Performance Analysis of Space Time Frequency Coding For Mimo Ofdm System

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Abstract

This paper proposes Pre-DFT processing for MIMO OFDM with space time frequency coding. For the application such as high speed computer network, multimedia services and the demand of flexible drive requires a necessity of the broadband wireless communication systems thus the available frequency spectrum is insufficient for the future systems. To accomplish the demand, spectral efficiency must be enhanced in order to increases link throughput and network capacity. A promising approach is to use multiple antennas at both transmitter and receiver i.e., a Multiple-Input Multiple-Output (MIMO) system. Subcarrier based space processing was conventionally employed in Orthogonal Frequency Division Multiplexing (OFDM) systems under Multiple-Input and Multiple Output (MIMO) channel to obtain a optimal performance. At the receiver multiple Discrete Fourier Transform (DFT) blocks in which each of them is corresponding to one receiver antenna are used. In this paper, we present a pre-DFT processing scheme at the receiver of MIMO-OFDM systems embedded with space-time frequency coding. In the proposed system the number of DFT blocks at the receiver will be any number of the receiving antennas which enables effective complexity and performance trade off. The number of the input signal at the space-time frequency decoder can be reduced by using the pre-DFT processing. Therefore a high dimensional MIMO system can be compressed into an equivalent low dimensional. Due to the reduction of the dimension both the complexity of channel and the complexity of the decoder estimation can be reduced. The weighting coefficients estimation for the pre-DFT processing should be relevant to the specific space-time-frequency code. In this paper we propose a simple universal weighting coefficient calculation algorithm that can be used to obtain a excellent performance for most practical space-time-frequency coding. It makes the design of the pre-DFT processing scheme independent of the optimization of the space-time-frequency coding, which is appropriate for multiplatform systems

Keyword: Orthogonal frequency division multiplexes (OFDM), Multiple-input and multiple outputs (MIMO), Space time frequency code.

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1. Introduction

For high data rate wideband wireless communications Orthogonal Frequency division multiplexing (OFDM) is embedded with Multiple-Input Multiple-Output (MIMO) technology to achieve better performance. In the conventional MIMO-OFDM systems, subcarrier based space processing was embedded to obtain optimal performance [1]-[9].however, it requires multiple discrete Fourier transform/ inverse DFT (DFT/IDFT) blocks, each corresponding to one receiver and one transmit antenna.DFT/IDFT can be efficiently implemented using fast Fourier transform/inverse FFT (FFT/IFFT),but the complexity will be higher in OFDM implementation [10].MIMO is one of the most promising technologies to improve the performance in the wireless link.MIMO refers to a radio link with multiple transmitter and receiver antennas. In wireless link, from the transmitter radio signal travels in space, reflected by object and reaches to receiving

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antenna over the multiple paths. While travelling through multiple path causes interference and signal fading occurs.MIMO has an advantage of multi paths by multiplexing those signals using a advance DSP algorithm to enhance the wireless bandwidth efficiency and range.

Wireless system using MIMO can improve the spectral efficiency of a system. Multiple antennas is required at the baseband signal processing components to handle multiple input signals, thus it have complexity for the decoder and the channel estimator at the receiver to decrease the complexity of such systems, several schemes [10]-[15] were presented in the literature. For the OFDM systems with multiple transmit antennas, explicity or implicitly considered that the channel state information (CSI) is known at the transmitter. In mobile communication, where the channel varies rapidly, it is difficult to maintain the CSI at the transmitter up-to-date without substantial system overhead [16].Space-time-frequency codes were proposed for OFDM systems to obtain an advantage of the frequency diversity and spatial diversity presented in frequency selective fading channels without the requirement of the availability of CSI at the transmitter. For such system, traditional subcarrier based space processing which introduce considerable complexity. In this paper, we proposed to use pre-DFT processing to reduce the receiver complexity of MIMO-OFDM systems with Space-time-frequency coding. In our proposed work received signals at the receiver are first weighted and then combined before the DFT processing. In the DFT processing blocked required at the receiver can be reduced, and a high dimensional MIMO system can be compressed into an equivalently low dimension. Both reduce the complexity.

An important issue in the proposed pre-DFT processing scheme for MIMO-OFDM systems with space-time-frequency coding is the estimation of the weighting coefficient before the DFT processing. In this paper, we propose a universal weighting coefficient calculation algorithm that can be applied in most practical space-time-frequency codes such as those presented in [1]-[4], [6],[8] and [9]. Design of the pre-DFT processing scheme is independent of the optimization of the space-time-frequency coding. In general, the weighting coefficient is determined before the DFT processing assuming that the CSIs are explicitly available. In this paper we will exhibit that the weighting coefficient can also be obtained by signal space method [10], [11] without the knowledge of the CSIs. This will reduce the complexity of the

channel estimation required by the space-time-frequency decoding science the number of the equivalent channel branches required to calculate the proposed scheme can reduced from the number of the receiver antenna. This paper is organized as follows. In section II, describes MIMO-OFDM system. The calculation of weighting coefficient with/without explicit CSIs is described in the Section 3 and 4. Some concept about the proposed pre-DFT processing scheme are given in Section 5. simulation result is explained in the Section 6. Finally 7 contain conclusion of tour work.

2. System Model

We analyze a MIMO-OFDM system with N subcarriers as shown in figure.1. In this system, there are F transmit antennas and M receiver antennas. At the nth OFDM symbol period, the output of the space-time-frequency encoder is considered as follows:

$$C^{t} = c_{0,1}^{t},....,c_{N-1,1}^{t},...,c_{o,F}^{t},...,c_{N-1,F}^{t}$$
$$t = O, 1, ..., T-1 \text{ (1)}$$

Where $c_{n,f}^{(t)}$ is the coded information symbol at the nth subcarrier and t^{th} OFDM symbol period transmitted from the fth transmit antenna and T is the number of OFDM symbols in a space-time-frequency code reduces to space-frequency code. After the IDFT processing at the tth OFDM symbol period the lth sample at the fth transmit antenna is given by

$$s_{t,l}^{f} = \frac{1}{N}\sum_{n=0}^{N-1} c_{n,f}^{t} e^{j2\pi ln/N},$$
$$-N_{g} \le l < N, f = 1,...,F, t = 0,...,T-1$$

(2)

Where N_{g} is the length of the cyclic prefix, and we consider that the $(N_{g}+1) < N$ to keep high transmission efficiency. We consider that the channel does not vary over the period of one space-time-frequency codeword. We assume that the channel impulse responses (CIRs) decay to zero during the cyclic extension, or $L \le (N_{g}+1)$ where L is the maximum length.

At m^{th} receive antenna, the lth sample at the tth OFDM symbol period is given by

$$r_{t,l}^{m} = \sum_{f=1}^{r} h_{l}^{m,f} * s_{t,l}^{f} + z_{t,l}^{m}$$

$$-N_g \leq l < N, m = 1,...,M, t = 0,...,T-1 \quad (3)$$

Where * denotes the convolution product, $h_i^{(m,f)}$ denotes the CIR between the fth transmit antenna and m^{th} receive antenna and $z_{t,l}^{(m)}$ denotes the additive white Gaussian noise (AWGN) component at the mth receive antenna. Before the DFT processing, the M data streams from the output of the M receive antenna are weighted and then combined to form Q branches. After the removal of guard interval the weighted and combined signals are then applied to the DFT processors. There are Q branches, and the number of the DFT blocks required at the receiver is Q and the result is compared with the conventional receiver structure [1]-[9].M DFT blocks are used, the number of DFT blocks employed at the receiver can be decreased when the PRE-DFT processing is employed.

For the qth branch, the output of the DFT processor at the tth OFDM symbol period is given as

Where

$$v_{n,q}^{(t)} = \sum_{f=1}^{F} \sum_{m=1}^{M} \omega_{m,q} H_n^{(m,f)} c_{n,f}^{(t)} + \sum_{m=1}^{M} \omega_{m,q} \hat{Z}_{t,n}^{(m)}$$

$$H_n^{(m,f)} = \sum_{l=0}^{L-1} h_l^{(m,f)} e^{-j2\pi ln/N}$$

$$\hat{Z}_{t,n}^{(m)} = \sum_{l=0}^{N-1} Z_{t,l}^{(m)} e^{-j2\pi ln/N}$$

And $\omega_{m,q}$ is the weighting coefficient for the mth receiver antenna at the qth branch. To keep the noise white and its variance at different branch are the same. We consider as the weighting coefficient are normalized i.e., $\Omega^H \Omega = I_Q$, where

Ω is a $M \times Q$ matrix with the $(m,q)^{th}$ entry given by $\omega_{m,q}$ and I_Q is $Q \times Q$ identify matrix.

3. Weighting Coefficients Calculation With Explicit Csi

This section describes the way of calculating the weighting coefficients for the proposed pre-DFT processing. When the ML decoder is employed, the pair-wise error probability (PEP) can be used to denote the performance, which is further calculated by the pair-wise codeword distance [11]. The pair-wise codeword distanced d^2 $C, E \mid H$ between a favored coded sequences.

$$E = \left[E^0, E^1, ..., E^{T-1} \right]^T$$

Where

$$E^t = \left[e_{0,1}^t, ..., e_{N-1,1}^t, ..., e_{0,F}^t, ..., e_{N-1,F}^t \right]$$

$\forall t \in 0, T-1$ And the transmitted coded sequence

$$C = [C^{(0)}, C^{(1)}, ...C^{(T-1)}]^T$$

Where C^t $\forall t \in 0, T-1$ is defined in (1), is given by

$$d2 \ C, E \mid H = \sum_{t=0}^{T-1} \sum_{q=1}^{Q} \sum_{n=0}^{N-1} \sum_{m=1}^{M} \sum_{m'=1}^{M} \omega_{m,q} \omega_{m',q}^* \quad (4)$$

$$\sum_{f=1}^{F} \sum_{f'=1}^{F} H_n^{m,f} \ H_n^{m',f'^{*}} \ c_{n,f}^t - e_{n,f}^t \ c_{n,f'}^t - e_{n,f'}^{t^{*}} \quad (7)$$

According to [11] maximizing the pair-wise codeword distance is equivalent to minimizing the pair-wise error probability is given by (7). Equation (7) indicates that the optimal weighting coefficients are related to the specific codeword pair. To make the codeword pair and weighting coefficients independent, we average (7) over all codeword's pair ensemble

As a result, we obtained

$$\overline{d^2 \ C,E|H} = \sum_{q=1}^{Q}\sum_{m=1}^{M}\sum_{m'=1}^{M} \omega_{m,q}\omega_{m',q}^{*} \sum_{n=0}^{N-1}\sum_{f=1}^{F}\sum_{f'=1}^{F} H_n^{(m,f)}(H_n^{m',f'})^{*} \sum_{t=0}^{T-1}(c_{n,f}^{(t)}-e_{n,f}^{(t)})(c_{n,f'}^{(t)}-e_{n,f'}^{(t)})^{*}$$

(8)

Where the over bar stands for the average of all code words pair. In order to rewrite the equation(8) into matrix form, consider C_n is an $F \times T$ matrix with the $(f,t)^{th}$ entry given by $c_{n,f}^{(t)}$, E_n be an matrix with the $(f,t)^{th}$ entry given by $e_{n,f}^{(t)}$, and $H_n(n=0,...,N-1)$ be an $M \times F$ matrix with the $(m,f)^{th}$ entry given by $H_n^{(m,f)}$. With these equation (8) can be written into

$$\overline{d^2 \ C,E|H} = trace \ \Omega^T \Phi \Omega^*$$

(9)

Where

$$\Phi = \sum_{n=0}^{N-1} H_n k_n H_n^{H}$$

(10)

With

$$k_n = \overline{C_n - E_n \ \ C_n - E_n}^{H}.$$

(11)

Let the Eigen values of Φ be $\lambda_q (q=1,.....,Q)$ with $\lambda_1 \geq \lambda_2 \geq ..... \geq \lambda_M$ and $w_q(q=1,....,Q)$ be the ith column of Ω. It is well known that when $\omega_q(q=1,....,Q)$ conjugate of the Eigen values are $\lambda_q(q=1,...,M)$, the maximum of $d^2 \ C,E|H$ is achieved and given by[22]

$$\overline{d^2 \ C,E|H}_{max} = \sum_{q=1}^{Q}\lambda_q \ \Phi .$$

(12)

In general, to obtain Φ in (14), we need knowledge of both the CSIs and the space-time-frequency code since k_n is dependent on the specific space-time-frequency code. Transmitter
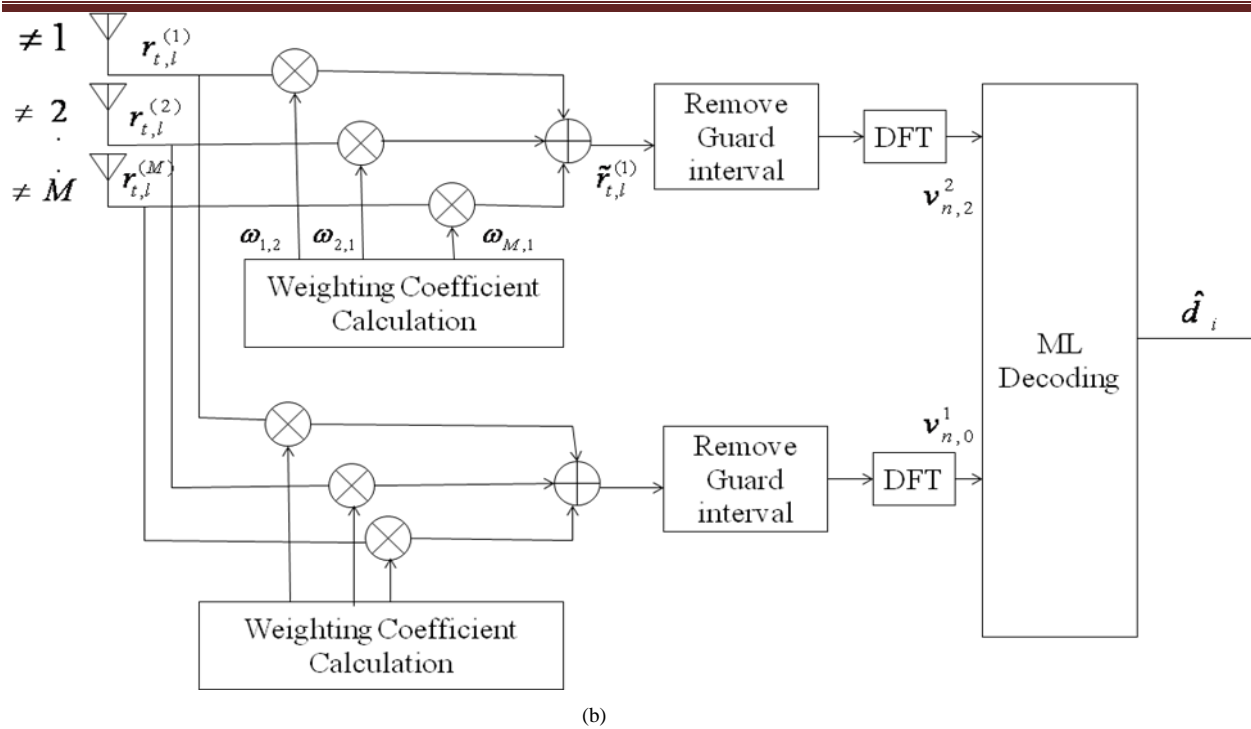

(a)

(b)
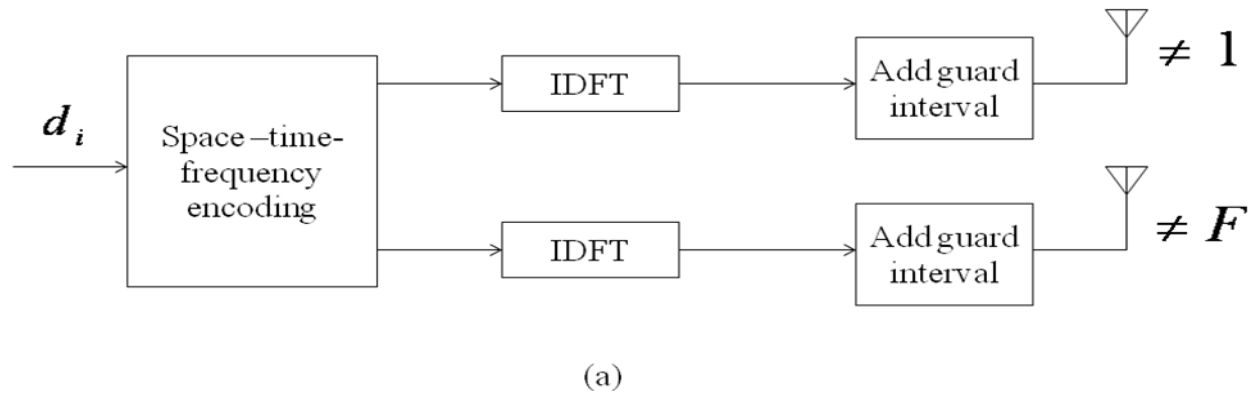
Figure 3.1 .PreDFT processing for a MIMO OFDM system (a) Transmitter, (b) Receiver

Does not contain any channel information. The space-time-frequency coding will not favor a particular sub-carrier or a particular transmit antenna. As a result, in the following, it will be given that for most practical space-time-frequency codes, it is reasonable to assume that Φ is in the following form

$$\Phi = k \sum_{n=0}^{N-1} H_n H_n^H \quad (13)$$

Where k is a constant which is independent of n. As the result the weighting coefficients are conjugate of the Eigen value of, are independent of the specific space-time-frequency coding for the space-time-frequency codes proposed in [8] and [9],A, k_n can be expressed as follows:

$$k_n = k_1 . diag \quad \beta_{\tau\ n} \quad (14)$$

Where k_1 a   constant   number   is   independent   of

n, $\beta_{r(n)} = [0,...,0,....,0]$ is an F dimensional standard

basis vector with 1 in its $\tau(n)$ component and 0 elsewhere, and $\tau(n)$ is determined by the space-frequency coding. Φ Can be proved to be in the form of (13) with $k = k_1/F$ for space-time-frequency codes (STBC) [17][18] is employed as a inner code. Using the orthogonal property of STBC, we can prove that

$$k_n = diag\left(\overline{2\left|c_{n,1}^0 - e_{n,1}^0\right|^2}, ..., \overline{2\left|c_{n,F}^0 - e_{n,F}^0\right|^2} \right)$$

(15)

It is reasonable to consider that the signal at the input of the inner encoder have the same distribution for the different subcarriers and different transmit antennas, especially when an interleaves is between the outer encoder and the inner encoder. Thus k_n can be written as

$$k_n = kI_F \quad (16)$$

Therefore,(10) can be simplified into(13) for these codes. For general space-time-frequency code as proposed in [6], simulation results in Section VI will also exhibits the excellent performance achieved by using the weighting coefficient determined based on given by (13)

4. Weighting Coefficient Calculation Without Exxplicit Csi

In this Section we propose a way to obtain the weighting coefficient without explicit CSI. This is important for differential modulation and CSI is not supposed to be explicitly required for the weighting coefficients calculation, the complexity in the channel estimation can be reduced by number of receiver antennas to the number of DFT branches. The covariance matrix of the received signal vector can be given by

$$r_{t,l} = \left[r_{t,l}^1, r_{t,l}^2, ..., r_{t,l}^M \right]^T$$ Can be given as

$$R = \mathrm{E}\left[r_{t,l} r_{t,l}^{\mathrm{H}} \right] \quad (17)$$

The $(m, m')^{th}$ element of r is given by

$$\rho_{m,m'} = \mathrm{E}\left[r_{t,l}^m \quad r_{t,l}^{m'}{}^* \right]$$

$$= \sum_{f=1}^{F} \sum_{f'=1}^{F} \sum_{\upsilon=l-L+1}^{l} \sum_{\upsilon'=l-L+1}^{l} h_{l-\upsilon}^{m,f} \quad h_{l-\upsilon'}^{m',f'}{}^*$$

$$\mathrm{E}\left[s_{t,\upsilon}^f \quad s_{t,\upsilon'}^{f'}{}^* \right] + N_0 \delta \quad m - m'.$$

(18)

When a large number of subcarriers are used, it is reasonable to consider that the transmitted signal is white, that is

$$\mathrm{E}\left[s_{t,\upsilon}^f \quad s_{t,\upsilon'}^f{}^* \right] = \mathrm{E}_s \delta \quad f - f' \quad \delta \quad \upsilon - \upsilon'$$

(19)

Where E_s is the average energy of the coded symbol, Hence, by substituting (19) into (18) after some manipulations $\rho_{m,m'}$ can be given by

$$\rho_{m,m'} = \frac{E_s}{N} \left(\sum_{f=1}^{F} \sum_{n=0}^{N-1} \mathrm{H}_n^{m,f} \quad \mathrm{H}_n^{m',f}{}^* \right) + N_0 \delta \quad m - m'$$

(20)

Where N_0 is the variance of the noise, using (13),we get

$$\rho_{m,m'} = \frac{E_s}{Nk} \Phi_{m,m'} + N_0 \delta \quad m - m' \quad (21)$$

Where $\Phi_{m,m'}$ is the $(m, m')^{th}$ entry of Φ.

From (21), it can be seen that the eigenvectors of Φ are the same as those of R. As a result, we can obtain the weighting coefficient directly from R without explicit knowledge of CSI.

5. Complexity Consideration

The proposed MIMO-OFDM system consists of pre-DFT weighting and combining, weighting coefficient calculation, DFT-processing channel estimation and ML decoding. By weighting and combining before the DFT processing. The number of branches to be handled by the ML, decoder is reduced from M to Q. As a result, comparing with subcarrier based processing [1]-[9], the complexity of ML decoding can be decreased.

As for the complexity coming from the DFT processing the pre-DFT weighting and combining, the number of multiplications needed in between the proposed scheme and subcarrier based scheme is given as:

$$\eta = \frac{QN \log_2 N + QMN}{MN \log_2 N} = \frac{Q}{M} \left(\frac{\log_2 N + M}{\log_2 N} \right)$$

(22)

From (22), it can be seen that, when $\log_2 N \gg M$. η Is close to Q/M. from (12), the number of DFT blocks at the receiver, Q.is determined by the rank of Φ . After some manipulations, we have

$$rank \quad \Phi \leq \min \quad rank \quad \tilde{R}^{1/2} \quad , M, FL \quad.$$

(23)

Where

$$\tilde{R}^{1/2} = \left[R_0^{1/2}, R_1^{1/2}, ..., R_{L-1}^{1/2} \right]. \quad (24)$$

While $R_l = R_l^{1/2} R_l^{1/2}$ are the receive correlation matrix as defined in [19]. From (23), we obtain Φ is singular when $R^{1/2}$ is not of full row rank or FL is smaller than M. In this case, the number of DFT blocks required will be less than the number of receiver antennas to achieve optimal performance. When is non singular, still it is possible to achieve good performance with a limit number of DFT blocks due to the small contribution of the small eigen value to the average pair-wise codeword distance $\overline{d^2 \; C, E \mid H} \;..$
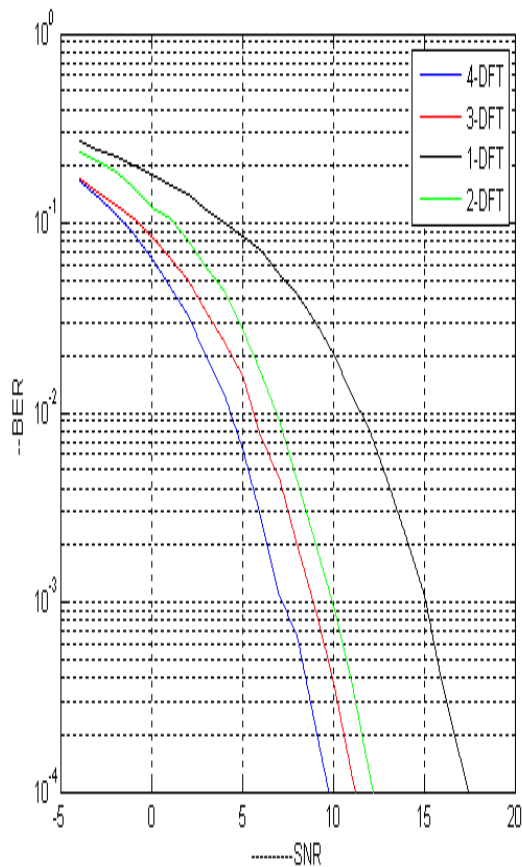
6. Simulation Results

Figure 6.1 Performance analysis for 2×4 antenna system simulated under QPSK modulation for different DFT's
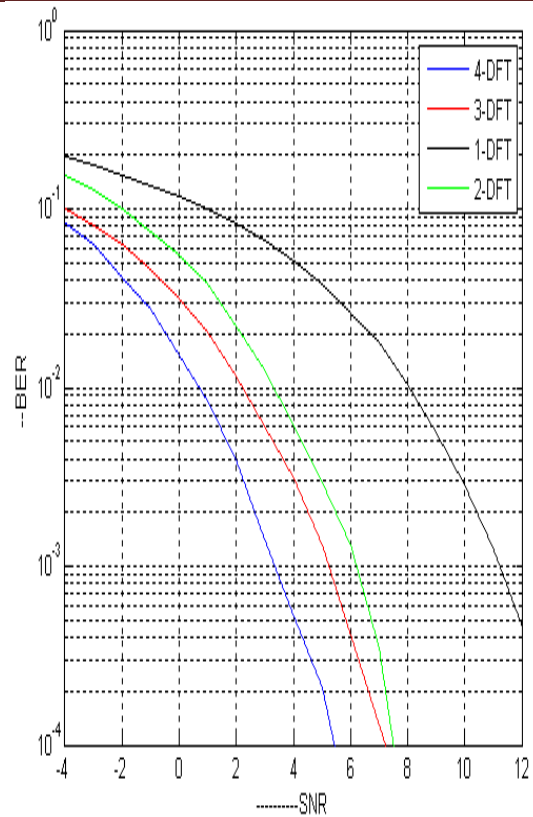

Figure 6.2 Performance analysis for 4×4 antenna system simulated under QPSK modulation for different DFT's.

7. Conclusion

In this paper pre-DFT processing method was proposed for a MIMO-OFDM system with Space-time frequency coding. Combination of MIMOOFDM supports the increasing demand of the user in communication network. The proposed system decreases the complexity and enhances the performance. Simple weighting coefficient algorithm is derived. The theoretical analysis and simulation result shows that the algorithm can be applied for most existing practical space-time-frequency codes. MIMO OFDM is robust for wide range of channel condition. In the proposed system the number of DFT blocks required is reduced to achieve near optimal system performance.

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