Detection Schemes for Distributed Space-Time Block Coding in Time-Varying Wireless Cooperative Systems

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Abstract— **In this paper, we investigate the influence of time-varying channels on the bit-error-rate (BER) performance for distributed space-time block coded system. We make use of pilot symbol assisted modulation (PSAM) to estimate the time-varying channel coefficients. We assess the performance of different reception methods, one of which is our newly considered receiver called cooperative maximum likelihood detector. The others are maximum likelihood detection in [12], Alamouti's receiver, zero-forcing detection and decision-feedback detection. Our results show that all detection methods, except cooperative maximum likelihood detection, achieve nearly the same BER performance over time-varying channels, unlike the results obtained in [12] by assuming perfect channel estimation. In many cases, cooperative maximum likelihood detection performs better by about 5dB due to the diversity gain. In addition, we have found that the time-varying nature contributes to the error flooring effect.**

Keywords-**Cooperative communications, space-time block coding, fading channels, time-varying.**

I. Introduction

To mitigate the fading in wireless channels, spatial diversity offers significant improvement in link reliability and spectral efficiency through the use of multiple antennas at the transmitter and/or receiver side [1]-[4]. Yet, employing a large antenna array might not be practical at the mobile terminals, e.g. cellular mobile devices, due to the size and power limitations of them.

Recently, cooperative diversity has been demonstrated to provide an effective way of improving spectral and power efficiency of wireless networks without the additional complexity of multiple antennas [5]-[8]. In particular, distributed space-time block coding (DSTBC) is of great interest because the conventional version of orthogonal STBC can be readily applied in a distributed fashion [7]. In [9], the authors analyze DSTBC operating in the amplify-and-forward (AF) mode, and show that the original design criteria for conventional STBC still apply to the design of DSTBC schemes.

In [11], the effect of imperfect channel estimation for a DSTBC is studied. However, most of the current work still assumes that the channels are static. In practice, the channels are time-varying, which leads to performance degradation and complexity of the receiver design. In [12], some performance analysis of transmit diversity over time-varying channels has been discussed, yet under the assumption of perfect channel estimation over a fast time-varying environment, which is not realistic. Thus, in this paper, we study different reception schemes with the help of pilot symbol assisted modulation (PSAM). With these estimated channel coefficients, we can then compare their BERs. To the best of our understanding, PSAM has not been studied before in the context of a user cooperation scenario with time-varying channels. In Section II, we will present the time-varying system model. PSAM will be introduced in Section III. Then, we will analyze different receivers in Section IV. Numerical results will be presented in Section V. We then draw our conclusions in Section VI.

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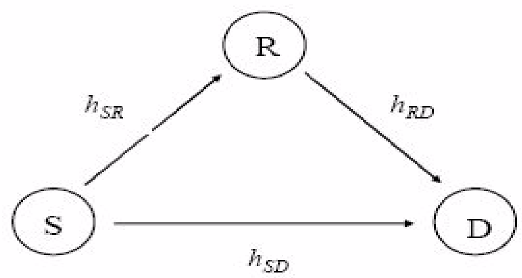

Fig. 1. Schematic representation of relay-assisted transmission.

II. System Model

A wireless communication scenario where the source terminal S transmits information to the destination terminal D with the help of the relay terminal R is considered (see Fig. 1). All terminals are equipped with a single antenna. We assume a frequency-flat Rayleigh fading channel and adopt the user cooperation protocol proposed by [9]: S communicates with R during the first signaling interval. In the second signaling interval, both R and S communicate with D. For the $R \to D$ link[1], an AF mode is used, in which R amplifies and forwards the signal received from S in the first signaling interval. Following [10], we can write the received signal in one period as

$$r = \alpha h_1 x_1 + \beta h_2 x_2 + n, \tag{1}$$

where $h_1 = h_{SR}h_{RD}$ and $h_2 = h_{SD}$. Here, h_{SR}, h_{RD} and h_{SD} denote the fading coefficients for the $S \to R$, $R \to D$ and $S \to D$ links respectively. They are modeled as independent zero-mean complex Gaussian random variables with unit variance, leading to a Rayleigh fading channel. n is a zero mean, complex Gaussian random variable with variance $N_0/2$ per dimension.

[1] $X \to Y$ denotes the link from terminal X to terminal Y

In (1), α and β are defined as

$$\alpha = \sqrt{\frac{E_{SR}/N_0}{1 + E_{SR}/N_0 + E_{RD}/N_0}}\sqrt{E_{RD}}, \tag{2}$$

$$\beta = \sqrt{\frac{1 + E_{SR}/N_0}{1 + E_{SR}/N_0 + E_{RD}/N_0}}\sqrt{E_{SD}}. \tag{3}$$

respectively. We now introduce space-time coding across the transmitted signals, i.e. x_1 and x_2. In our case, we need to use STBC designed for two transmit antennas (i.e. Alamouti's scheme) where the code matrix is defined as

$$\begin{bmatrix} x_1 & x_2 \\ -x_2^* & x_1^* \end{bmatrix}. \tag{4}$$

To model the time-varying channels, let $\tilde{h}_{SR}$, $\tilde{h}_{RD}$ and $\tilde{h}_{SD}$ denote the fading coefficients of $S \to R$, $R \to D$ and $S \to D$ links respectively during the second block period. Thus, the received signals are now given as

$$r_1 = \alpha h_1 x_1 + \beta h_2 x_2 + n_1, \tag{5}$$

$$r_2 = -\alpha \tilde{h}_1 x_2^* + \beta \tilde{h}_2 x_1^* + n_2 \tag{6}$$

where $\tilde{h}_1 = \tilde{h}_{SR}\tilde{h}_{RD}$ and $\tilde{h}_2 = \tilde{h}_{SD}$. In matrix form,

$$\begin{bmatrix} r_1 \\ r_2^* \end{bmatrix} = \begin{bmatrix} \alpha h_1 & \beta h_2 \\ \beta \tilde{h}_2^* & -\alpha \tilde{h}_1^* \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \end{bmatrix} + \begin{bmatrix} n_1 \\ n_2^* \end{bmatrix} \tag{7}$$

or

$$\mathbf{r} = \mathbf{H}\mathbf{x} + \mathbf{n}. \tag{8}$$

We employ Jakes' time-varying model [13] for the above system and let ρ be the correlation between two successive channel realizations in two symbol intervals. Then,

$$\rho = J_0\left(2\pi f_D T\right) \tag{9}$$

where $J_0(.)$ is the zero-order Bessel function of the first kind, T is symbol interval and $f_D = \nu f_c/c$ is the maximum Doppler frequency, where ν is the vehicle speed in ms^{-1}. f_c is the carrier frequency, and c is the speed of light ($3 \times 10^8 ms^{-1}$). Throughout this paper, we make the following assumptions:

- The pairs $(h_i, \tilde{h}_i)$ are independent for $i = 1, 2$.
- Temporally symmetric Rayleigh fading so that the correlation ρ between h_i and $\tilde{h}_i$ is the same for $i = 1, 2$, i.e. $E\left[h_i \tilde{h}_i^*\right] = \rho$.

In practice, as we are not able to obtain the perfect channel state information over time-varying channels, PSAM is the common way to estimate the time-varying channel coefficients.

III. Pilot Symbol Assisted Modulation (PSAM) for DSTBC

In most wireless applications, pilot symbols are inserted in the data frames for practical implementation of channel estimation. PSAM [14] achieves coherent demodulation in a fading environment by using pilot symbols to estimate the channel on a minimum-mean-squared-error (MMSE) basis.

In our PSAM scenario, the pilot symbols are sent separately over different links at different time slots. Specifically, in time slot k, the source terminal S broadcasts the pilot symbol to the destination terminal D and the relay terminal R. In the time slot $k+1$, R transmits the received version of pilot symbol (after energy normalization) to D. There is no transmission from S to D within this period. These pilot symbols are placed at the front end, followed by $M-1$ data symbols in a data frame, where M is the size of data frame. At the receiver, D then extracts the L nearest pilot symbols in data frames and constructs two different sets that represent the pilot symbols received over the $S \to R \to D$ and $S \to D$ links. Subsequently, for each set of the received pilot symbols, the channel estimator gives interpolation by incorporating a Wiener filter among the samples to construct a fading estimate for every symbol period.

Let p represent the pilot symbol, where $p^2 = 1$. The received pilot signals at the destination terminal are given as

$$r_{S\to D}^{k} = \beta h_2^k p + n^k \tag{10}$$

$$r_{S\to R\to D}^{k+1} = \alpha h_1^k p + n^{k+1} \tag{11}$$

where n^k and n^{k+1} are independent samples of complex Gaussian random variables with zero mean and variance of $N_0/2$ per dimension. Based on the received signals corresponding to pilot symbol transmissions, the destination terminal employs the Wiener filter to estimate the fading coefficients. As depicted in Fig. 2, $\lfloor L/2 \rfloor$ pilot symbols from the following frames and $\lfloor (L-1)/2 \rfloor$ pilot symbols from the previous and current frames are employed in this estimation. Defining

$$\bar{r}_{S\to D}^{k} = p \cdot r_{S\to D}^{k} \tag{12}$$

$$\bar{r}_{S\to R\to D}^{k+1} = p \cdot r_{S\to R\to D}^{k+1}, \tag{13}$$

and furthermore introducing the received signal vectors representing the L nearest received pilot symbols, i.e.

$$\bar{\mathbf{r}}_{S\to D} = [\bar{r}_{S\to D}^{-\lfloor (L-1)/2 \rfloor} \cdots \bar{r}_{S\to D}^{\lfloor L/2 \rfloor}] \tag{14}$$

$$\bar{\mathbf{r}}_{S\to R\to D} = [\bar{r}_{S\to R\to D}^{-\lfloor (L-1)/2 \rfloor} \cdots \bar{r}_{S\to R\to D}^{\lfloor L/2 \rfloor}], \tag{15}$$

by using the Wiener filter, say for $S \to D$ link,

$$\mathbf{w}_{S\to D}(m) = [w_{S\to D}^{-\lfloor (L-1)/2 \rfloor}(m) \cdots w_{S\to D}^{\lfloor L/2 \rfloor}(m)] \tag{16}$$

and the corresponding fading estimate at the m^{th} symbol period is denoted by

$$\hat{h}_2(m) = \mathbf{w}_{S\to D}(m)\bar{\mathbf{r}}_{S\to D}. \tag{17}$$

The Wiener filter, which is regarded as the optimal interpolator, in the sense of MMSE, is the one that minimizes

$$\varepsilon^2(m) = \frac{1}{2}E\left[|h_2(m) - \hat{h}_2(m)|^2\right]. \tag{18}$$

It can be obtained by solving

$$\frac{\partial}{\partial w_{S\to D}^{n}(m)}\varepsilon^2(m) = 0, \text{ for } n = -\lfloor (L-1)/2 \rfloor, \ldots, \lfloor L/2 \rfloor, \tag{19}$$

leading to

$$\mathbf{w}_{S\to D}(m) = \mathbf{\Phi_{h_2\bar{r}_{S\to D}}(m)\Phi^{-1}_{\bar{r}_{S\to D}\bar{r}_{S\to D}}}. \tag{20}$$

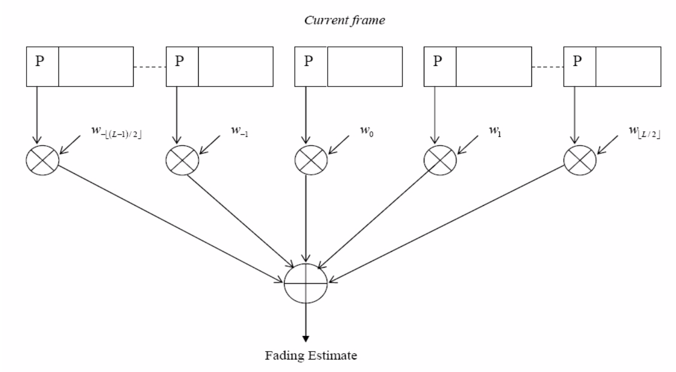

Fig. 2. Fading interpolation in PSAM.

So, the channel estimate for the m^{th} bit is

$$\hat{h}_2(m) = \mathbf{\Phi_{h_2\bar{r}_{S\to D}}(m)\Phi^{-1}_{\bar{r}_{S\to D}\bar{r}_{S\to D}}\bar{r}_{S\to D}}. \tag{21}$$

Similarly, the channel estimate $\hat{h}_1(m)$ for $S \to R \to D$ link is given by

$$\hat{h}_1(m) = \mathbf{\Phi_{h_1\bar{r}_{S\to R\to D}}(m)\Phi^{-1}_{\bar{r}_{S\to R\to D}\bar{r}_{S\to R\to D}}\bar{r}_{S\to R\to D}}. \tag{22}$$

With these channel estimates, we can then construct different receivers and analysis their performances in terms of BER.

IV. Reception Methods

In this section, we examine five reception methods for signals detection. They are the maximum-likelihood receiver in [12], the cooperative maximum-likelihood receiver, Alamouti's receiver, the zero-forcing receiver, the decision-feedback receiver, assuming binary phase shift keying (BPSK). Here, we denote $g_1, g_2, \tilde{g}_1$ and $\tilde{g}_2$ as the estimated values of $h_1, h_2, \tilde{h}_1$ and $\tilde{h}_2$ respectively.

A. Maximum-Likelihood Detection (ML) in [12]

In the presence of AWGN, the maximum-likelihood (ML) detection is equivalent to

$$\arg\min_{\mathbf{x}} ||\mathbf{r} - \hat{\mathbf{H}}\mathbf{x}||^{\mathbf{2}} \tag{23}$$

where

$$\hat{\mathbf{H}} = \begin{bmatrix} \alpha g_1 & \beta g_2 \\ \beta\tilde{g}_2^* & -\alpha\tilde{g}_1^* \end{bmatrix}. \tag{24}$$

Let $\mathbf{R} = \hat{\mathbf{H}}^*\hat{\mathbf{H}}$, where

$$\mathbf{R} = \begin{bmatrix} \alpha^2|g_1|^2 + \beta^2|\tilde{g}_2|^2 & \alpha\beta g_1^* g_2 - \alpha\beta\tilde{g}_1^*\tilde{g}_2 \\ \alpha\beta g_1 g_2^* - \alpha\beta\tilde{g}_1\tilde{g}_2^* & \alpha^2|\tilde{g}_1|^2 + \beta^2|g_2|^2 \end{bmatrix}. \tag{25}$$

Now that $\mathbf{R}$ is Hermitian, we can express $\mathbf{R} = \mathbf{G}^*\mathbf{G}$ by the Cholesky factorization, where G is given by

$$\mathbf{G} = \frac{1}{\sqrt{\alpha^2|\tilde{g}_1|^2+\beta^2|g_2|^2}} \cdot \begin{bmatrix} |\alpha^2 g_1\tilde{g}_1^* + \beta^2 g_2\tilde{g}_2^*| & 0 \\ \alpha\beta g_1 g_2^* - \alpha\beta\tilde{g}_1\tilde{g}_2^* & \alpha^2|\tilde{g}_1|^2 + \beta^2|g_2|^2 \end{bmatrix}. \tag{26}$$

Then, we can convert (23) into

$$\arg\min_{\mathbf{x}} \ |\mathbf{z} - \mathbf{G}\mathbf{x}||^{\mathbf{2}} \tag{27}$$

where

$$\mathbf{z} = \mathbf{G}^{-*}\hat{\mathbf{H}}^*\mathbf{r} \tag{28}$$

by multiplying (23) by the unitary matrix $\mathbf{G}^{-*}\hat{\mathbf{H}}^*$. Multiplying both sides in (8) by $\mathbf{G}^{-*}\hat{\mathbf{H}}^*$ gives the following:

$$\mathbf{z} = \mathbf{G}\mathbf{x} + \mathbf{w} \tag{29}$$

where $z = [z_1, z_2]^T$ and $\mathbf{w}$, which is the white Gaussian noise, has the same statistical properties as $\mathbf{n}$.

B. Cooperative Maximum-Likelihood Detection (CML)

In our transmission protocol of interest, in the first signaling period, if the destination D is idle, it may receive some portions of the signal sent from the source S. Thus, the received signals can be given by

$$r_{11} = \sqrt{E_{SD}}h_2 x_1 + n_{11}, \tag{30}$$

$$r_{12} = \alpha h_1 x_1 + \beta h_2 x_2 + n_{12}, \tag{31}$$

$$r_{21} = -\sqrt{E_{SD}}\tilde{h}_2 x_2^* + n_{21}, \tag{32}$$

$$r_{22} = -\alpha\tilde{h}_1 x_2^* + \beta\tilde{h}_2 x_1^* + n_{22} \tag{33}$$

and the cooperative maximum-likelihood (CML) detection is as follows

$$\arg\min_{\mathbf{x}} ||\mathbf{r} - \hat{\mathbf{P}}\mathbf{x}||^{\mathbf{2}} \tag{34}$$

where

$$\hat{\mathbf{P}} = \begin{bmatrix} \sqrt{E_{SD}}g_2 & 0 \\ \alpha g_1 & \beta g_2 \\ 0 & -\sqrt{E_{SD}}\tilde{g}_2^* \\ \beta\tilde{g}_2^* & -\alpha\tilde{g}_1^* \end{bmatrix}. \tag{35}$$

This time, all the signals sent either from the source or from the relay or both can be taken into account for demodulation at the destination. In order words, we make use of all available information. And, the performance of CML detection is expected to be better than that of ML detection due to the diversity gain if not all the channels are fully correlated. For the sake of simplicity, we ignore the power gains and consider the following code matrix

$$\mathcal{C} = \begin{bmatrix} 0 & x_1 & 0 & 0 \\ x_1 & x_2 & 0 & 0 \\ 0 & 0 & 0 & -x_2^* \\ 0 & 0 & -x_2^* & x_1^* \end{bmatrix} \tag{36}$$

in

$$\begin{bmatrix} r_{11} \\ r_{12} \\ r_{21} \\ r_{22} \end{bmatrix} = \begin{bmatrix} 0 & x_1 & 0 & 0 \\ x_1 & x_2 & 0 & 0 \\ 0 & 0 & 0 & -x_2^* \\ 0 & 0 & -x_2^* & x_1^* \end{bmatrix} \begin{bmatrix} h_1 \\ h_2 \\ \tilde{h}_1 \\ \tilde{h}_2 \end{bmatrix} + \begin{bmatrix} n_{11} \\ n_{12} \\ n_{21} \\ n_{22} \end{bmatrix}, \tag{37}$$

If all channels are independent, the above code matrix $\mathcal{C}$ can achieve third order diversity as its rank equals three. In order to maximize the diversity benefit or in other words, to obtain the quasi-independent channels, we separate the transmission stream into two parts. At the receiver, the signals are recombined before entering the CML detector.

C. Alamouti's Receiver (AR)

The detection scheme used to combine and decode the symbols becomes

$$\hat{x}_1 = sign(g_1^* r_1 + g_2 r_2^*) \hat{x}_2 = sign(\tilde{g}_2^* r_1 - \tilde{g}_1 r_2^*) \quad (38)$$

As it simply does nothing with the received signals to suppress the noise, it is expected to be the worst reception scheme.

D. Zero-forcing Linear Detection (ZF)

This detection tries to force the crosstalk to zero

$$\mathbf{y} = \mathbf{Cr} \quad (39)$$

A linear operation $\mathbf{C}$ tries to make a decision on x_i based on y_i, for $i = 1, 2$. That means, $\mathbf{C}\hat{\mathbf{H}}$ is a nonnegative diagonal matrix or $\mathbf{C} = \mathbf{A}\hat{\mathbf{H}}^{-1}$ for some nonnegative diagonal matrix $\mathbf{A}$. To keep the noise variance the same, it is easy to show that

$$\mathbf{C} = \frac{|\alpha^2 g_1 \tilde{g}_1^* + \beta^2 g_2 \tilde{g}_2^*|}{\alpha^2 g_1 \tilde{g}_1^* + \beta^2 g_2 \tilde{g}_2^*} \mathbf{M} \begin{bmatrix} \alpha \tilde{g}_1^* & \beta g_2 \\ \beta \tilde{g}_2^* & -\alpha g_1 \end{bmatrix} \quad (40)$$

where

$$\mathbf{M} = \begin{bmatrix} \sqrt{\alpha^2 |\tilde{g}_1|^2 + \beta^2 |g_2|^2} & 0 \\ 0 & \sqrt{\alpha^2 |g_1|^2 + \beta^2 |\tilde{g}_2|^2} \end{bmatrix} \quad (41)$$

and

$$\mathbf{A} = |\alpha^2 g_1 \tilde{g}_1^* + \beta^2 g_2 \tilde{g}_2^*| \, \mathbf{M}. \quad (42)$$

Hence, it is easily verified that $\mathbf{A}\mathbf{R}^{-1}\mathbf{A}$ has ones on the diagonal so that the noise variance does not change. Substituting (40) and (8) into (39) gives

$$\mathbf{y} = \mathbf{Ax} + \tilde{\mathbf{n}} \quad (43)$$

where $\tilde{\mathbf{n}} = \mathbf{A}\mathbf{H}^{-1}\mathbf{w}$, $\tilde{n}_1$ and $\tilde{n}_2$ are zero-mean complex Gaussian with variance $N_0/2$ per dimension. The suboptimal decision on $\hat{x}_i$ regarding to x_i is obtained by quantizing y_i only, for $i = 1, 2$.

E. Decision-feedback Detection (DF)

From (26), there is only one symbol x_1 contributing to the first received signal z_1 in (29). By quantizing z_1, we can obtain a suboptimal decision $\hat{x}_1$. We assume this detection is correct and then subtract it off in z_2. Then, we can quantize the resulting difference, denoted by D, to obtain $\hat{x}_2$, where

$$D = z_2 - \frac{\alpha\beta \left(g_1 g_2^* - \tilde{g}_1 \tilde{g}_2^* \right)}{\sqrt{\alpha^2 |\tilde{g}_1|^2 + \beta^2 |g_2|^2}} \hat{x}_1 \quad (44)$$

In other words, the decoded symbols are

$$\hat{x}_1 = sign\left(z_1\right) \hat{x}_2 = sign\left(D\right) \quad (45)$$

For ideal but unrealistic cases, we simply substitute $h_1, h_2, \tilde{h}_1$ and $\tilde{h}_2$ into $g_1, g_2, \tilde{g}_1$ and $\tilde{g}_2$ respectively.

V. Numerical Results

In this section, we consider the performance of different receivers over time-varying channels with PSAM. We choose the frame size $M = 7$ and the number of interpolation coefficients $L = 5$. For CML detection, we separate the transmitted signals into two parts so as to increase the independence of the channels while for the other schemes, the transmitted signals are closely packed as sparse packing does not improve the BER performance. For $E_{SR}/N_0 = 30$dB, the balanced links, i.e., $E_{SD} = E_{RD}$, and fading $R \to D$ link, the simulation results for different settings are plotted. In Fig. 3, for $\rho = 0.99$ $(f_D T = 0.03)$, the results are shown. We observe that, mainly ascribed to diversity gain, CML detection prevails over the others by about 4dB. On the other hand, the performances of the other reception methods do not differ from each other. Fig. 4 shows that for $\rho = 0.9755$ $(f_D T = 0.05)$, again, ZF, AR, DF and ML have similar performance while CML detection still outperforms the others by about 5dB as its diversity gain can compensate the effect of channel estimation error to a certain extent. With practical consideration in time-varying channels, by employing PSAM, most detectors have the similar performance, which contradicts the results obtained in [12] by assuming perfect channel state information available at the receiver. For $\rho = 0.9037$ $(f_D T = 0.1)$, the results become worse due to the accumulated errors in channel estimation over fast time-varying channels, which are plotted in Fig. 5. We see an immediate error floor, though the CML receiver has a diversity benefit. We see that the time-varying channels may greatly reduce the performance. In such a fast time-varying environment, exploiting CML detection will help maintain reliable communications to a certain extent.

VI. Conclusion

To sum up, we have employed a realistic approach to estimate the time-varying channel coefficients by PSAM. With these estimates, we can then assess the BER of different reception methods (ML. CML, AR, ZF and DF) by simulations. Our results have shown that all detection methods, except CML detection, achieve nearly the same BER performance over time-varying channels. In many cases, CML detection performs better by about 5dB due to the diversity gain. However, the performance degrades when the channel becomes less correlated. In addition, we have found that the influence of time-varying channels results in the error flooring effect.

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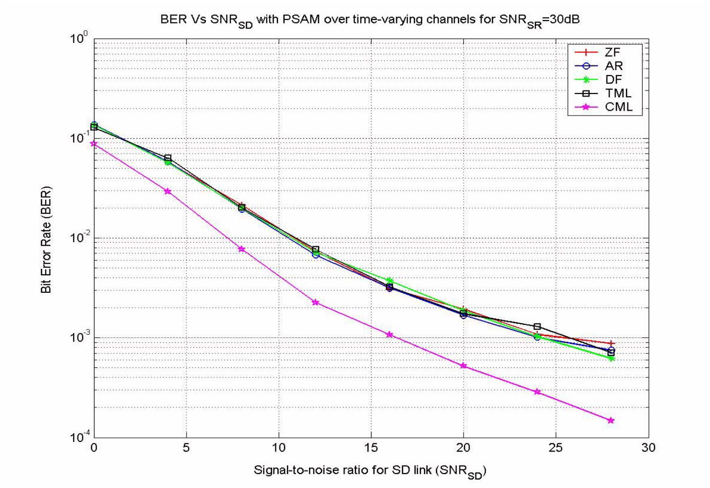

Fig. 3. BER of different detectors with PSAM over time-varying channels with $\rho = 0.99$ when $E_{SR}/N_0 = 30$dB.

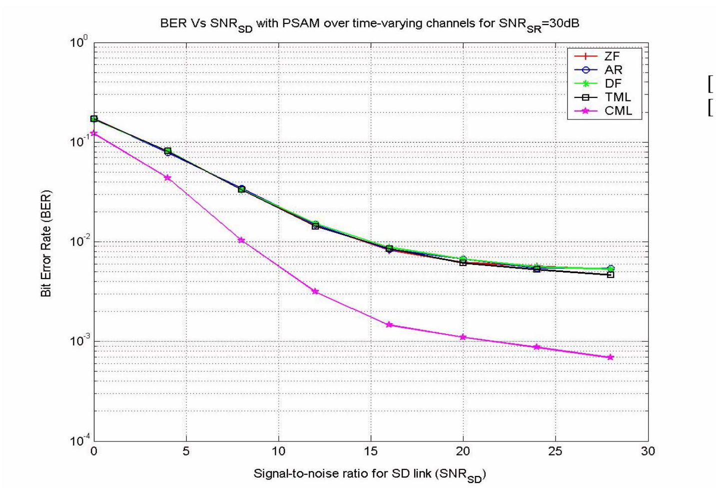

Fig. 4. BER of different detectors with PSAM over time-varying channels with $\rho = 0.9755$ when $E_{SR}/N_0 = 30$dB.

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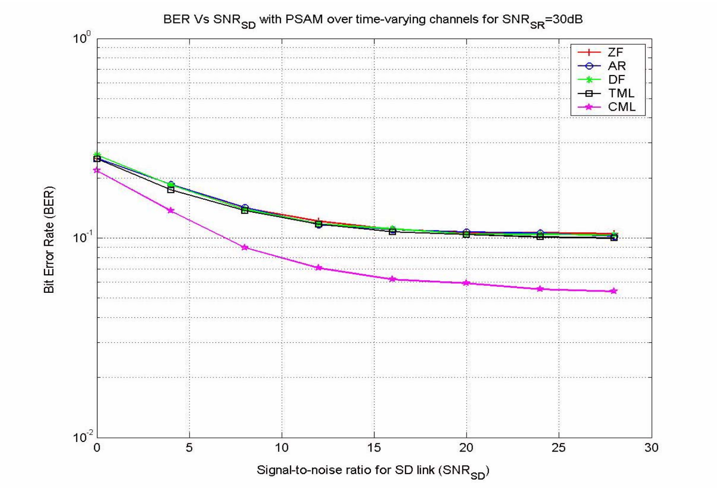

Fig. 5. BER of different detectors with PSAM over time-varying channels with $\rho = 0.9037$ when $E_{SR}/N_0 = 30$dB.

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