear relationship in (21,22), is encoded to the nearest half-sampled delay interval, i.e. $\{0, 1/2, \ldots, 15\}$ therefore providing a resolution to the nearest half-sampling interval. This SCQLPI scheme reduces the feedback overhead in our simulations to 12-bits [providing a feedback overhead saving in excess of 90%], where 4-bits account for the quantized phase rotation for the first sub-carrier [k=0] and the remaining 8-bits determine the delay offset correction between sub-carriers. As shown in Fig.4 SCQLPI performs favorably in contrast to SCPQ, and the feedback overhead is proportional to the maximum asynchronous delay between relay nodes in contrast to the SCPQ scheme which is proportional to the OFDM symbol length.

Note: orthogonal STBC designs with a code-rate less than unity are not included in the discussion due to inferior theoretical channel capacity and throughput performance in virtual antenna arrays [7].

5. CONCLUSION

This paper demonstrates a novel space-time transmission scheme for asynchronous cooperative relay networks utilizing four relay nodes. Using a feedback channel to the relay nodes was shown to increase the diversity order of the quasi-orthogonal coding scheme from second to fourth order, thereby improving the robustness of the wireless links between source and destination nodes in the presence of fading. The simulation results verify that in the event of feedback failure, this proposed scheme still outperforms all other full-rate STBC designs for amplify-and-forward relaying. Also the paper presents a simple maximum-likelihood decoder, demonstrating that decoding can be performed on a symbol-by-symbol basis; assuming that in the event of asynchronous delay between relay nodes, orthogonality between sub-carriers is preserved through cyclic-prefix insertion. Finally, the paper demonstrates that the overhead for the feedback channel in the practical system can be substantially reduced through phase rotation interpolation over the various OFDM sub-carriers.

6. ACKNOWLEDGMENT

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