A COMPLEXITY REDUCTION TECHNIQUE FOR MULTI HOP RELAY NETWORK WITH OUT SIGNAL INFORMATION

A.VENKATA HANUMALEELA[1], M.NAGA MANI[2]

[1] *A.V.H.LEELA, PG Student, Department of Electronics &Communication Engineering, SRKIT, Vijayawada, India*
[2]*M.NAGA MANI, Assistant professor, Department of Electronics &Communication Engineering, SRKIT, Vijayawada, India*

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Abstract - *Improved Differential distributed space-time coding (I-DDSTC) has been consider to improve both diversity and data rate in the absence of channel (signal) information [CSI]. This paper proposes a new differential encoding and decoding process for D-DSTC systems with two relays. The proposed method is robust against synchronization errors and does not require any channel information at the destination. Simulation results shows less complexity for various synchronization errors without CSI at receiver*.

Key Words: **Differential Distributed space time coding, diversity, differential encoding and decoding, synchronization errors, OFDM, relay networks.**

1. INTRODUCTION

Cooperative communication techniques are used in a network can listen to a source during its transmission phase, they are able to re-broadcast the received data to the destination in another phase. Therefore, the overall diversity and performance of a network would benefit cooperatively by multiple users. Depending on the protocol, relays are used to process and re-transmit the received signal to the destination.
On the other hand, due to the dispensed nature of relay networks, the received signals from relays at the destination are not always aligned in the symbol level. They are known as synchronization errors between relays, causes inter symbol interference (ISI).

1.1 DSTC Network

A differential encoding and decoding process is designed to contest synchronization error when neither CSI nor synchronization delay are available at the destination. We consider the case that a source communicates with a destination via two relays and the received signals from the two relays may not be aligned. All channels are assumed to be Rayleigh flat-fading and slowly changing over time. Differential encoding and decoding are combined with an OFDM approach to circumvent both channel estimation and the ISI. At the source, differential encoding and the Inverse Discrete Fourier Transform (IDFT) are employed. At the relays, essential configuration and a protecting guard are applied as will be detailed later. At the destination, the Discrete Fourier Transform (DFT) and differential decoding are utilized to obtain a symbol-by-symbol decoding with low complexity. The proposed method does not require any CSI or the amount of synchronization error and provides significant performance improvement compared to cases with symbol misalignment.In DSTC networks, the relays cooperate to combine the received symbols by multiplying them with a fixed or variable factor and forward the resulting signals to the destination. The cooperation is such that a space-time code is effectively constructed at the destination. Coherent detection of transmitted symbols can be achieved by providing the instantaneous channel state information (CSI) of all transmission links at the destination. Although this requirement can be accomplished by sending pilot (training) signals and using channel estimation techniques.

1.2SYSTEM MODEL

Consider one source, two relays and one destination. All nodes have one antenna and transmission is half-duplex. The wireless channels between the nodes are all flat-fading channels.

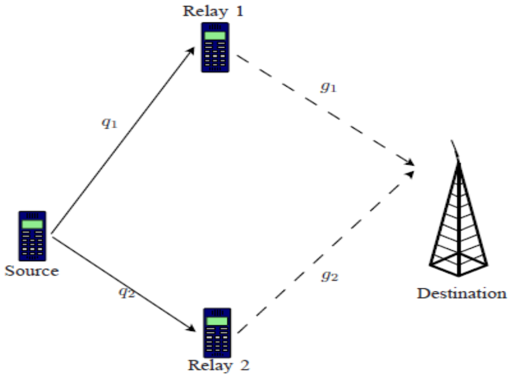

Fig.1. Cooperative network under consideration, Source communicates with Destination through two relays.

In the conventional DDSTC system, information bits are converted to symbols from constellation set V (such as PSK, QAM) at Source. Let us assume that two symbols $v_1, v_2 \in V$ are going to be sent from Source to Destination. The transmission process is divided into two phases and sending two symbols from Source to Destination in two phases is referred to as "one transmission block", indexed by $k \in Z$. First, symbols are encoded to a Unitary space-time coding (USTC) matrix as

$$V^{(k)} = \frac{1}{\sqrt{|v_1|^2 + |v_2|^2}} \begin{bmatrix} v_1 & -v_2^* \\ v_2 & v_1^* \end{bmatrix} \quad (1)$$

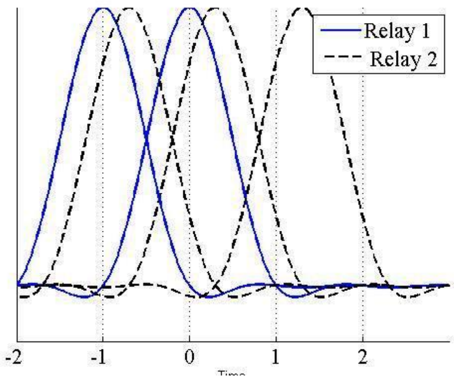

Fig2. Received signals from Relays 1 and 2 at Destination after the matched filter using a raised-cosine pulse-shape with roll-off factor $\beta = 0.9$.

The received signals at Destination at block-index (k) can be written as

$$y_1^{(k)} = g_1 x_{11}^{(k)} + g_{20} x_{21}^{(k)} + g_{21} x_{22}^{(k-1)} + n_1^{(k)}$$
$$y_2^{(k)} = g_1 x_{12}^{(k)} + g_{20} x_{22}^{(k)} + g_{21} x_{21}^{(k)} + n_2^{(k)} \quad (2)$$

Where $g_{20} = p(\tau) g_2, g_{21} = p(T_S - \tau) g_2$, and $n_j^{(k)} \sim CN(0, N_0)$, j=1,2 are the noise element at Destination. Thus, the effect of synchronization error is collected into quantities g_{20} and g_{21}.Depending on the number of side-lobes of the pulse-shape filter, more terms may appear in (2). In our model, the small contributions of the side-lobes of $p(t)$ are neglected .if the information is not available (system under consideration), one can treat the ISI as noise. In this case, assuming that channel coefficients are constant during two consecutive blocks, the data symbols can be conventionally decoded as

$$\hat{v}1, \hat{v}2 = \arg\min_{V} \left\| y^{(k)} - V^{(k)} y^{(k-1)} \right\| \quad (3)$$

1.3REVIEW OF PREVIOUS METHOD

In this section consider a GSM system with ordered block MMSE detector. The MIMO channel is assumed to be of quasi-static frequency flat fading, and hence can be represented by a $N_r X N_t$ channel matrix H. Consequently, the received signal y can be described as y = Hx + w, where x denotes the transmitted symbol vector, and w denotes the zero-mean circularly-symmetric complex Gaussian random noise vector with covariance matrix .

In GSM system, only N_p transmit antennas are activated at each time instance. Consequently, there are $C_{N_p}^{N_t}$ possible TACs. Among these TACs, N = $2^{\log_2 (C_{N_p}^{N_t})}$ TACs are chosen to convey $\log_2 N$ bits of information. Since only antennas are active, and are assumed to be drawn i.i.d. from a M-ary constellation set. We first denote the channel matrix associated to the TAC as $H_I = [B,c]$, where B and c are formed by the first Np – 1 column vectors and the last column vector of H_I respectively. By using the property of projection matrices, we have $P_{H_I} = P_{[B,c]} = P_B$ and $P_{(I-P_B)c}$, where P_B and $P_{(I-PB)c}$ denote projection matrices associated to the subspaces spanned by the columns of B and $(I-P_B)c$ respectively. It follows that for any two arbitrary vectors f $C^{N_r \times 1}$ and g $C^{N_r \times 1}$, one can compute

$$f^H P_{H_I} g = f^H P_{[B,c]} g = f^H P_{H_B} g + f^H P_{(I-P_B)C} g$$

. for special case f=g=y then we obtain

$$y^H P_{H_I} y = y^H P_{[B,c]} y = \frac{y^H P_B y + |c^H y - c^H P_B y|^2}{\|c\|^2 - c^H P_B c}$$

With the above procedure, we can write the explicit expression of $y^H P_{H_I} y$ for any Np - 2 in terms of inner products among the columns H_I and y. Here we provide the explicit expression for the Np = 3 as an example.
From the above discussion, it is clear that the computation Of $u_j = y_H P_{H_I} y$ can always be decomposed into some basic arithmetic operations of inner products between the columns of H_I and the data vector y, and also the inner products between the columns of H_I . Let the weighting factors be sorted as $u_{m_1} \_ u_{m_2} \_ . . . \_ u_{m_N}$. Since there are only finitely many (more precisely) possible channel vectors, it is expected that when we compute u_{mj} for some relatively large j, many of the required inner product terms may have already been computed when computing $u_{m_1} \_ u_{m_2} \_ . . . \_ u_{m_{(j-1)}}$. This motivates the development of the previous CECML computation algorithm in which the redundant computation of these inner product terms is avoided.
A summary of the proposed CECML ordering algorithm for the case of Np = 3 is described as in Table I.

TABLE1:summary of CECML algorithm

input:TAC sets $\{I_j\}_{j=1}^N$ and channel matrix $H_{\hat{I}} C^{N_r X N_t}$
output:weighting factors $\{u_j\}_{j}^{N=1}$
1.set $\Gamma_{hh}=0^{N_t X N_t}$, $\Phi_{hh}=0^{N_t X N_t}$, $\gamma_{hy}=0^{N_t X1}$
2.for l=1 to N_t do
3. $[\Gamma_{hh}]_{l,l}=\|h_l\|^2$, $[\Phi_{hh}]_{l,l}=1$, $[\gamma_{hy}]_l=h_l^H y$
4.end for
5.for j=1 to N do
6.if $[\Phi_{hh}]_{j_2,j_1}==0$ then
7.$[\Gamma_{hh}]_{j_2,j_1}==h_{j_2}^H h_{j_1}$, $[\Phi_{hh}]_{j_2,j_1}=1$
8.$[\Gamma_{hh}]_{j_1,j_2}==[\Gamma_{hh}]^*_{j_2,j_1}$, $[\Phi_{hh}]_{j_1,j_2}=1$
9.end if
10.if $[\Phi_{hh}]_{j_3,j_1}==0$ then
11.$[\Gamma_{hh}]_{j_3,j_1}==h_{j_3}^H h_{j_1}$, $[\Phi_{hh}]_{j_3,j_1}=1$
12.$[\Gamma_{hh}]_{j_1,j_3}==[\Gamma_{hh}]^*_{j_3,j_1}$, $[\Phi_{hh}]_{j_1,j_3}=1$
13.end if
14.if $[\Phi_{hh}]_{j_3,j_2}==0$ then
15.$[\Gamma_{hh}]_{j_3,j_2}==h_{j_3}^H h_{j_2}$, $[\Phi_{hh}]_{j_3,j_2}=1$
16.$[\Gamma_{hh}]_{j_2,j_3}==[\Gamma_{hh}]^*_{j_3,j_2}$, $[\Phi_{hh}]_{j_2,j_3}=1$
17.end if
18.compute u_j using $y^H P_{[t1,t2,t3]} y = y^H P_{[t1,t2]} y$+num/den with following quantites
$t_1^H y=[\gamma_{hy}]_{j_1}$, $t_2^H y=[\gamma_{hy}]_{j_2}$, $t_3^H y=[\gamma_{hy}]_{j_3}$,
$\|t_1\|^2=[\Gamma_{hh}]_{j_1,j_1}$, $\|t_2\|^2=[\Gamma_{hh}]_{j_2,j_2}$
$\|t_3\|^2=[\Gamma_{hh}]_{j_3,j_3}$
$t_1^H t_2=(t_1^H t_2)^*=[\Gamma_{hh}]_{j_1,j_2}$
$t_1^H t_3=(t_1^H t_3)^*=[\Gamma_{hh}]_{j_1,j_3}$
$t_2^H t_3=(t_2^H t_3)^*=[\Gamma_{hh}]_{j_2,j_3}$
19.end for

The algorithm first initializes the elements in Γ_{hh}, $\Phi_{h,h}$, and γ_{hy} to be all zeros. The (l,m)th element of Γ_{hh} will used to store the value of $h_l^H h_m$ after it has been computed. The (l,m)th element of $\Phi_{h,h}$ is set to 1 if $[\Gamma_{hh}]_{l,m}$ has been computed and then stored, while lth element of γ_{h_y} is used to store the value of $h_l^H y$ after it has been computed. The CECML algorithm provides substantial complexity reduction at high SNR region and hence exhibits a better performance-complexity having channel state information at destination. Here, receiver already knows that which information can be transmitted from source to destination .if the source and destination are in synchronization or without delay of information at each time slot (synchronous condition) we are using MMSE detector to reduce the bit error rate. To reduce the bit error rate (complexity) without using any detector in destination we proposed a method in next section.

2. PROPOSED METHOD

In this section, we propose a method for combating the synchronization error in the above system. The method combines differential encoding and decoding with an OFDM approach and is referred to as Differential OFDM (D-OFDM) DSTC. To establish the notation, first a brief review of OFDM systems is provided.

2.1 OFDM System

Frequency selective channels are usually modeled with finite impulse response (FIR) filters in the base-band. The channel output is the convolution of the channel impulse response and input sequence which leads to ISI. OFDM is a low complexity approach to deal with the ISI encountered in frequency-selective channels as explained in the following. Let $\{x[n]\}$, $n=0,\ldots,N-1$ represent the data symbols of length N and $\{h_0,\ldots,h_{L-1}\}$ represent the discrete time channel of length L. The N-point IDFT define as $X[m]=\text{IDFT}\{x[n]\}=1/\sqrt{N}\sum_{n=0}^{N-1} x[n]\exp(j2\pi nm/N)$ is applied to obtain sequence. Let us assume that the additive noise is zero. The channel output sequence, after

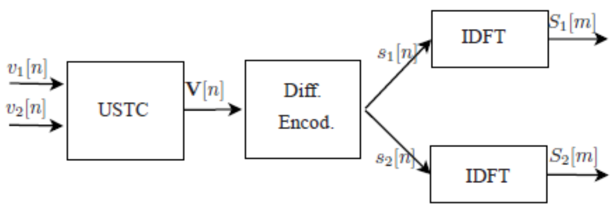

Fig4. Encoding process at Source

Removing the first L received symbols, then we get $Y[m]=h_1 \text{Ä} X[m]$, $m=0,\ldots,N-1$.From $y[n]=\text{DFT}\{Y[m]\}=1/\sqrt{N}\sum_{m=0}^{N-1} Y[m]\exp(-j2\pi mn/N)$ is applied. Using OFDM, the ISI is removed and the L-tap frequency-selective channel is converted to N parallel flat-fading channels.

2.2Differential OFDM DSTC

Using Eq(2), the effect of synchronization error is modeled by a frequency-selective channel with two taps and the OFDM method is utilized to remove the ISI. Similar to the conventional method, a two-phase transmission process is employed. However, instead of two symbols, a sequence of symbols will be transmitted during each phase. In Phase I, Source encodes data information as depicted in Fig. 4 and transmits 2N symbols to the relays. Then, the relays apply a special configuration, append $2L$ symbols and transmit 2(N+L) symbols to Destination in Phase II. Finally, Destination removes the symbols $2L$ and decodes the 2N symbols. Transmission of $2N$ symbols from Source to Destination in two phases is referred to as “one block transmission”, indexed by $k \in Z$. The two sequences are then encoded to USTC matrices based on (1) to differentially encoded as

Obtain $\{V[n]\}$, $n=1,\ldots,N$. Next, matrices $\{V[n]\}$ are

$$s[n]^{(k)}=V[n]^{(k)}s[n]^{(k-1)}=\begin{bmatrix} s_1[n] \\ s_2[n] \end{bmatrix}$$

$$s[n]^{(0)}=[1 \quad 0]^t, \quad n=0,\ldots,N-1, \qquad (9)$$

for simplicity of notations, the block-index (k) is omitted.

Then, the N-point IDFT is applied to $\{s_1[n]\}$, $\{s_2[n]\}$ then $S_1[m]=\text{IDFT}\{s_1[n]\}$ and $S_2[m]=\text{IDFT}\{s_2[n]\}$, $m=0,\ldots,N-1$. The obtained sequences $S_1[m]$ and $S_2[m]$ are then transmitted consecutively from Source to the relays over two sub-blocks, in Phase I. The received signals at the relays for $m=0,\ldots,N-1$ are expressed as

$$R_{ij}[m]=\sqrt{2P_0}\,q_i S_j[m]+Z_{ij}[m], \quad i,j=1,2 \qquad (10)$$

Where $Z_{ij}[m]$ □ $CN(0,N_0)$ are the noise elements at Relay i and sub-block j. The received signals at Relays 1 and 2 for $m=0,\ldots,N-1$ are configured as

$$X_{1j}[m]=A\,R_{1j}[m], \qquad j=1,2,$$
$$X_{21}[m]=-A\,\overset{\circ}{R}{}^{*}_{22}[m],$$
$$X_{22}[m]=A\,\overset{\circ}{R}{}^{*}_{21}[m], \qquad (11)$$

Where A is the amplification factor and $\overset{\circ}{R}_{2j}[m]$ is the circular time-reversal [15] of $R_{2j}[m]$ defined as

$$\overset{\circ}{R}_{2j}[m]=\begin{cases} R_{2j}[0], & m=0 \\ R_{2j}[N-m], & \text{otherwise.} \end{cases} \qquad (12)$$

Before transmission, the last L symbols of sequences $\{X_{ij}[m]\}$,i, j=1,2 are appended to their beginnings as the cyclic prefix to obtain

$$\{X_{ij}[N-L],\ldots,X_{ij}[N-1],\ldots,X_{ij}[N-1]\}$$

With length $N+L$. Here, L is the cyclic prefix length determined based on the amount of delay between the received signals from both relays and will be discussed shortly.

In Phase II, Relay 1 transmits sequences $\{X_{11}[m]\}$ and $\{X_{12}[m]\}$, while Relay 2 transmits $\{X_{21}[m]\}$ and

$\{X_{22}[m]\}$, for m=L,...,.,...,N-1, during two consecutive sub blocks or 2(N+L) symbols, to Destination.

At Destination, without loss of generality, let us assume that the received signal from Relay 2 is $(dT_S+\tau)$ seconds delayed with respect to that of Relay 1, where d is an integer number and $0\le\tau\le T_S$. Thus, to avoid ISI, the cyclic-prefix length is determined as L>d. If the delay, as shown in Fig. 2, is less than one symbol duration, L=1is enough. In practice the relays do not need to know the delay and, based on the propagation environment, the maximum value of d in the network can be estimated and used to determine the cyclic prefix length.

In this case, the received signals during two sub-blocks, after removing the first L symbols, can be expressed as

$$\begin{aligned}Y_1[m]&=g_1X_{11}[m]+g_{20}X_{21}[m-d]+g_{21}X_{21}[m-1-d]+W_1[m]\\&=g_1X_{11}[m]+(g_2[m]\otimes X_{21}[m-d])+W_1[m],\ m=0,\dots,N-1\end{aligned}\quad(13)$$

$$\begin{aligned}Y_2[m]&=g_1X_{12}[m]+g_{20}X_{22}[m-d]+g_{21}X_{22}[m-1-d]+W_2[m]\\&=g_1X_{12}[m]+(g_2[m]\otimes X_{22}[m-d])+W_2[m],\ m=0,\dots,N-1\end{aligned}\quad(14)$$

Where $g_2[m]=\sum_{l=0}^{1}g_{2l}\delta[m-l]$.

By substituting (11) and (10) into the above equations, one obtains, for $m=0,\dots,N-1$

$$\begin{aligned}Y_1[m]&=A\sqrt{2P_0}\left(h_1S_1[m]-h_2[m]\otimes \overset{o}{S}{}_2^*[m-d]\right)\\&\quad+A\left(g_1Z_{11}[m]-g_2[m]\otimes \overset{o}{Z}{}_{22}^*[m-d]\right)+W_1[m]\end{aligned}\quad(15)$$

$$\begin{aligned}Y_2[m]&=A\sqrt{2P_0}\left(h_1S_2[m]+h_2[m]\otimes \overset{o}{S}{}_1^*[m-d]\right)\\&\quad+A\left(g_1Z_{12}[m]+g_2[m]\otimes \overset{o}{Z}{}_{21}^*[m-d]\right)+W_2[m]\end{aligned}\quad(16)$$

where sequences $\overset{o}{S}_j[m]$ and $\overset{o}{Z}_{2j}[m]$ are the circular time reversal sequences $S_j[m]$ and $Z_{2j}[m]$, respectively. By taking the N-point DFT of $Y_1[m]$, $Y_2[m]$ sequences and the properties of circular time-reversal1 sequences, for n=0,...,N-1 , one derives

$$\begin{aligned}y_1[n]&=A\sqrt{2P_0}\left(h_1s_1[n]-H_2[n]s_2^*[n]\right)+\tilde{w}_1[n],\\y_2[n]&=A\sqrt{2P_0}\left(h_1s_2[n]+H_2[n]s_1^*[n]\right)+\tilde{w}_2[n],\end{aligned}\quad(17)$$

with

$$\begin{aligned}&H_2[n]=q_2^*G_2[n],\\&G_2[n]=\left(g_{20}+g_{21}e^{-j2\pi d/N}\right),\\&\tilde{w}_1[n]=A\left(g_1z_{11}[n]-G_2[n]z_{22}^*[n]\right)+w_1[n],\\&\tilde{w}_2[n]=A\left(g_1z_{12}[n]+G_2[n]z_{21}^*[n]\right)+w_2[n],\\&z_{11}[n]=\mathrm{DFT}\{Z_{11}[m]\},\ z_{22}[n]=\mathrm{DFT}\{Z_{22}[m]\},\\&z_{12}[n]=\mathrm{DFT}\{Z_{12}[m]\},\ z_{21}[n]=\mathrm{DFT}\{Z_{21}[m]\},\\&w_1[n]=\mathrm{DFT}\{W_1[m]\},\ w_2[n]=\mathrm{DFT}\{W_2[m]\}.\end{aligned}\quad(18)$$

Clearly, $z_{ij}[n]\sim CN(0,N_0)$ and $w_j[n]\sim CN(0,N_0)$ for $i,j=1,2$ .The received signals for the block-index (k), $y[n]^{(k)}=[y_1[n]\quad y_2[n]]^t$, in the matrix form, can be expressed as

$$y[n]^{(k)}=A\sqrt{2P_0}\begin{bmatrix}s_1[n]&-s_2^*[n]\\s_2[n]&s_1^*[n]\end{bmatrix}\begin{bmatrix}h_1\\H_2[n]\end{bmatrix}+\begin{bmatrix}\tilde{w}_1[n]\\\tilde{w}_2[n]\end{bmatrix}\quad(19)$$

It is pointed out that, for given g_1, g_2 the equivalent noise $[\tilde{w}1[n]\quad \tilde{w}2[n]]^t\sim CN(0,\sigma^2[N]I_2)$, where

$$\sigma^2[n]=N_0\left(1+A^2\left(|g_1|^2+|g_2|^2c[n]\right)\right),\quad(20)$$

$$c[n]=\left|p(\tau)+p(T_S-\tau)e^{-j2\pi n/N}\right|^2.\quad(21)$$

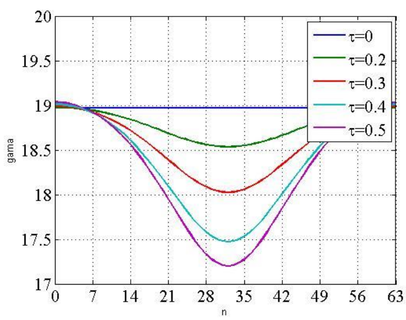

Fig5. Average received SNR vs. n and τ

Also, the received SNR per symbol, for given g_1, g_2 can

$$\gamma[n,\tau]=\frac{A^2P_0\left(|g_1|^2+|g_2|^2c[n]\right)}{N_0\left(1+A^2\left(|g_1|^2+|g_2|^2c[n]\right)\right)}\quad(22)$$

With raised-cosine filter, $p(\tau)=1$ and $p(T_S-\tau)=0$ for $\tau=0$ and hence $c[n]=1$. Thus, the noise variance and the received SNR of the proposed system are the same as that of the conventional D-DSTC for $\tau = 0$. However, for $\tau \neq 0$ the average received SNR is a function of τ and n. To see this dependency, $\gamma[\tau,n]$ is plotted versus n and τ inFig.5,when N=64, L=1, $P/N_0 = 25\text{dB}$, $P_0 = P/2$, $P_r = P/4$, and for simplicity $|g_1|^2 = |g_2|^2 = 1$. As can be seen, $\gamma[\tau,n]$ is symmetric around its minimum at $n = N/2-1$ Also, overall, $\gamma[\tau,n]$ decreases with increasing τ and reaches its minimum value at $\tau = 0.5T_S$. This phenomena yields the same average BER for symmetric values of τ around $0.5T_S$, as will be seen in the simulation results.

$$\hat{v}_1[n],\hat{v}_2[n]=\arg\min_{v}\left\|y[n]^{(k)}-V[n]^{(k)}y[n]^{(k-1)}\right\|, \quad (23)$$

decode 2N data symbols. Because of the orthogonality of V[n], symbols $v_1[n], v_2[n]$ are decoded independently, without any knowledge of CSI or delay. It is easy to see that, due to the structure of Eq.(19), the desired diversity of two is achieved in this system.

2.3 SIMULATION RESULTS

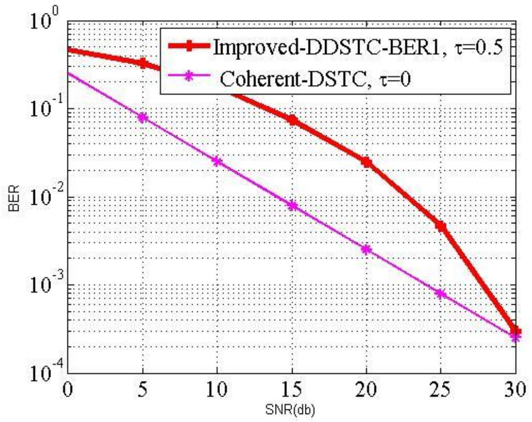

Fig6. Simulation for BER of D-OFDM DSTC (proposed method, $N = 64$, $L = 1$,)and coherent DSTC using BPSK for delays.

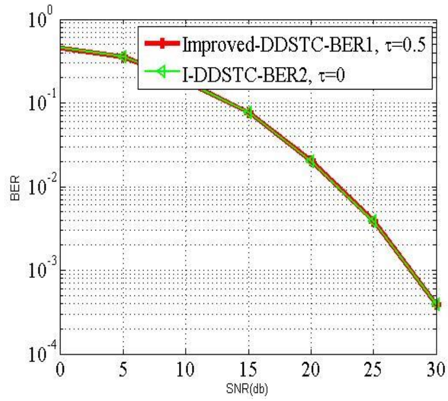

Fig7 Simulation for BER of D-OFDM DSTC (proposed method, $N = 64$, $L = 1$,), D-DSTC , and using BPSK for various values of delays

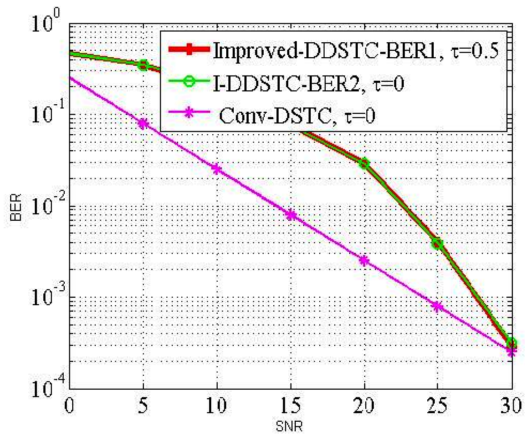

Fig 6: Simulation for BER of D-OFDM DSTC (proposed method, N=64, L=1,), D-DSTC, and coherent DST using BPSK for various values of delays.

As shown in the figure, the performance of the conventional D-DSTC system is severely degraded and an error floor appears in the BER curves. On the other hand, the proposed method is able to deliver the desired performance for all values of the delays. As explained in Section III, the BER curves are symmetric around $\tau = 0.5T_S$.

3. CONCLUSIONS

While collecting channel information is challenging, synchronization error is also inevitable in distributed space-time relay networks. Hence, in this paper a method was proposed that does not require any channel information and is very robust against synchronization error. The method combines differential encoding and decoding with an OFDM-based approach to circumvent channel estimation and deal with synchronization error. It was shown through simulations that the method works well for various synchronization error values.

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BIOGRAPHIES

Ms A.V.H.Leela received her B.Tech degree in Electronics and Communication Engineering from Sarojini Institute of Technology, Telaprolu, in 2014 and pursuing M.Tech in SRK Institute of Technology, Vijayawada, Andhra Pradesh, India. Her Research interests are Wireless Communications.

Mrs. M.Naga Mani received B.Tech Degree in Electronics and Communication Engineering from Sarathi College of Engineering, nuzividu in 2009 and M.Tech Degree in System And Signal Processing from Lakireddy Bali Reddy College Of Engineering, Andhra Pradesh, India in 2011. Now, She is working as a Asst.Professor in S.R.K Institute of Technology, Vijayawada, Andhra Pradesh, India. Her, research interests are analog and digital communications, cellular mobile communication, wireless communications and Networking.