CHANNEL ESTIMATION FOR FAST-VARYING MIMO-OFDM SYSTEMS

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ABSTRACT

Multiple input-multiple output (MIMO) systems hold the potential to drastically improve the spectral efficiency and link reliability in future wireless communications systems. A particularly promising candidate for next-generation fixed and mobile wireless systems is the combination of MIMO technology with Orthogonal Frequency Division Multiplexing (OFDM). OFDM has become the standard method because of its advantages over single carrier modulation schemes on multi-path, frequency selective fading channels. Doppler frequency shifts are expected in fast-moving environments, causing the channel to vary in time, that degrades the performance of OFDM systems. In this paper, we present a time-varying channel modelling and estimation method based on the Discrete Evolutionary Transform to obtain a complete characterization of MIMO-OFDM channels. Performance of the proposed method is evaluated and compared on different levels of channel noise.

1. INTRODUCTION

The major challenges in future wireless communications systems are increased spectral efficiency and improved link reliability. The wireless channel constitutes a hostile propagation medium, which suffers from fading (caused by destructive addition of multipath components) and interference from other users. Diversity provides the receiver with several (ideally independent) replicas of the transmitted signal and is therefore a powerful means to combat fading and interference and thereby improve link reliability. Common forms of diversity are time diversity (due to Doppler spread) and frequency diversity (due to delay spread). In recent years the use of spatial (or antenna) diversity has become very popular, which is mostly due to the fact that it can be provided without loss in spectral efficiency. Receiver diversity, that is, the use of multiple antennas on the receiver side of a wireless link, is a well-studied subject [1]. Driven by mobile wireless applications, where it is difficult to deploy multiple antennas in the handset, the use of multiple antennas on the transmitter side combined with signal processing and coding has become known under the name of space-time coding [2] and is currently an active area of research. The use of multiple antennas at both ends of a wireless link (multiple-input multiple-output (MIMO) technology) has been demonstrated to have the potential of achieving extraordinary data rates [3]. The corresponding technology is known as spatial multiplexing [4] or BLAST [5] and yields an impressive increase in spectral efficiency. Most of the previous work in the area of MIMO wireless has been restricted to narrowband systems. Besides spatial diversity broadband MIMO channels, however, offer to higher capacity and frequency diversity due to delay spread. Orthogonal frequency division multiplexing (OFDM) [6] significantly reduces receiver complexity in wireless broadband systems. The use of MIMO technology in combination with OFDM, i.e., MIMO-OFDM [4], therefore seems to be an attractive solution for future broadband wireless systems.
However, inter-carrier interference (ICI) due to Doppler shifts, phase offset, local oscillator frequency shifts, and multi-path fading severely degrades the performance of OFDM systems [7, 8]. Most of the channel estimation methods assume a linear time–invariant model for the channel, which is not valid for the next generation, fast-moving environments [9]. A time-varying model of the channel can be obtained by employing time-frequency representation methods. Here we present a time–varying MIMO-OFDM channel estimation based on the discrete evolutionary representation of channel output. The Discrete Evolutionary Transform (DET) provides a time- frequency representation of the received signal by means of which the spreading function of the multi-path, fading and frequency selective channel can be modelled and estimated.

2. WIRELESS CHANNEL MODEL

In wireless communications, the multi-path, fading channel with Doppler frequency shifts may be modelled as a linear time-varying system with the following discrete-time impulse response [10]

$$h(m,\ell)=\sum_{\rho=0}^{L_p-1}\alpha_\rho\, e^{j\psi_\rho m}\;\delta(\ell-N_\rho) \qquad (1)$$

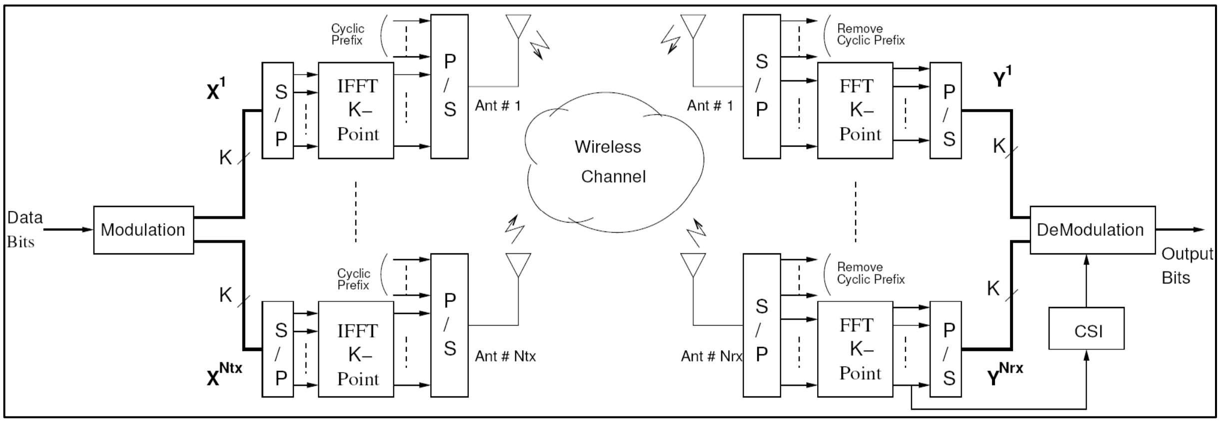

Figure 1- MIMO-OFDM System Model

Where L_p is the total number of transmission paths, ψ_ρ represents the Doppler frequency, α_ρ is the relative attenuation, and N_ρ time delay caused by path ρ. Time-varying transfer function of this linear channel is calculated by taking the discrete Fourier transform (DFT) of the impulse response with respect to ℓ, i.e.,

$$H^{(i,j)}(m,\omega_k) = \sum_{\rho=0}^{L_p-1} \alpha_\rho e^{j\psi_\rho m}\, e^{-j\omega_k N_\rho}$$

Where $\omega_k = \frac{2\pi}{K}k,\ k = 0,1,\cdots,K-1$. Now, the bi-frequency function $B(\Omega_s,\omega_k)$ is found by computing the discrete Fourier transform of $H(m,\omega_k)$ with respect to time variable, *m*:

$$B(\Omega_s,\omega_k) = \sum_{\rho=0}^{L_p-1} \alpha_\rho\, e^{-j\omega_k N_\rho}\delta(\Omega_s - \psi_\rho)$$

$\Omega_s = \frac{2\pi}{K}s,\ s = 0,1,\cdots,K-1$. Furthermore, the spreading function of the channel is obtained by calculating the DFT of $h(m,\ell)$ with respect to *m*, or by taking the inverse DFT of $B(\Omega_s,\omega_k)$ with respect to ω_k

$$S\ (\Omega_s,\ell) = \sum_{\rho=0}^{L_p-1} \alpha_\rho \delta(\Omega_s - \psi_\rho)\delta(\ell - N_\rho)$$

which displays peaks located at the time-frequency positions determined by the delays and the corresponding Doppler frequencies, and with α_ρ as their amplitudes [10]. If we extract this information from the received signal, we will be able to eliminate the effects of the time-varying channel and estimate the transmitted data symbol.

3. MIMO-OFDM SYSTEM MODEL

In an OFDM communication system, the available bandwidth B_d is divided into K subchannels. The input data is also divided into K-bit parallel bit streams, and then mapped onto some transmit symbols $X_{n,k}$ drawn from an

arbitrary constellation points where n is the time index, and $k = 0,1,\cdots,K-1$, denotes the frequency or subcarrier index.

We then insert some pilot symbols, at some pilot positions (n',k'), known to the receiver: $(n',k') \in P = \{(n',k') \mid n' \in Z, k' = iS + (n' mod(S)), i \in [0, P-1]\}$ where P is the number of pilots, and the integer $S = K/P$ is the distance between adjacent pilots in an OFDM symbol [7]. The n^{th} OFDM symbol $s_n(m)$ is obtained by taking the inverse DFT and then adding a cyclic prefix of length L_{CP} where L_{CP} is chosen such that $L \le L_{CP} + 1$, and L is the time-support of the channel impulse response. This is done to mitigate the effects of intersymbol interference (ISI) caused by the channel time spread [8].

The system model is given in Fig.1, a MIMO-OFDM system with N_{tx} transmit and N_{rx} receive antennas, is assumed. The system has K subcarriers in an OFDM block, and another L_{CP} subcarriers are added as a guard band (cyclic prefix, CP). The incoming bits are modulated to form $X_{n,k}^{(i)}$, where i is the transmit index. For each modulated signal, an inverse DFT of size K is performed, and the CP is added to mitigate for the residual ISI due to previous OFDM symbol. After parallel-to-serial (P/S) conversion, signal is

$$s_n^{(i)}(m) = \frac{1}{\sqrt{K}} \sum_{k=0}^{K-1} X_{n,k}^{(i)} e^{j\omega_k m}$$

$m = -L_{CP}, -L_{CP}+1,\cdots,0,\cdots,K-1$ where again $\omega_k = \frac{2\pi}{K}k$, and each OFDM symbol has $N = K + L_{CP}$ length. The channel output suffers from multi-path propagation, fading and Doppler frequency shifts:

$$\begin{aligned} y_n^{(j)}(m) &= \sum_{i=1}^{N_{tx}} \left(\sum_{\ell=0}^{L_p^{(i,j)}-1} h^{(i,j)}(m,\ell)\, s_n^{(i)}(m-\ell) \right) \\ &= \sum_{i=1}^{N_{tx}} \left(\sum_{\rho=0}^{L_p^{(i,j)}-1} \alpha_\rho^{(i,j)}\, e^{j\psi_\rho^{(i,j)} m}\, s_n^{(i)}(m - N_\rho^{(i,j)}) \right) \\ &= \sum_{i=1}^{N_{tx}} \left(\frac{1}{\sqrt{K}} \sum_{k=0}^{K-1} X_{n,k}^{(i)} \sum_{\rho=0}^{L_p^{(i,j)}-1} \alpha_\rho^{(i,j)}\, e^{j\psi_\rho^{(i,j)} m}\, e^{j\omega_k (m - N_\rho^{(i,j)})} \right) \\ &= \sum_{i=1}^{N_{tx}} \left(\frac{1}{\sqrt{K}} \sum_{k=0}^{K-1} H_n^{(i,j)}(m,\omega_k) e^{j\omega_k m}\, X_{n,k}^{(i)} \right) \end{aligned}$$

where $(.)^{(j)}$ denotes the j^{th} receiver $(.)^{(i,j)}$ is indexing for the wireless time varying channel between the i^{th} transmitter and the j^{th} receiver. The transmit signal is also corrupted by additive white Gaussian noise $\eta^{(j)}(m)$ over the channel. The received signal for the n^{th} frame can then be written as

$r_n^{(j)}(m) = y_n^{(j)}(m) + \eta_n^{(j)}(m)$. The receiver discards the Cyclic Prefix and demodulates the signal using a K-point DFT as

$$\begin{aligned} R_{n,k}^{(j)} &= \frac{1}{\sqrt{K}} \sum_{m=0}^{K-1} \left[y_n^{(i)}(m) + \eta_n^{(i)}(m) \right] e^{-j\omega_k m} \\ &= \sum_{i=1}^{N_{tx}} \left(\frac{1}{\sqrt{K}} \sum_{s=0}^{K-1} X_{n,s}^{(i)} \sum_{\rho=0}^{L_\rho^{(i,j)}-1} \alpha_\rho^{(i,j)} e^{j\omega_s m - N_\rho^{(i,j)}} \right. \\ &\qquad \left. \times \sum_{m=0}^{K-1} e^{j\psi_\rho^{(i,j)} m} e^{j(\omega_s - \omega_k)m} + Z_{n,k}^{(j)} \right) \end{aligned} \tag{2}$$

If the Doppler effects in all the channel paths are negligible, $\psi_\rho^{(i,j)} = 0, \forall i$, then the channel is almost time--invariant within one OFDM symbol. In that case, above equation becomes

$$\begin{aligned} R_{n,k}^{(j)} &= \sum_{i=1}^{N_{tx}} X_{n,k}^{(i)} \sum_{\rho=0}^{L_\rho^{(i,j)}-1} \alpha_\rho^{(i,j)} e^{-j\omega_k N_\rho^{(i,j)}} + Z_{n,k}^{(j)} \\ &= \sum_{i=1}^{N_{tx}} H_{n,k}^{(i,j)} X_{n,k}^{(i)} + Z_{n,k}^{(j)} \end{aligned}$$

where $H_{n,k}^{(i,j)}$ is the frequency response of the channel between the i^{th} transmitter and j^{th} receiver antenna, and $Z_{n,k}^{(j)}$ is the Fourier transform of the channel noise at the j^{th} receiver. If there are large Doppler frequency shifts in the channel, then the time-invariance assumption above is no longer valid. Here we consider time-varying channel modelling and estimation and approach the problem from a time-frequency point of view [10, 12].

4. TIME-VARYING CHANNEL ESTIMATION FOR MIMO-OFDM SYSTEMS

In the following we briefly explain the Discrete Evolutionary Transform as a tool for the time--frequency representation of wireless channel output.

4.1 Time-Frequency Analysis by DET

A non-stationary signal, $x(n), 0 \le n \le N-1$, may be represented in terms of a time-varying kernel $X(n,\omega_k)$ or its corresponding bi-frequency kernel $X(\Omega_s,\omega_k)$. The time-frequency discrete evolutionary representation of $x(n)$ is given by [13],

$$x(n) = \sum_{k=0}^{K-1} X(n,\omega_k) e^{j\omega_k n},$$

where $\omega_k = 2\pi k / K$, K is the number of frequency samples, and $X(n,\omega_k)$ is the evolutionary kernel. The discrete evolutionary transformation (DET) is obtained by expressing the kernel $X(n,\omega_k)$ in terms of the signal. This is done by using conventional signal representations [10]. Thus, for the representation above, the DET that provides the evolutionary kernel $X(n,\omega_k)$, $0 \le k \le K-1$, is given by

$$X(n,\omega_k) = \sum_{\ell=0}^{N-1} x(\ell) \mathrm{w}_k(n,\ell) e^{-j\omega_k \ell},$$

where $\mathrm{w}_k(n,\ell)$ is, in general, a time and frequency dependent window. Details of how the windows can be obtained are given in [13]. However, for the representation of multipath wireless channel outputs, we need to consider signal-dependent windows that are adapted to the Doppler frequencies of the channel.

4.2 MIMO-OFDM Channel Estimation

We will now consider the computation of the spreading function by means of the evolutionary transformation of the received signal. The output of the channel, after discarding the cyclic prefix, for the n^{th} OFDM symbol can be written as,

$$\begin{aligned} y_n^{(j)}(m) &= \sum_{i=1}^{N_{tx}} \left(\frac{1}{\sqrt{K}} \sum_{\rho=0}^{L_\rho^{(i,j)}-1} \sum_{k=0}^{K-1} \alpha_\rho^{(i,j)} e^{j\psi_\rho^{(i,j)} m} e^{j\omega_k (m - N_\rho^{(i,j)})} X_{n,k}^{(i)} \right) \\ &= \frac{1}{\sqrt{K}} \sum_{\rho=0}^{L_\rho^{(i,j)}-1} H_n^{(i,j)}(m,\omega_k) e^{j\omega_k m} X_{n,k}^{(i)} \end{aligned} \tag{3}$$

The equation above can be rewritten in matrix form as,

$$\mathbf{y}^{(j)} = \mathbf{H}^{(j)} \mathbf{x} \;, \tag{4}$$

where

$$\begin{aligned} \mathbf{y}^{(j)} &= \left[y_n^{(j)}(0), y_n^{(j)}(1), \ldots, y_n^{(j)}(K-1) \right]^T; \\ \mathbf{x}^{(i)} &= \left[X_{n,0}^{(i)}, X_{n,1}^{(i)}, \ldots, X_{n,K-1}^{(i)} \right], \\ \mathbf{x} &= \left[\mathbf{x}^{(1)}, \mathbf{x}^{(2)}, \ldots, \mathbf{x}^{(N_{tx})} \right]^T; \\ \mathbf{H}^{(i,j)} &= \left[a_{m,k} \right]_{K \times K}, \quad a_{m,k} = \frac{H_n^{(i,j)}(m,\omega_k) e^{j\omega_k m}}{\sqrt{K}}, \\ \mathbf{H}^{(j)} &= \left[\mathbf{H}^{(1,j)}, \mathbf{H}^{(2,j)}, \ldots, \mathbf{H}^{(3,j)} \right]. \end{aligned}$$

From (4), the input-output relation for the whole system results as

$$\mathbf{y} = \mathbf{H}\mathbf{x}$$

where

$$\mathbf{H} = \begin{bmatrix} \mathbf{H}^{(1)} \\ \mathbf{H}^{(2)} \\ \vdots \\ \mathbf{H}^{(N_{rx})} \end{bmatrix}, \quad \mathbf{y} = \begin{bmatrix} \mathbf{y}^{(1)} \\ \mathbf{y}^{(2)} \\ \vdots \\ \mathbf{y}^{(N_{rx})} \end{bmatrix}.$$

If $\mathbf{H}$ is known and then input symbols can be estimated via the following relation,

$$\hat{\mathbf{x}} = \mathbf{H}^{-1}\mathbf{y} \tag{5}$$

Now calculating the discrete evolutionary representation of $y_n^{(j)}(m)$, we get

$$y_n^{(j)}(m) = \sum_{k=0}^{K-1} Y_n^{(j)}(m,\omega_k) e^{j\omega_k m} \tag{6}$$

Comparing the representations of $y_n^{(j)}(m)$ in (3) and (6), we get the kernel as

$$Y_n^{(j)}(m,\omega_k) = \sum_{i=1}^{N_{tx}} \left(\frac{1}{\sqrt{K}} H_n^{(i,j)}(m,\omega_k) X_{n,k}^{(i)} \right)$$

Above relation is also valid at the preassigned pilot positions $k = k'$;

$$
\begin{aligned}
Y_n^{(j)}(m,\omega_p) &= Y_n'^{(j)}(m,\omega_{k'}) \\
&= \sum_{i=1}^{N_{tx}} \left(\frac{1}{\sqrt{K}} H_n'^{(i,j)}(m,\omega_{k'})\, X_{n,k'}^{(i)} \right)
\end{aligned}
$$

where $p = 1,2\cdots,P$ and $H_n'^{(i,j)}(m,\omega_{k'})$ is a decimated version of the $H_n^{(i,j)}(m,\omega_k)$. Note that P is again the number of pilots, and $d = K/P$ is the distance between adjacent pilots. Finally if the pilot symbols are chosen to be orthogonal to each other the decimated frequency response of the channel between the i^{th} transmitter and the j^{th} receiver my be obtained as:

$$
H_n'^{(i,j)}(m,\omega_p) = \frac{\sqrt{K}}{\left\| X_{n,p}'^{(i)} \right\|} \left\langle Y_n'^{(j)}(m,\omega_p), X_{n,p}'^{(i)} \right\rangle \tag{7}
$$

Taking the inverse DFT of $H_n'^{(i,j)}(m,\omega_p)$ with respect to ω_p and DFT with respect to m , we obtain the down-sampled spreading function $S'(\Omega_s,\ell)$,

$$
S_n'^{(i,j)}(\Omega_s,\ell) = \frac{1}{d} \sum_{\rho=0}^{L_\rho^{(i,j)}-1} \alpha_\rho^{(i,j)}\, \delta_n^{(i,j)}(\Omega_s - \psi_\rho^{(i,j)})\; \delta_n^{(i,j)}\left(\frac{\ell - N_\rho^{(i,j)}}{d}\right)
$$

By comparing $S_n'^{(i,j)}(\Omega_s,\ell)$ and $S_n^{(i,j)}(\Omega_s,\ell)$, we observe that the channel parameters $\alpha_\rho^{(i,j)}$ $N_\rho^{(i,j)}$ and $\psi_\rho^{(i,j)}$ calculated from $S_n^{(i,j)}(\Omega_s,\ell)$, can also be estimated from the down-sampled spreading function $S_n'^{(i,j)}(\Omega_s,\ell)$. The time-frequency evolutionary kernel of the channel output in the j^{th} receiver is obtained as

$$
\begin{aligned}
Y_n^{(j)}(m,\omega_k) &= \sum_{\ell=0}^{K-1} y_n^{(j)}(\ell)\mathrm{w}_k(m,\ell) e^{-j\omega_k \ell} \\
&= \sum_{i=1}^{N_{tx}} \Bigg(\frac{1}{\sqrt{K}} \sum_{s=0}^{K-1} X_{n,s}^{(i)} \sum_{\rho=0}^{L_\rho^{(i,j)}-1} \alpha_\rho^{(i,j)}\, e^{-\omega_s N_\rho^{(i,j)}} \\
&\qquad \times \sum_{\ell=0}^{N-1} \mathrm{w}_k(m,\ell)\, e^{j\left(\psi_\rho^{(i,j)} + \omega_s - \omega_k\right)\ell} \Bigg)
\end{aligned}
$$

We consider windows of the form $\mathrm{w}_p(m,\ell) = e^{j\psi_p(m-\ell)}$, for $0 \le \psi_p \le \pi$ presented in [9] that depends on the Doppler frequency ψ_p . This window will give us the correct representation of $Y_n^{(j)}(m,\omega_k)$ only when $\psi_p = \psi_\rho^{(i,j)}$, in fact, using the window $\mathrm{w}_\rho(m,\ell) = e^{j\psi_\rho^{(i,j)}(m-\ell)}$, above representation of $Y_n^{(j)}(m,\omega_k)$ becomes,

$$
Y_n^{(j)}(m,\omega_k) = \sum_{i=1}^{N_{tx}} \left(\sqrt{K} H_n^{(i,j)}(m,\omega_k)\, X_{n,k}^{(i)} \right)
$$

which is the expected result multiplied by K .

4.3 Time-Frequency Receiver

After estimating the spreading function and the corresponding frequency response $H_n^{(i,j)}(m,\omega_k)$ of the channel, data symbols $X_{n,k}^{(i)}$ can be detected using a time-frequency receiver given in (5). On the other hand the channel output in equation (2) can be rewritten as

$$
\begin{aligned}
R_{n,k}^{(j)} &= \sum_{i=1}^{N_{tx}} \left(\frac{1}{K} \sum_{s=0}^{K-1} \left\{ \begin{array}{c} \sum_{m=0}^{K-1} H_n^{(i,j)}(m,\omega_k) \\ \times e^{j(\omega_s - \omega_k)m} \end{array} \right\} X_{n,s}^{(i)} \right) + Z_{n,k}^{(j)} \\
&= \frac{1}{K} \sum_{s=0}^{K-1} B_n^{(i,j)}(\omega_k - \omega_s, \omega_s) X_{n,s} + Z_{n,k}. \\
&= \sum_{i=1}^{N_{tx}} \left(\frac{1}{K} \sum_{s=0}^{K-1} B_n^{(i,j)}(\omega_k - \omega_s, \omega_s)\, X_{n,s}^{(i)} \right) + Z_{n,k}^{(j)}
\end{aligned}
$$

where $B_n^{(i,j)}(\Omega_s,\omega_k)$ is the bi-frequency function of the channel during n^{th} OFDM symbol, and above equation indicates a circular convolution with the data symbols. Based on above equality it is possible to write the MIMO-OFDM system consisting of N_{tx} transmitter and N_{rx} receiver antennas in a matrix form as

$$
\begin{pmatrix} \mathbf{r}^{(1)} \\ \vdots \\ \mathbf{r}^{(N_{rx})} \end{pmatrix} = \begin{pmatrix} \mathbf{B}^{(1,1)} & \cdots & \mathbf{B}^{(N_{tx},1)} \\ \vdots & \cdots & \vdots \\ \mathbf{B}^{(N_{rx},1)} & \cdots & \mathbf{B}^{(N_{tx},N_{rx})} \end{pmatrix} \square \begin{pmatrix} \mathbf{x}^{(1)} \\ \vdots \\ \mathbf{x}^{(N_{tx})} \end{pmatrix} + \begin{pmatrix} \mathbf{z}^{(1)} \\ \vdots \\ \mathbf{z}^{(N_{rx})} \end{pmatrix}
$$

or

$$
\mathbf{r} = \mathbf{Bx} + \mathbf{z}
$$

where $\mathbf{B}^{(i,j)} = \left[b_{s,k} \right]_{K\times K} = B_n^{(i,j)}(\omega_k - \omega_s, \omega_s)$ is a $K\times K$ matrix; $\mathbf{r}$, $\mathbf{x}$ and $\mathbf{z}$ are $K\times 1$ vectors defined by $\mathbf{r}^{(j)} = \left[R_{n,1}^{(j)}, R_{n,2}^{(j)}, \ldots, R_{n,K}^{(j)} \right]^T$, $\mathbf{x}^{(j)} = \left[X_{n,1}^{(j)}, X_{n,2}^{(j)}, \ldots, X_{n,K}^{(j)} \right]^T$, and $\mathbf{z}^{(j)} = \left[Z_{n,1}^{(j)}, Z_{n,2}^{(j)}, \ldots, Z_{n,K}^{(j)} \right]^T$ respectively. Finally, data symbols $X_{n,k}^{(j)}$ can be estimated by using a simple time-frequency: $\hat{\mathbf{x}} = \mathbf{B}^{-1}\mathbf{r}$.

5. SIMULATIONS

In the experiments, a 2-input, 2-output MIMO OFDM system is considered and the wireless channels are simulated randomly, i.e., the number of paths, $1 \le L \le 5$, the delays, $0 \le N_\rho \le L_{CP} - 1$ and the Doppler frequency shift $0 \le \psi_\rho \le \psi_{max}, (\rho = 0,1,\cdots,L_p - 1)$ of each path are picked randomly. Input data is QPSK coded and modulated onto K= 128 sub-carriers. The Signal-to-Noise Ratio (SNR) of the channel noise is changed between 0 and 15 dB, for a fixed value of the maximum Doppler $\psi_{max} = 500\text{Hz}$ on each path, and the bit error rate (BER) is calculated by five different approaches: 1) no channel estimation, 2) 8 pilots, 3) 16 pilots, 4) Pilot symbol assisted (PSA) time-invariant channel estimation 5) Perfect Channel State Information (CSI) . The spreading function, hence all the parameters of the channel are estimated by the proposed method. Fig.1 depicts an example of the estimated spreading function for one 2×2 MIMO-OFDM system, during one OFDM symbol. Finally, Fig. 2 shows the BER versus SNR for the above methods.

6. CONCLUSIONS

In this work, we present a time-varying estimation of MIMO-OFDM channels for fast-moving communication systems by means of discrete evolutionary transform. This approach allows us to obtain a representation of the time-dependent channel transfer functions from the noisy channel output. Examples show that, the BER performance of our proposed method improves as we increase the number of pilot symbols, as expected. Moreover, our time-frequency based method outperforms the pilot symbol assisted LTI channel estimation approach [9].

7. REFERENCES

[1] W. C. Jakes, Microwave mobile communications. New York: Wiley, 1974.

[2] S. M. Alamouti, "A simple transmit diversity technique for wireless communications," IEEE J.Select. Areas Commun., vol. 16, no. 8, pp. 1451-1458, Oct. 1998.

[3] A. J. Paulraj and T. Kailath, "Increasing capacity in wireless broadcast systems using distributed transmission/directional reception," U. S. Patent, no. 5,345,599, 1994.

[4] H. Bolcskei, D. Gesbert, and A. J. Paulraj, "On the capacity of OFDM-based spatial multiplexing systems," IEEE Trans. Commun., vol. 50, no. 2, pp. 225-234, Feb. 2002.

[5] G. J. Foschini and M. J. Gans, "On limits of wireless communications in a fading environment when using multiple antennas," Wireless Personal Communications, vol. 6, pp. 311-335, 1998.

[6] A. Peled and A. Ruiz, "Frequency domain data transmission using reduced computational complexity algorithms," in Proc. IEEE ICASSP-80, Denver, CO, 1980, pp. 964-967.

[7] Petropulu, A., Zhang, R., and Lin, R., "Blind OFDM Channel Estimation Through Simple Linear Precoding," *IEEE Trans. On Wireless Com.*, Vol. 3, No. 2, pp. 647–655, Mar. 2004.

[8] Stuber, G.L., Barry, J.R., Mc Loughlin, S.W., Li, Y.G., Ingram, M.A., Pratt, T.G., "Broadband MIMO-OFDM wireless communications," *IEEE Proceedings*, Vol. 92, No. 2, pp. 271–294, Feb. 2004.

[9] Kang, S.G., Ha, Y.M., Joo, E.K., "A comparative investigation on channel estimation algorithms for OFDM in mobile communications," *IEEE Trans. on Broadcasting*, Vol. 49, No. 2, pp. 142–149, June 2003.

[10] Hahm, M.D., Mitrovski, Z.I., and Titlebaum, E.L., Deconvolution in the Presence of Doppler with Application to Specular Multipath Parameter Estimation," *IEEE Trans. on Signal Proc.*, Vol. 45, No. 9, pp. 2203–2219, Sep. 1997.

[11] Schafhuber, D., Matz, G., Hlawatsch, F., "Adaptive Wiener filters for time–varying channel estimation in wireless OFDM systems," *IEEE ICASSP'03*, Vol. 4 , pp. IV - 688–691, Hong Kong, Apr. 6-10, 2003.

[12] Chaparro, L.F., and Alshehri, A.A., "Channel Modelling for Spread Spectrum via Evolutionary Transform," *IEEE ICASSP'04*, Vol. II, pp. 609–612, Montreal, Quebec, Canada, May 17–21, 2004.

[13] Suleesathira, R., Akan, A. and Chaparro, L. F., "Discrete Evolutionary Transform for Time-Frequency Signal Analysis," *J. Franklin Institute*, pp. 347-364, Vol. 337, No. 4, Jul. 2000.

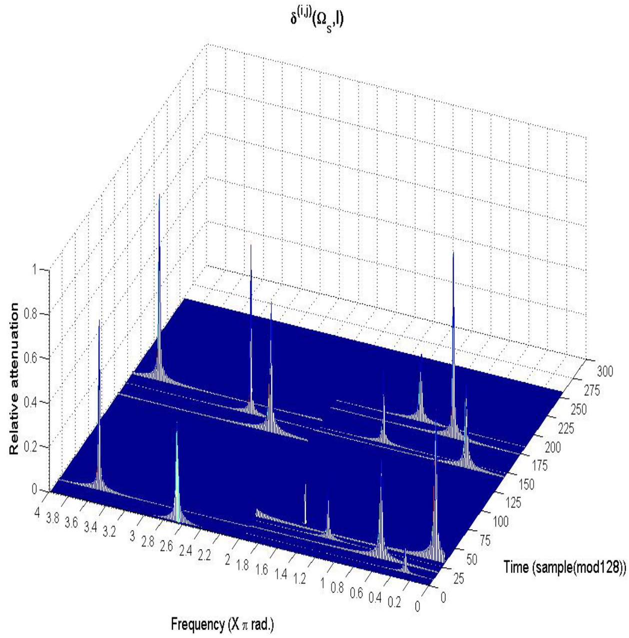

Figure 1- An Example of estimated spreading function for The 2×2 MIMO-OFDM system

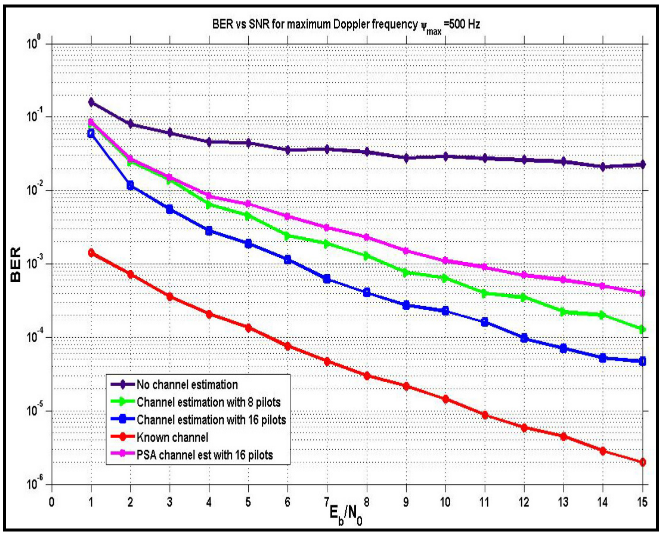

Figure 2- BER performance of Q-PSK modulated 2×2 MIMO-OFDM system without using STC .