# Distributed Double Differential Space-Time Coding with Amplify-and-Forward Relaying

Sylvia Ong Ai Ling, Hushairi Zen, Al-Khalid Hj Othman and Khairuddin Bin Hamid
*Faculty of Engineering, Universiti Malaysia Sarawak, Kota Samarahan, Kuching, Sarawak, Malaysia.*
*15010028@siswa.unimas.my*

***Abstract*—This paper provides the double differentially modulated distributed space-time coding for amplify-and-forward (AF) relaying cooperative communications system under time-varying fading channels. In many wireless systems, the communication terminals are mobile. In such case, frequency offsets arise subjected to Doppler's effect and frequency mismatch amongst the terminals' local oscillators. The double differential coding is proposed to overcome the problem of frequency offsets that present in the channel due to the rapidly fast moving nodes. The advantage of the double differential is that the scheme requires neither channel nor frequency offset knowledge for decoding process at the desired destination. However, the conventional two-codeword approach fails to perform and leads to error floor, a region where the error probability performance curve flattens for high signal-to-noise ratio (SNR) regime in fast fading environment. Hence, a low complexity multiple-codeword double differential sphere decoding (MCDDSD) is proposed. The simulation results show that the proposed MCDDSD significantly improve the system performance in time-varying environment.**

***Index Terms*—Distributed Space-Time Coding; Double Differential Detection; Frequency Offsets; Sphere Decoding; Time-Varying Relaying Channel**

## I. INTRODUCTION

Recently, cooperative diversity has stimulated much attention in the wireless communication system due to its ability to achieve diversity gain and improve the system performance. Previous studies have primarily assumed that either perfect channel knowledge is known at relay and destination or the channel is over slow fading condition. Generally, channel knowledge estimation acquired the use of pilot symbols or blind detection at the cost of spectral efficiency loss and excessive computational complexity, especially for fast fading channels. Thus, it is particularly important to explore differential schemes that obviate the need for channel estimation at the destination. The differential space-time block code relaying system is widely studied in [1–3]. The differential system works for channels that remain fixed over at least two symbol interval time. However, the system fails to perform in rapidly fading mobile environments due to the presence of frequency offsets. Frequency offsets happen because of the relative motion amongst the nodes and are proportional to the Doppler shift. Another cause of frequency offsets is the mismatch carrier frequency between the terminals' oscillators. To address the problem of frequency offsets, double differential coding regardless of channel estimation is studied.

Several space-time block coded modulation based on double differential and their performance analyses can be found in [4–8]. Liu et al. in [4] and Bhatnagar et al. in [5] implemented double differential coding under time-selective fading channel and flat block-fading correlated Rayleigh channel, respectively in multiple-input multiple-output (MIMO) system. The low complexity scheme enables diversity gain foregoes the knowledge of channel and frequency offsets. However, the proposed scheme in [5] requires partially known channel fading statistics at the source. The double differential for cooperative communication system was further proposed in [6] and [7]. However, the studies only consider relayed communication with no direct link between the source and destination. Furthermore, the model in [7] assumed that oscillators' synchronization is achieved between the source and relay.

Although the above researchers showed many interesting results indicating the potential of the differential based space-time coding to overcome the complex computational channel knowledge and frequency offsets estimation, the network performance employing codeword based detection, however, experiences degradation (i.e. error floor) at high signal-to-noise ratio (SNR) in a relatively fast fading environment. This is the motivation behind the present study. Multiple codeword differential detections (MCDD) has been investigated in [9–11] for unitary space-time codes in multiple-input multiple-output (MIMO) systems and dual-hop cooperative communication network, accordingly to overcome the performance degradation of the two-codeword based detection at the cost of an increased complexity. In general, MCDD processes blocks of received codeword and improve the power efficiency as the block size increases. However, the search space increases exponentially with the block size and the complexity of MCDD becomes highly prohibited when maximum likelihood (ML) decoding is applied. ML decoder is based on full search over the codeword that is continuously transmitted from the source and relays. In order to reduce the exhaustive search of the MCDD detection, Lampe et. al. in [12] developed a sphere decoding algorithm so as to reduce the search area during the decoding process. The sphere decoding algorithm has near ML performance with reasonably low complexity. In the context of relay networks, the optimum decision metric does not yield a closed-form solution due to the complexity of the distributed received signals at the desired destination. Thus, a low-complexity alternative decision rule is proposed to replace the optimum decision rule. This research outperforms the previously suggested schemes and is able to additionally integrate the direct link between the source and destination terminals.

Moreover, the obtained results show that the proposed multiple-codeword sphere decoding (MCSD) based on double differential scheme achieves significant performance improvement as compared with the conventional codeword based detection.

The rest of the paper is presented as follows. Section II describes the system model. Section III provides the details of the proposed MCSD double differential detection for AF distributed space-time block coding under time-varying environment. Next, the simulation results are presented and discussed in Section IV. Finally, the conclusion is drawn.

## II. SYSTEM MODEL

The system model is designed using the following strategy in [13] that consists of a source, two relays and a destination as depicted in Figure 1. The proposed model allows the communicating terminals to use the direct link and relayed link via the relays to achieve a higher diversity gain.

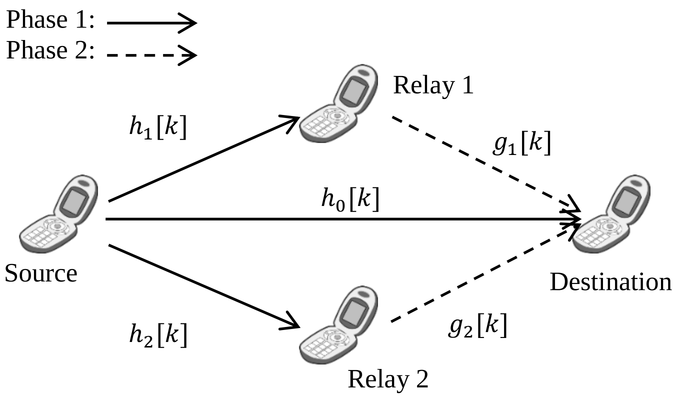


Figure 1: System Model for Distributed Double Differential Space-Time Coding Using AF Relaying Protocol

Considering all the terminals are equipped with a single antenna, the communications between the terminals are in a half-duplex manner. The channel coefficients at time $k$ from source to relay, relay to destination and source to destination are depicted as $h_i[k], g_i[k]$ and $h_0[k]$. The channels of all links are assumed to be Rayleigh flat-fading (i.e. $h_i \sim \mathcal{CN}(0, \sigma_i^2), i = 1,2,3$), spatially uncorrelated and changing over time due to the mobility of the terminals. The channel variation is described by the auto-correlation between two channel coefficients, $n$ blocks apart based on Jake's model [14]:

$$\varphi_{s,r}(n) = E\{h_i[k]h_i^*[k+n]\} = \sigma_i^2 J_0(2\pi f_{s,r} nR) \quad (1)$$
$$\varphi_{r,d}(n) = E\{g_i[k]g_i^*[k+n]\} = \sigma_i^2 J_0(2\pi f_{r,d} nR) \quad (2)$$

where $J_0(\cdot)$ denotes the zeroth-order Bessel function of the first kind, $f_{s,r}$ and $f_{r,d}$ are the maximum normalized Doppler frequency of the source-relay and relay-destination channels, respectively. It is noted that the auto-correlation value equals 1 for fixed channels and decreases in time-varying channels. It is also assumed that all links are perturbed with frequency offsets. Generally, the frequency offsets is caused by the mismatch between the nodes' oscillator and relative motion among the nodes. In this paper, it is consider that the frequency offsets is due to the Doppler effects.

At the source, the information bits $log_2 M$ are converted to symbols using M-PSK as $c[k] \in \mathcal{C}$ where $\mathcal{C}$ is $R \times R$ unitary matrices depending on the number of relays $R$. $\mathcal{C}$ can be written as:

$$\mathcal{C} = \{C_l | C_l C_l^H = C_l^H C_l = I_R, l = 1,2,\dots,L\} \quad (3)$$

where $L$ represents the length of codeword.

Before the transmission process, the symbols are transformed into Alamouti codeword [15] $C[k] \in \mathcal{C}$ and double differentially encoded as follows:

$$s[k] = C[k]s[k-1]s[k-2], \quad s[0] = [1 \;\; 0 \;\; \dots \;\; 0]^T \quad (4)$$

The transmission is divided into two phase. For the transmission process, either codeword-by-codeword or block of codeword can be transmitted. In both cases, the analysis is similar. During the first phase, let $P_S$ as the average power per transmission by the source and $w \in [-\pi, \pi]$ be the frequency offset between the source and destination, which is independent of time $k$. The frequency offset is assumed to remain static over the transmission period of at least three double differential encoded matrices [4],[5]. The received vector at the destination and $i$th relay can be expressed as:

$$y_0[k] = \sqrt{P_S} s[k] e^{j2w[k]} h_0[k] + w_0[k] \quad (5)$$
$$r_i[k] = \sqrt{P_S R} s[k] e^{j2w[k]} h_i[k] + u_i[k] \quad (6)$$

where $w_0[k] \sim \mathcal{CN}(0, N_0 I_0)$ and $u_i[k] \sim \mathcal{CN}(0, N_0 I_R)$ denote the noise vector at the destination and $i$th relay, respectively. At the relay(s), the received vector is linearly combined with its conjugate as:

$$x_i[k] = G(A_i r_i[k] + B_i r_i^*[k]) \quad (7)$$

where $A_i$ and $B_i$ are the combining matrices and determined based on the space-time code. Referring to Alamouti code $A_i$ and $B_i$ is written as:

$$A_1 = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}, \quad B_1 = 0_T$$
$$A_2 = 0_T, \quad B_2 = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix} \quad (8)$$

$G$ represents the amplifier gain at the relay that can either be constant or varies. In the environment when no channel properties and frequency offsets are known, a constant gain is employed using the variance of the source-relay channels as follows:

$$G = \sqrt{\frac{P_R}{P_S \sigma_{s,r}^2 + N_0}} \quad (9)$$

where $P_R$ is the average power per transmission by the relay(s). All the relay transmissions are distributed in the same channel at the same time and superimpose at the destination, thus the corresponding received vector at the destination during second phase of the transmission is:

$$y[k] = \sum_{i=1}^{R} x_i[k]\, g_i[k] + w_i[k] \tag{10}$$

where $w_i[k] \sim \mathcal{CN}(0, N_0 I_R)$ is the noise vector at destination. By substituting (7) into (10) yields:

$$y[k] = G\,\sqrt{P_S R} S[k] e^{j2w[k]} q[k] D[k] + w[k] \tag{11}$$

where:

$$S[k] = \left[\hat{A}_1 \hat{s}_1 \dots \hat{A}_R \hat{s}_R\right] = C[k] S[k-1] S[k-2]$$
$$q[k] = [q_1[k] \dots q_R[k]]^T$$
$$D[k] = diag[1, e^{jwk}, \dots, e^{jwk[N-1]}]$$
$$w[k] = G \sum_{i=1}^{R} g[k] \hat{A}_i\, \hat{u}_i[k] + w_i[k]$$

and

$$\left.\begin{matrix} \hat{A}_i = A_i, & q_i[k] = h_i[k] g_i[k], \\ \hat{u}_i[k] = u_i[k], & \hat{s}_i[k] = s[k] \end{matrix}\right\} \; if \;\; B_i = 0$$

$$\left.\begin{matrix} \hat{A}_i = B_i, & q_i[k] = h_i^*[k] g_i[k], \\ \hat{u}_i[k] = u_i^*[k], & \hat{s}_i[k] = s^*[k] \end{matrix}\right\} \; if \;\; A_i = 0$$

represents the distributed space-time coding matrix, the equivalent relayed channel vector as well as the equivalent noise vector.

## III. Multiple-Codeword Double Differential Detection

Multiple-codeword double differential detection is proposed to circumvent the limitation of two-codeword double differential detection under fast fading environments.

By omitting the notation $k$ from (13) and (14) for simplicity, considering the model under time-varying channel for which $N$ received symbols (in vector form) at the destination from the source and relay is expressed as:

$$\left.\begin{matrix} \hat{A}_i = B_i, & q_i[k] = h_i^*[k] g_i[k], \\ \hat{u}_i[k] = u_i^*[k], & \hat{s}_i[k] = s^*[k] \end{matrix}\right\} \; if \;\; A_i = 0 \tag{12}$$

$$\bar{y}_{s,d} = \sqrt{P_S}\, diag\{\bar{s}\}\, diag\{\bar{h}_{s,d}\} e^{jwk} diag\{D\} + \bar{w}_{s,d} \tag{13}$$

$$\bar{y}_{r,d} = G\sqrt{P_S R}\; diag\{\bar{s}\}\, diag\{\bar{h}_{r,d}\} \bar{h}_{s,r}\, e^{jwk} diag\{D\} + \bar{w}_{r,d} \tag{14}$$

where

$\bar{s} = s[1], s[2], \dots, s[N]]$

$\bar{y}_{r,d} = [{y_{r,d}}^T[1], {y_{r,d}}^T[2], \dots, {y_{r,d}}^T[N]]^T$

$\bar{y}_{s,d} = [{y_{s,d}}^T[1], {y_{r,d}}^T[2], \dots, {y_{r,d}}^T[N]]^T$

$\bar{h}_{r,d} = [h_{r,d}\,[1], h_{r,d}[2], \dots, h_{r,d}[N]]^T$

$\bar{h}_{s,r} = [{h_{s,r}}^T[1], {h_{s,r}}^T[2], \dots, {h_{s,r}}^T[N]]^T$

$\bar{h}_{s,d} = [h_{s,d}\,[1], y_{s,d}[2], \dots, y_{s,d}[N]]^T$

$\bar{w}_{r,d} = [{w_{r,d}}^T[1], {w_{r,d}}^T[2], \dots, {w_{r,d}}^T[N]]^T$

$\bar{w}_{s,d} = [{w_{s,d}}^T[1], {w_{s,d}}^T[2], \dots, {w_{s,d}}^T[N]]^T$

$D = [1, e^{jw}, \dots, e^{jw}[N-1]]$

$s$ is a unitary block diagonal matrix ($s^H s = s s^H = I R_N$) and it contains $N$ transmitted codewords corresponding to $N-2$ data codewords collected in:

$$\bar{C} = diag\{C[1], \dots, C[N-1], C[N-2]\} \tag{15}$$

such that $s[n+2] = C[n]s[n]s[n+1]$ and $s[n] = I_R$ is set as initial symbol.

Therefore, conditioned on $\bar{s}$, $\bar{y}_{s,d}$ is a circularly symmetric complex Gaussian vector with the following pdf:

$$P(\bar{y}_{s,d}|\bar{s}) = \frac{1}{\pi^N \det\{\Sigma_{\bar{y}_{s,d}}\}} \exp\left(-{\bar{y}_{s,d}}^H {\Sigma_{\bar{y}_{s,d}}}^{-1} \bar{y}_{s,d}\right) \tag{16}$$

Similarly, conditioned on $\bar{s}$ and $\bar{h}_{r,d}$, $\bar{y}_{r,d}$ is a circularly symmetric complex Gaussian vector with the following pdf:

$$P(\bar{y}_{r,d}|\bar{s}, \bar{h}_{r,d}) = \frac{1}{\pi^N \det\{\Sigma_{\bar{y}_{r,d}}\}} \exp\left(-{\bar{y}_{r,d}}^H {\Sigma_{\bar{y}_{r,d}}}^{-1} \bar{y}_{r,d}\right) \tag{17}$$

In (16) and (17), the matrices $\Sigma_{\bar{y}_{s,d}}$ and $\Sigma_{\bar{y}_{r,d}}$ are conditional covariance matrices of $\bar{y}_{s,d}$ and $\bar{y}_{r,d}$, defined as:

$$\begin{aligned} \Sigma_{\bar{y}_{s,d}} &= E\{\bar{y}_{s,d}\, {\bar{y}_{s,d}}^H \,|\, s\} \\ &= P_S\, diag\{\bar{s}\} \Sigma_{\bar{h}_{s,d}}\, diag\{\bar{s}^*\} + N_0 I_N \\ &= diag\{\bar{s}\}\left(P_S \Sigma_{\bar{h}_{s,d}} + N_0 I_N\right) diag\{s^*\} \end{aligned} \tag{18}$$

and

$$\begin{aligned} \Sigma_{\bar{y}_{r,d}} &= E\{\bar{y}_{r,d}\, {\bar{y}_{r,d}}^H \,|\, s, \bar{h}_{r,d}\} \\ &= G^2 (P_S R)\, diag\{\bar{s}\}\, diag\{\bar{h}_{r,d}) \Sigma_{\bar{h}_{s,r}} \\ &\times diag\{{\bar{h}_{r,d}}^*) diag\{\bar{s}^*\} + \Sigma_{\bar{w}} \end{aligned} \tag{19}$$

where the covariance matrices of $\bar{h}_{s,d}$, $\bar{h}_{s,r}$ and $\bar{w}$ is defined, accordingly as follows:

$$\Sigma_{\bar{h}_{s,d}} = E\left\{\bar{h}_{s,d}\, {\bar{h}_{s,d}}^H\right\} = C_{\bar{h}_{s,d}} \otimes I_R \tag{20}$$

$$C_{\bar{h}_{s,d}} = toeplitz\{\varphi_{s,d}(0), \dots, \varphi_{s,d}(N-2)\} \tag{21}$$

and

$$\Sigma_{\bar{h}_{s,r}} = E\left\{\bar{h}_{s,r}\, {\bar{h}_{s,r}}^H\right\} = C_{\bar{h}_{s,r}} \otimes I_R \tag{22}$$

$$C_{\bar{h}_{s,r}} = toeplitz\{\varphi_{s,r}(0), \dots, \varphi_{s,r}(N-2)\} \tag{23}$$

and

$$\Sigma_{\bar{w}} = E\{\bar{w}\, \bar{w}^H\} = C_{\bar{w}} \otimes I_R \tag{24}$$

$$C_{\bar{w}} = N_0\, diag\{(1 + G^2 \textstyle\sum_{i=1}^{R} |\bar{h}_{r,d_i}[1]|^2), \dots, (1 + G^2 \sum_{i=1}^{R} |\bar{h}_{r,d_i}[N]|^2)\} \tag{25}$$

Given (16) and (17), the non-coherent ML detection without the channel and frequency offsets estimation can be expressed as:

$$\hat{s} = \arg\max_{s\in M^{N-2}} \{\frac{1}{\pi^N \det\{\Sigma_{\bar{y}_{s,d}}\}} \exp\left(-\bar{y}_{s,d}{}^H \Sigma_{\bar{y}_{s,d}}{}^{-1} \bar{y}_{s,d}\right)\} \times \underset{\bar{h}_{r,d}}{\mathrm{E}} \{\frac{1}{\pi^N \det\{\Sigma_{\bar{y}_{r,d}}\}} \exp\left(-\bar{y}_{r,d}{}^H \Sigma_{\bar{y}_{r,d}}{}^{-1} \bar{y}_{r,d}\right) \tag{26}$$

As can be seen from (15), the ML metric needs the expectation over the distribution of $\bar{h}_{r,d}$, which does not yield a closed-form expression. As an alternative, it is proposed to use the following decision metric:

$$\hat{s} = \arg\max_{s\in M^{N-2}} \{\frac{1}{\pi^N \det\{\Sigma_{\bar{y}_{s,d}}\}} \exp\left(-\bar{y}_{s,d}{}^H \Sigma_{\bar{y}_{s,d}}{}^{-1} \bar{y}_{s,d}\right)\} \times \{\frac{1}{\pi^N \det\{\bar{\Sigma}_{\bar{y}_{r,d}}\}} \exp\left(-\bar{y}_{r,d}{}^H \bar{\Sigma}_{\bar{y}_{r,d}}{}^{-1} \bar{y}_{r,d}\right)\} \tag{27}$$

where

$$\bar{\Sigma}_{\bar{y}_{r,d}} = \underset{\bar{h}_{r,d}}{\mathrm{E}} \{\Sigma_{\bar{y}_{r,d}}\} = G^2 P_S R \bar{s}(C_{\bar{h}} \otimes I_R)\bar{s}^H + (1 + G^2 \sigma_{r,d}^2 R) N_0 (I_N \otimes I_R) = \bar{s}(C \otimes I_R)\bar{s}^H \tag{28}$$

and

$$C = G^2 P_S R \bar{s} C_{\bar{h}} + N_0\left(1 + G^2 \sigma_{r,d}^2 R\right) I_N$$
$$C_{\bar{h}} = toeplitz\{\varphi_{s,r}(0)\varphi_{r,d}(0), \dots, \varphi_{s,r}(N-2)\, \varphi_{r,d}(N-2)\} \tag{29}$$

Using the rule $\det\{AB\} = \det\{BA\}$, the determinant in (27) is no longer dependent to $\bar{s}$ and the modified decision matric can be simplified as:

$$\begin{aligned}\hat{\bar{s}} &= \arg\min_{\bar{s}\in M^{N-2}} \{\bar{y}_{r,d}{}^H\, \Sigma_{\bar{y}_{r,d}}{}^{-1} \bar{y}_{r,d} + \bar{y}_{s,d} \Sigma_{\bar{y}_{s,d}}{}^{-1} \bar{y}_{s,d}\} \\ &= \arg\min_{\bar{s}\in M^{N-2}} \{\bar{y}_{r,d}{}^H\, diag\{\bar{s}\}(C^{-1} \otimes I_R)\, diag\{\bar{s}\}^*\, \bar{y}_{r,d} + \bar{y}_{s,d} diag\{\bar{s}\}\left(C_{s,d}^{-1} \otimes I_R\right) diag\{\bar{s}\}^* \bar{y}_{s,d}\} \\ &= \arg\min_{\bar{s}\in M^{N-2}} \{\bar{y}_{r,d}{}^H\, diag\{\bar{s}\}\left(U_{r,d}^H \otimes I_R\right) \times \left(U_{r,d} \otimes I_R\right) diag\{\bar{s}\}^*\, \bar{y}_{r,d} + \bar{y}_{s,d} diag\{\bar{s}\} \times \left(U_{s,d}^H \otimes I_R\right)\left(U_{s,d} \otimes I_R\right) diag\{\bar{s}\}^* \bar{y}_{s,d}\}\} \\ &= \arg\min_{\bar{s}\in M^{N-2}} \{\|U\|^2 + \left\|U_{s,d}\right\|^2\end{aligned} \tag{30}$$

where $U$ is an upper triangular matrix obtained by the Cholesky decomposition of $C^{-1} = U^H U$ and thus, the new matrices $U_{\bar{s}}$ is defined as:

$$U_{\bar{s}} = \begin{bmatrix} \sum_{j=1}^{N} U_{\bar{s}1,j} \bar{s}^H[j]\, \bar{y}[j] \\ \sum_{j=2}^{N} U_{\bar{s}2,j} \bar{s}^H[j]\, \bar{y}[j] \\ \vdots \\ u_{N,N} \bar{s}^H[N] y[N] \end{bmatrix} \tag{31}$$

where $U_{\bar{s}_{i,j}} = [U_{s,d_{i,j}}, U_{r,d_{i,j}}]^T$ is the specific vector component in row $i$ and column $j$. $U_{s,d_{i,j}}$ and $U_{r,d_{i,j}}$ denote the entry of $U_{s,d}$ and $U_{r,d}$ in row $i$ and column $j$, respectively.

Since $s[N] = I_R$, the last term of vector $U_{\bar{s}}$ does not have any effect on the minimization and can be ignored. By substituting $\bar{s}^H[n] = \bar{s}^H[n+2]\, \bar{s}^H[n+1] V_n$ into (19), it follows that:

$$\hat{\bar{V}} = \arg\min_{\hat{s}\in M^N} \{ \sum_{n=1}^{N-2} \left\| u_{n,n} V[n]\, \bar{y}[n] + \bar{s}[n+1] \sum_{j=n+1}^{N-1} u_{n,j} \bar{s}^H[j] \bar{y}[j] + \bar{s}[n+2] \sum_{j=n+2}^{N} u_{n,j} \bar{s}^H[j] \bar{y}[j] \right\|^2 \} \tag{32}$$

Then, the minimization can be computed using the sphere decoding algorithm to obtain $N-2$ codewords with low complexity.

## IV. Simulation Results

In this section, the double differential AF relay network simulation is performed under symmetric channel qualities (i.e. when the channel variance for all channels equal to unity) employing both two-codeword and the proposed multiple-codeword for distributed space-time block coding. The channel qualities are based on the fading powers of the channels. Thus, the total power $P_T$ is divided equally between the source and relay terminals, with $\frac{P_S}{2}$ and $\frac{P_R}{2R}$, respectively where $R$ represents the number of relays. The simulation results are obtained from $10^5$ realization. Information data is double differentially modulated with BPSK and QPSK constellations. In the simulation, all the uncorrelated Rayleigh faded channels are generated using the Sum-of Sinusoids (SoS) [16].

### *(a) Performance of Two-Codeword and Multiple-Codeword in Relayed Transmission and Relayed with Direct Transmission*

To motivate the design of multiple-codeword over two-codeword, the performance of MCDDSD is evaluated under two scenarios. The first scenario involves the dual-hop (i.e. source-relay-destination without a direct link) while the second scenario considers both dual-hop and direct (i.e. source-destination) link. The simulation is carried out over a time-varying Rayleigh fading channel, for which the normalized frequency at each link is $f_{s,r} = 0.05; f_{r,d} = 0.04; f_{s,d} = 0.04$. The channels are assumed to be constant for three consecutive blocks. It can be observed in Figure 2 that the BER performance of both two-codeword and multiple-codeword in the two scenarios decreases with respect to the P. This shows that cooperative diversity is achieved for the distributed space time coding. Higher diversity order is obtained when direct link is included in the relayed network, which outperforms the performance in the scenario of a dual-hop link network.

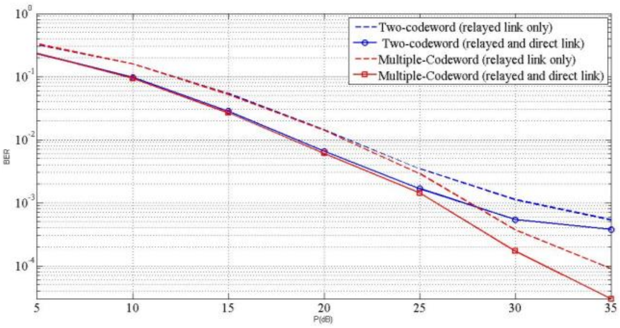


Figure 2: BER performance comparison of the B for the two-codeword and proposed multiple-codeword based on double differential transmission when $f_{s,r} = 0.06; f_{r,d} = 0.04; f_{s,d} = 0.04$

*(b) Performance of Two-Codeword and Multiple-Codeword in Cooperative Transmission Under Time-Varying Channel Environment*

The proposed scheme is further implemented and simulated in various channels fading scenario. The results are depicted in Figure 3. In case I, it is considered that the source and destination are moving in slow velocity (i.e. $f_{s,r} = f_{r,d} = f_{s,d} = 0.001$). The normalized frequency of each channel show that all the channels are under slow fading. In case II, both source and destination are moving with fairly fast fading channels (i.e. $f_{s,r} = 0.05$; $f_{r,d} = 0.05$; $f_{s,d} = 0.05$). In case III, the source and destination is considered to have high mobility with fast fading channels (i.e. $f_{s,r} = 0.01$; $f_{r,d} = 0.01$; $f_{s,d} = 0.01$). It can be seen from Figure 3 that for two-codeword and scheme in case I, the BER is monotonically decreasing when SNR increases. However, in case II, the curve shows deviation from the results in case I at around 20dB and experiences error floor at 35dB. In case III, the performance degrades more severe as compared to case II. The degradation starts earlier at around 15 dB and reaches error floor at 30dB.

Due to the poor performance of the two-codeword scheme in both case II and III, multiple-codeword sphere decoding based on double differential modulation with $N = 10$ is proposed and simulated to case II and III for BPSK and QPSK constellations. The simulation results are plotted in Figure 3 and Figure 4. It can be seen that the MCDDSD scheme significantly outperform the two-codeword in both case II and III.

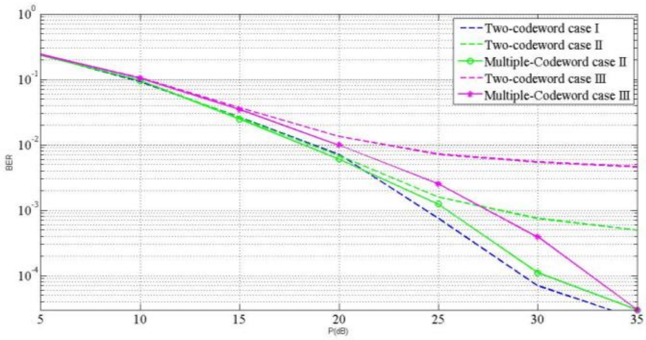


Figure 3: Performance comparison of two-codeword and MCDDSD under different fading channels for BPSK constellations.

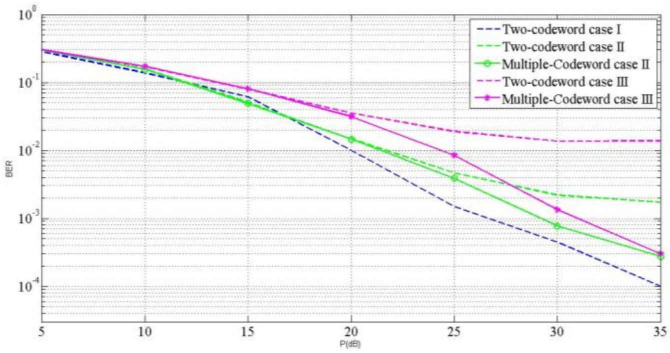


Figure 4: Performance comparison of two-codeword and MCDDSD under different fading channels for QPSK constellations.

As shown in Figure 3 and 4, the BER curve of case I show the best performance achieved in both system employing BPSK and QPSK constellations since it is simulated in a slow fading environment. Hence, the plot is served as a benchmark in order to show the effectiveness of the proposed method in the time-varying channels. From the plots, it can be observed that the proposed method is able to reduce the error floors in fast fading environment and simultaneously increase the performance by bringing the curve closer to that of case I.

## V. CONCLUSION

This paper presented the performance enhancement of a relayed communication link which includes a direct link without requiring the channel and frequency offsets estimation. The simulation results also show that the two-codeword transmission for double differential distributed space-time coding in wireless cooperative diversity networks performs poorly under fast fading environment. A suboptimal multiple-codeword double differential detection is designed and implemented into the network. Since the complexity of the designed detection scheme increases with $N$, sphere decoding is integrated to reduce its complexity. Simulation results show that the proposed scheme significantly improve the poor performance of two-codeword scheme for fast fading environments.

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