International Journal of Computer Science and Engineering Communications
Vol.3, Issue 3, 2015, Page.1092-1102
ISSN: 2347–8586
www.scientistlink.org

Channel Estimation for MIMO OFDM System

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Abstract: A multiple-input multiple-output (MIMO) communication system combined with the orthogonal frequency division multiplexing (OFDM) technique can achieve reliable high data rate transmission over broadband wireless channels. Channel information for MIMO system based on training based, blind and semi blind methods. In training based channel estimation algorithms training symbols or pilot tones are known a priori to the receivers are multiplexed along with the data stream for channel estimation. The blind channel estimation is evaluating the statistical information of the channel and certain properties of the transmitted signals. The advantage of Blind Channel Estimation has no overhead loss and it is only applicable to slowly time-varying channels due to a long data record. Semi-blind channel technique is hybrid of blind, training technique utilizing pilots and other natural constraints to perform channel estimation.
The performance of MIMO OFDM channel estimation is evaluated on the basis of Mean Square Error (MSE) level. Training method algorithms are LSE, MMSE and LMS. MMSE gives low estimation error comparing with the LSE and LMS.LMS algorithm has low estimation error comparing with the LSE and low computational complexity comparing with the MMSE method because it does not require calculating the correlation matrix. Blind method of SVD-Iteration gives high estimation error and low computational complexity comparing with the SVD-Iteration. Semi-blind channel technique performance is much better than other methods like training based and blind channel estimation methods for the MIMOOFDM system.

***Keywords**: LSE, MMSE, LMS, AWGN, STBC, SVD, LTE**,** ETSI, ADSL, DAB.*

1. Introduction

As wireless services become more and more advanced higher data rates achieved in the communication systems. The limiting factor of the available bandwidth, MIMO antenna systems have been proposed to increase channel capacity over a fixed spectrum. To tackle another large problem in wireless communication is caused by multipath fading channels, OFDM has been proposed. The technique simplifies the channel equalization process by using MIMO systems in combination with OFDM an efficient communication system can be formed with high spectral efficiency and low sensitivity to multipath fading. To increase the spectral efficiency of these systems several efficient techniques for channel estimation like Training-Based, Blind and Semi-blind channel estimations are employed[1].

2. Orthogonal Frequency Division Multiplexing

The basic principle of OFDM is to split a high-rate data stream into a number of lower rate streams are transmitted simultaneously over a number of subcarriers. Because the symbol duration increases for the lower rate parallel subcarriers, the relative amount of dispersion in time caused by multipath delay spread is decreased by introducing a guard time in every OFDM symbol. The ideal channel is suffer from attenuation ,interference and fading due to multi-path propagation and the non-ideal response of channels lead to ISI is eliminated completely. The guard time of the OFDM symbol is cyclically extended to avoid inter carrier interference (ICI).
The concept of OFDM is contributing more advantages in high speed bi-directional wireless data communication. The OFDM technology is used in broad cast systems such as Asymmetric Digital Subscriber Line (ADSL), European Telecommunications standard Institute (ETSI), radio (DAB: Digital Audio broadcasting) and TV (DVB: Digital Video broadcasting-Terrestrial) [2].

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A. Generation of Subcarriers Using the IFFT:

An OFDM signal consists of a sum of subcarriers are modulated by using QAM. An OFDM symbol start at (t - ts) is given asS (t) =Re {exp (j2π (-) (t-ts))}, ts ≤ t ≤ ts+T (2.1)

S (t) =0, t<ts and t> ts+T

Where NS is the Number of subcarriers, T is the Symbol duration and fc is the carrier frequency. The real and imaginary parts correspond to the in phase and quadrature parts of the OFDM signals are multiplied by a cosine and sine of the desired carrier frequency to produce the final OFDM signal. The basic operation of the OFDM modulator is shown in fig.1.S (t) =Re {exp (j 2 π () (t-ts))},ts ≤ t ≤ ts+T (2.2)

S (t) =0, t<ts and t > ts+T

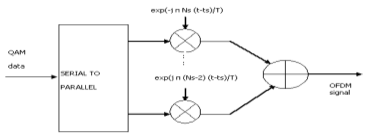

Figure 1. OFDM Modulator [4]

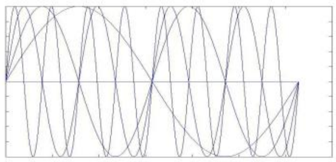

Figure 2: OFDM signal [4]

A one OFDM signal have four subcarriers are shown inFig.2 All subcarriers have the same phase and amplitude is modulated differently for each subcarrier has an exactly integer number of cycles in the interval Tb and the number of cycles between adjacent subcarriers differs by exactly one. This property resembles the orthogonality between the subcarriers. If the j^{th} subcarrier is demodulated by down converting the signal with a frequency of j/T and integrating the signal over T seconds. The intermediate result is a complex carrier is integrated over T seconds. The demodulated subcarrier j integration gives the desired output $d_{j+N/2}$ and all other subcarriers integration is zero because the frequency differences (i-j)/T produce an integer number of cycles within the interval T then the integration result is always zero.

$$\int_{t_s}^{t_s+T} \exp\left(-j2\pi \frac{j}{T}(t-t_s)\right) \sum_{i=\frac{N_s}{2}}^{\frac{N_s}{2}-1} d_{i+\frac{N_s}{2}} \exp\left(j2\pi \frac{i}{T}(t-t_s)\right) dt$$

$$= \sum_{i=\frac{N_s}{2}}^{\frac{N_s}{2}-1} d_{i+\frac{N_s}{2}} \int_{t_s}^{t_s+T} \exp(j2\pi \tfrac{i-j}{T}(t-t_s))\, dt = d_{j+Ns/2}\, T \qquad (2.3)$$

The spectrum of a single symbol is a convolution of a group of Dirac pulses located at the subcarrier frequencies with the spectrum of a square pulse is one for T-second period and zero otherwise [4].

3.Multiple Input Multiple Output

MIMO technology uses the multiple antennas at both the transmitter and receiver to improve communication performance. It has attracted attention in wireless communication because significant increases in data throughput and link range without additional bandwidth or transmits power it achieves higher spectral efficiency and reliability or diversity. MIMO is an important part of modern wireless communication standards such as IEEE 802.11n (Wi-Fi),3GPP Long Term Evolution, WiMAX and HSPA+ [6].

A. MIMO wireless communication

MIMO technology has rapidly gained in popularity over the past decade due to its powerful performance-enhancing capabilities. In Communication wireless channels is impaired predominantly by multi-path fading. Multi-path is the arrival of the transmitted signal at an intended receiver through differing angles, time delays and frequency (i.e., Doppler) shifts due to the scattering of electromagnetic waves in the environment. Consequently, the received signal power fluctuates in space (angle spread), frequency (delay spread) and time (Dopplerspread) through the random superposition of the impinging multi-path components. This random fluctuation in signal level known as fading it can affect the quality and reliability of wireless communication. Additionally, the constraints posed by limited power and scarce frequency bandwidth make the task of designing high data rate, high reliability wireless communication systems extremely challenging.

B. MIMO Channel Model

The MIMO channel model signal is composed by NT transmit and NR receive dimensions are **y** = **H x** + n(3.1)

Where **x**© NTx1 is the transmitted vector, **H**© NTx NR is the channel matrix, **y**© NRx1 is the received vector and **n**© NRx1 is the noise vector. This model represents a single transmission for a communication with multiple transmission signals are indexed with a time-discrete index as **y** (t) = Hs (t) +n (t) Where Hs (t)is the channel time-varying [7].

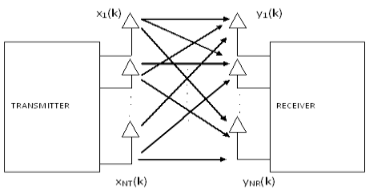

Figure 3. MIMO Chanel model in a scattering environment.

The signals at the transmit (N_T) and receive (N_R) antennas Array vectors are denoted by s(t) = [x1(t), x2(t), ..., xN_T(t)]^T andy(t) = [y1(t), y2(t), ..., yN_R(t)]^T, where $(\cdot)^T$ denotes transposition, xm(t) and ym(t) are the signals at the mth transmit and receive antenna ports. The MIMO radio channel describes theconnection between transmitter and receiver are expressed as

$$\mathrm{H} = \begin{pmatrix} H_{11} & H_{12} & \cdots & H_{1N_T} \\ H_{21} & H_{22} & \cdots & H_{2N_T} \\ \vdots & \vdots & \ddots & \vdots \\ H_{N_R 1} & H_{N_R 2} & \cdots & H_{N_R N_T} \end{pmatrix} \qquad (3.2)$$

Where H_{NM} is the complex transmission coefficient from antennas m at the transmitter to n at the receiver. Moreover, the path gains are correlated depending on spacing, the propagation environment and the polarization of the antenna elements. The relation between the vectors x (t) and **y** (t) are expressed as Y (t) =**H** (t) **x** (t) (3.3)
The two different spatial channel models are correlated and uncorrelated channels. The correlated Channel is widely used in a transmit RT and receiver RR of correlation matrices are an independent identically distributed (i.i.d.) complex Gaussian matrix are used $G \in N_C^{N_T \times N_R}$ and $H = R_R^{\frac{1}{2}} G (R_T^{\frac{1}{2}})^T$ (3.4)
Whereas the uncorrelated channel antenna elements arelocated far away from each other. The channel matrix H is i.i.d. complex Gaussian random variables with zero mean and unit variance is $H \sim N_C^{N_T \times N_R}(0,1)$ (3.5)
The correlation matrices are calculated as $R_R = E\{HH^H\}$ and $R_T = E\{H^H H\}$.Further, they are normalized in such a way that $E\{tr(HH^H)\} = N_R N_T$ $E\{tr(\mathrm{H}^H \mathrm{H})\} = N_R N_T$, as in the i.i.d. case.

4. Training Based Channel Estimation

The training based channel estimation algorithms, pilot tones are known as prior to the receiver are multiplexed along with the data stream for channel estimation. A dynamic estimation of the channel is necessary Pilot-based are widely used to estimate the channel properties and correct the received signal.

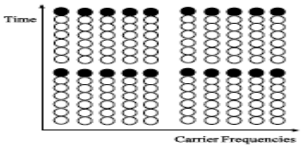

Figure 4. Block pilot

The first kind is a block-type pilot arrangement is shown in Fig 4. This pilot signal assigned to a particular OFDM block sent periodically in time-domain and it is suitable for slow fading radio channels because the training block contains all pilots channel interpolation in frequency domain is not required. Therefore, this type of pilot arrangement is relatively insensitive to frequency selectivity.

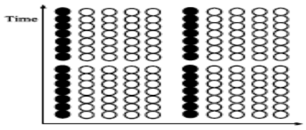

Figure 5 Comb pilot

The second kind is a comb-type pilot arrangement is shown in Fig 5 This pilot arrangement is uniformly distributed within each OFDM block; it has a higher re-transmission rate, better resistance to fast-fading channels. Since some sub-carriers contain the pilot signal and channel response of non-pilot subcarriers will be estimated by interpolating neighboring pilot sub-channels.

1) Least Square Estimator:

A least square estimator is minimizing the square distance between the received signal Y and original signal X then the signal model is $Y=HX+W$, where H is unknown channel and W is Gaussian noise. To minimize the Cost function C (H) =J= {(Y-XH) ^{T}(Y-XH)} = {(Y^T-H^TX^T) (Y-XH)} = {$Y^TY-Y^TXH-H^TX^TY+H^TX^TXH$}. Where $(.)^T$ is the conjugate transpose operator. For minimizations of J differentiate with respect to H and equate to zero

$$\frac{\partial}{\partial H}\{J\} = -X^TY + X^TXH = 0$$

$$\hat{H}_{LS} = X^{-1}Y \qquad (4.1)$$

In general, LS channel estimation technique for OFDM system has low complexity but it suffers from a high mean square error.

2) Minimum Mean Square Error (MMSE) Estimation:

It employs the second-order statistics of the channel conditions to minimize the mean-square error. Let the signal model is $Y=XH+W$ and $\hat{H} = MY$. MSE is expressed as $J(e) = E[(H-\hat{H})^2]$ $= E[(H-\hat{H})^H(H-\hat{H})]$. The orthogonality is to minimize the mean square error vector $e = H - \hat{H}$ has to be set orthogonal by the MMSE equalizer to the estimators input vector Y. $E[(H-\hat{H})Y^H] = 0$, $E[(H-MY)Y^H] = 0$, $E[HY^H] - ME[YY^H] = 0$ and $E[FhY^H] - ME[YY^H] = 0$

If the time domain channel vector h is Gaussian and uncorrelated with the channel noise W, then $FR_{hY} = MR_{YY}$, where $R_{hY} = E[hY^H]$, $R_{YY} = E[YY^H]$, $R_{hY} = E[hY^H] = E[h(XFh+w)^H]$ and $R_{hY} = R_{hh}F^HX^H$. h is uncorrelated with w because $E[hw^H] = 0$. $R_{YY} = E[YY^H] = E[(XFh+w)(XFh+w)^H]$ And $R_{YY} = XFR_{hh}F^HX^H + \sigma^2 I_N]$. Where σ^2 is the variance of noise; $M = FR_{hY}R_{YY}^{-1}$ and $\hat{H} = FR_{hY}R_{YY}^{-1}Y$ then the time domain MMSE estimate of h is given by $\hat{h}_{MMSE} = R_{hY}R_{YY}^{-1}Y$ (4.2)

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3) Least Mean Square (LMS) Algorithm:

The Least Mean Square (LMS) algorithm estimates the gradient vector and it incorporates an iterative procedure makes successive corrections to the weight vector in the direction of the negative gradient vector leads to the minimum mean square error is simple and it does not require correlation function.

A block diagram of the proposed channel estimator using the LMS algorithm of the time-domain preamble signal and aN-tap adaptive filters are shown in Fig. 6.and fig.7.

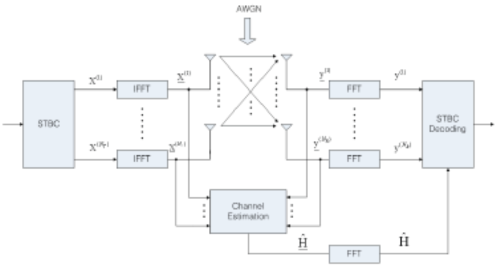

Figure 6. a block diagram of adaptive algorithm

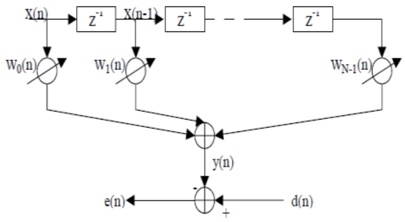

Figure 7. LMS algorithm for channel estimation.

$y(n)=\sum_{i=0}^{N-1} w_i(n)x(n-i)$ And *e (n) =d (n)-y (n)* (4.3)

The LMS algorithm changes the filter tap weights; e (n) is minimized in the mean-square sense. When the processes x (n) & d (n) are jointly stationary and converge to a set of tap-weights averages are equal to the Wiener-Hopf solution then the conventional LMS algorithm is a stochastic implementation of the steepest descent recursion algorithm is simply replaces the cost function. $\xi = E(e^2(n))$ Is an instantaneous coarse estimate $\hat{\xi} = (e^2(n)$. Substituting $\hat{\xi} = (e^2(n)$ in the steepest descent recursion ζ, is $\overline{W}(n+1) = \overline{W}(n) - \mu\nabla e^2(n)$ (4.4)

Where $\overline{W}(n+1) = [w_0(n), w_1(n), \ldots\ldots, w_{N-1}(n)]^T$

$$\nabla = \begin{bmatrix} \frac{\partial}{\partial w_0} & \frac{\partial}{\partial w_1} & \cdots & \frac{\partial}{\partial w_{N-1}} \end{bmatrix}^T$$

The i^{th} element of the gradient vector $\nabla e^2(n)$ is

$$\frac{\partial e^2(n)}{\partial w_i} = 2e(n)\frac{\partial e\ \ (n)}{\partial w_i}$$

$$= -2e(n)\frac{\partial y(n)}{\partial w_i} = -2e(n)x(n-i) \quad (4.5)$$

$\nabla e^2(n)$ = -2*e* (*n*) $\overline{x}(n)$ .Where $\overline{x}(n) = [x(n), x(n-1), \ldots\ldots, x(n-N+1)]^T$ .Finally $\overline{W}(n+1) = \overline{W}(n) + 2\mu e(n)\overline{x}(n)$ (4.6) The Advantages of LMS recursion algorithm are implementation, Stable and robust performance against different signal conditions. The architecture of the LMS algorithm for the channel estimation is shown in fig.8.

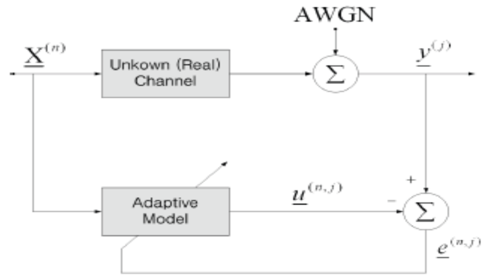

Figure 8. Architecture of the LMS algorithm

The output of the LMS algorithm is given by

$\underline{u}^{(n,j)}$(k) = $\underline{h}^{\wedge(n,j)H}(k)\underline{x}^{(n)}$(k) (4.7)

Where $\underline{h}^{\wedge(n,j)} \triangleq [\underline{h^{\wedge}{}_L}^{(n,j)}, \dots \dots \dots .., \underline{h_1}^{\wedge(n,j)}]^T$ is an estimated channel vector including channel taps for the (n, j) The transmit/receive antenna, $\underline{x}^{(n)}$(k) is the corresponding time domain input vector for the nth transmit antenna. The estimated channel vector $\underline{h}^{\wedge(n,j)}$(k) is given by

$\underline{h}^{\wedge(n,j)}$(k+1) = $\underline{h}^{\wedge(n,j)}$(k) + 2$\mu\underline{x}^{(n)}$(k) $\underline{e}^{(n,j)*}$(k) (4.8)

With estimation error $\underline{e}^{(n,j)}$(k) = $\underline{y}^{j}$(k) - $\underline{u}^{(n,j)}$(k) (4.9)

Where μ (> 0) is a step-size parameter to control a convergence speed of the algorithm and * is the complex conjugation. An output of the jth receive antenna can be given by $\underline{y}^{j}$ (k) = $\underline{h}^{(n,j)H}$ (k) $\underline{x}^{(n)}$ (k) + $\sum_{i=1,i\neq n}^{N_T} \underline{h}^{(i,j)H}(k)\underline{x}^{(i)}$(k)+$\underline{v}^{j}$ (k) When L is unknown, a scheme of estimating the number of taps or the tap vectors with the size of the number is expected to sufficiently greater than L.

4) LMS Based Channel Estimation:

The MIMO-OFDM system employs multiple transmit (N_T) and receive (N_R) antennas respectively. In the time-domain a received signal for the jth receive antenna at discrete sample index k can be modelled as

$\underline{r}^{(j)}$(k) = $\underline{h}^{(n,j)H}$(k)$\underline{s}^{(n)}$(k) + $\sum_{i=1,i\neq n}^{N_T} \underline{h}^{(i,j)H}(k)\underline{s}^{(i)}$(k)+$\underline{v}^{j}$ (k); n=1,2,…..N_T And j=1, 2…N_{TR} (4.10)

Where $\underline{h}^{(n,j)} = [\underline{h_L}^{(n,j)}, \dots \dots \dots .., \underline{h_1}^{(n,j)}]^T$ is a channel vector (size L) including (n,j)th channel taps, n and j are indices of the transmit and receive antennas respectively. L is the number of taps,$\underline{s}^{(n)}$(k)) is a time-domain signal mapped to the nth transmit antenna and $\underline{v}^{j}$ (k) is an additive white Gaussian noise (AWGN) with zero mean and variance σ^2. $\sum_{i=1,i\neq n}^{N_T} \underline{h}^{(i,j)H}(k)\underline{s}^{(i)}$(k) Serves as an interference signal when an (n, j)th channel $\underline{h}^{(n,j)}$ is estimated. In this paper the channel remains unchanged within one OFDM symbol Since, the signal after passing through the channel has a convolution form of the transmission signal and the tap vectors are expressed in order of L to 1, i.e., in reverse order.

B. MIMO OFDM System

A block diagram of the MIMO OFDM system is shown in fig. 4.6. The Alamouti Space Time Block Coding (STBC) scheme has full transmit diversity gain and low complexity decoder with the encoding matrix represented as $X = \begin{bmatrix} X_1 & -X_2^* \\ X_2 & X_1^* \end{bmatrix}$

Where $X_1 = (X[0], -X^*[1], X[2], -X^*[3], \dots.., X[0], -X^*[N-1])$ and

$X_2 = (X[1], X^*[0], X[3], X^*[2], \dots.., X^*[N-2])$ (4.11)

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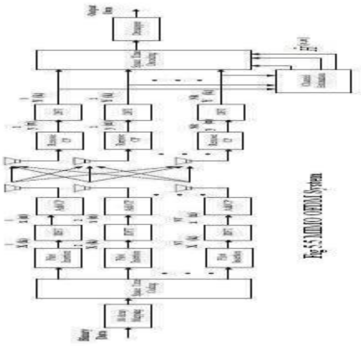

Figure 9. A MIMO OFDM System

The vectors X_1 and X_2 are modulated using the inverse fast Fourier transform (IFFT) and after adding a cyclic prefix as a guard time interval. The two modulated blocks x_1^g and x_2^g are generated and transmitted by the first and second transmit antennas respectively. The received signal will be the convolution of the channel and transmitted signal and the channel is static during an OFDM block at the receiver side after removing the cyclic prefix the FFT output as the demodulated received signal is expressed as

$$\begin{bmatrix} Y_1 \\ Y_2 \\ \vdots \\ Y_{N_R} \end{bmatrix} = \begin{bmatrix} H_{1,1} & H_{1,2} & \cdots & H_{1,N_T} \\ H_{2,1} & H_{2,2} & \cdots & H_{2,N_T} \\ \vdots & \vdots & \ddots & \vdots \\ H_{N_R,1} & H_{N_R,2} & \vdots & H_{N_R,N_T} \end{bmatrix} \begin{bmatrix} X_1 \\ X_2 \\ \vdots \\ X_{N_T} \end{bmatrix} + \begin{bmatrix} W_1 \\ W_2 \\ \vdots \\ W_{N_R} \end{bmatrix} \qquad (4.12)$$

Where [$W1$, $W2$…… W_{N_R}] is the AWGN and $H_{m,n}$ is the channel gain between the mth receive and nth transmit antenna pair. The nth column of **H** is the spatial signature of the nth transmit antenna across the receive antenna array. The channel information is known at the receiver, Maximum Likelihood (ML) detection is used for decoding of received signals for two antenna transmission systems can be written as

$$\tilde{s}[2k] = \sum_{i=1}^{N_R} H_{i,1}^*[2k]Y_i[2k] + H_{i,2}[2k]Y^*_i[2k+1] \quad \tilde{s}[2k+1] = \sum_{i=1}^{N_R} H_{i,2}^*[2k+1]Y_i[2k] - H_{i,1}[2k+1]Y^*_i[2k+1] \qquad (4.13)$$

Where k = 0, 1, 2…… (N/2)-1. The channel gains between two adjacent sub-channels $H_{i,1}[2k]=H_{i,1}[2k+1]$ and $H_{i,2}[2k]=H_{i,2}[2k+1]$ are equal.

1) MIMO OFDM Channel Estimation

The LS and MMSE channel estimations for MIMO-OFDM System between nth transmitter and mth receiver antennas are

$$\hat{H}_{LS}^{(n,m)} = (X^{(n)})^{-1} Y^{(m)} \quad (4.14)$$

$$\hat{H}_{MMSE}^{(n,m)} = F R_{hY} R_{YY}^{-1} Y^{(m)} \quad (4.15)$$

Where $R_{hY} = R_{hh}^{(m,n)} F^H (X^{(n)})^H$ and $R_{YY} = X^{(n)} F R_{hh}^{(n,m)} F^H (X^{(n)})^H + \sigma^2 I_N$

Where n = 1, 2...N_T, m = 1, 2...$N_R$and N_T, $N_R$are the number of transmit and receive antennas, X(n) is an N X N diagonal matrix whose diagonal elements correspond to the pilots of the n^{th} transmit antenna and Y(m) is received vector at receiver antenna m.

5. Blind Channel Estimation

The blind channel estimation is evaluating the statistical information of the channel and certain properties of the transmitted signals. It is only applicable to slowly time-varying channels due to a long data record. The Blind channel estimation of Singular Value Decomposition (SVD) iteration method.

A. Singular Value Decomposition(SVD) Iteration Method:

The system model has Q subcarriers and uses the numbered k_0 to k_0+D-1 for data block. The remaining Q-D un-modulated carriers are referred to as virtual carriers (VCs). The complex baseband data block can be written as $d_i(n) = [d_i(n,k_0), d_i(n,k_0+1),...., d_i(n,k_0+D-1)]$ (5.1)

Where i is the transmit antenna index from 1 to M. The matrices W (i) and W are associated with the IFFT as given byW (i) $\overset{\Delta}{=} \frac{1}{\sqrt{N}} \left[w_N{}^{ik_0}, w_N{}^{i(k_0+1)}, \ldots\ldots, w_N{}^{i(k_0+D-1)} \right]$ and

$$W \overset{\Delta}{=} [W(N-1)^T, \ldots, W(0)^T, W(N-1)^T, \ldots, W(N-P)T]^T \quad (5.2)$$

$$W \overset{\Delta}{=} I_J \otimes W \; W \otimes I_{Mt} \quad (5.3)$$

The received data can be written in matrix form as

$$r(n) = Hs(n) + \eta(n) = HWd(n) + \eta(n) \overset{\Delta}{=} \Xi d(n) + \eta(n) \quad (5.4)$$

Where *d (n)* is the source data before OFDM modulation, s (n) is the transmitted data after OFDM modulation and η (n) is a spatially and temporally uncorrelated complex Gaussian noise vector with the zero mean vector and the covariance matrix σ_η^2 $I_{(JK-1)N}$. J is the collecting OFDM symbols, *L* is the upper bound on the orders of these channels. *K=Q+P*. For received data r *(n)* on n time sample data autocorrelation matrix defined as follows
R (n) =λR (n-1) + (1-λ) r (n) rH (n) (5.5)
Where $0 \le \lambda \le 1$is a forgetting factor. When the autocorrelation matrix R (n) of the received signal vector r (n) is diagonalized through the eigen value decomposition, partitioning the eigenvectors U into the vectors spanning a signal subspace span (Us) and the vectors spanning a noise subspace (U_n) as

$$U = [U_S / U_n] = \left[u_1 \ldots\ldots\ldots, u_{JDM}, / u_{JDM+1} \ldots\ldots\ldots, u_{(JK-L)N} \right] \quad (5.6)$$

The orthogonal relationship is $u_k{}^H \Xi = 0$

For all $k \, \{n\}_{n=JDM+1}^{(JQ-L)N}$ (5.7)

Where M and N are number of transmitting and receiving antennas. Thus, when a MIMO channel is estimated by the orthogonal relationship are estimates only U_n of the eigenvectors spanning the noise subspace which are obtained by the Eigen value decomposition of R (n). Partitioning the eigenvector estimate u_k as given in

$$u_k = \begin{bmatrix} \hat{v}_1^{(k)} \\ \hat{v}_2^{(k)} \\ \vdots \\ \hat{v}_{(JK-L)}^{(k)} \end{bmatrix} \quad (5.8)$$

The matrix $\hat{V}_K$ as follows.

$$\hat{V} = \begin{bmatrix} \hat{v}_1^{(k)} & \hat{v}_2^{(k)} & \cdots & \hat{v}_{JK-L}^{(k)} & 0 & \cdots & 0 \\ 0 & \hat{v}_1^{(k)} & \hat{v}_2^{(k)} & \cdots & \hat{v}_{JK-L}^{(k)} & \cdots & 0 \\ \vdots & \vdots & \ddots & \ddots & \ddots & \ddots & \vdots \\ 0 & \cdots & 0 & \hat{v}_1^{(k)} & \hat{v}_2^{(k)} & \cdots & \hat{v}_{JK-L}^{(k)} \end{bmatrix} \quad (5.9)$$

The matrix $\Psi = \sum_{k=JDM+1}^{(JK-L)} \hat{V}_k (I_J \otimes W^* W^T) \hat{V}_K^H$ (5.10)

Where Wfrom an eq. (5.10) write a cost function $\sum_{i=1} h_i{}^H \Psi h_i$ equivalent to C (H). By imposing the constraints such as $\|h_i\|_2 l2$ norm for $1 \le i \le M$, where M is number of transmitting antennas to avoid trivial solutions to estimate $\hat{H}$ of the channel coefficient matrix H is obtained by $\hat{H} = \arg\min_{\|h_i\|_2} (\sum_{i=1}^{M} h^H{}_i \Psi h_i)$, when h_i satisfying $\partial(\sum_{i=1}^{M} (h_i{}^H \Psi h_i + \lambda_i (1 - \|h_i\|_2^2))) / \partial h_i^H = 0$ with a Lagrange multiplier λ_i, the estimates $\hat{h}_i$ of the channel response vectors h_i are the eigenvectors associated with the smallest *M* eigen values of Ψ. Since the orthogonal relationship $u_k{}^H \Xi = 0$ can be rewritten as $(I_J \otimes W^T) \hat{v}_k^H h_i = 0$ for $1 \le i \le M$, we find h_i closely orthogonal to column vectors of $\hat{V}_k (I_J \otimes W^*)$ with an estimate $\hat{v}_k$ of v_k . In addition, $[h_1 \quad h_2 \quad \cdots \quad h_M]$ should be linearly independent. If the eigenvectors associated with the smallest Meigen values of Ψare denoted as $[\hat{h}_1 \quad \hat{h}_2 \quad \cdots \quad \hat{h}_M]$ respectively, the estimated channel coefficient matrix $\hat{H}$ as $\hat{H} = [\hat{h}_1 \quad \hat{h}_2 \quad \cdots \quad \hat{h}_M]$.

6. Semi-Blind Channel Estimation

Semi blind channel estimation consists of combining pilot and blind channel estimations and it allow a significant reduction in number of pilot symbols, the reduce error in severe reception conditions. The received data can be written in matrix form as $r = \mathcal{H}d + z = \mathcal{H}\mathcal{F}^H X + z = \mathcal{g}X + z$(6.1)

Where received signal $r = [[r(0)]^T, \dots, [r(N-1)]^T]^T$, N is number of subcarriers transmitted signal $d = [s(N-L), \dots, (s(N-1), s(0), \dots, s(N-1)]^T$,

$\mathcal{F} = [F(N-L), \dots\dots, F(N-1), F(0), \dots\dots, F(N-1)]$

Where F is Discrete Fourier Transform (DFT) matrix is given as $F(i) = \frac{1}{\sqrt{N}}[e^{-j2\Pi i(0)/N}, \dots\dots, e^{-j2\Pi i(N-1)/N}]^T$ and $F = [F(0), \dots, F(N-1)]$. The channel matrix is given as $\mathcal{H} = \begin{bmatrix} h(L) & \cdots & h(0) & 0 & \cdots & 0 \\ \vdots & \ddots & \vdots & \vdots & \ddots & \vdots \\ 0 & \cdots & 0 & h(L) & \cdots & h(0) \end{bmatrix}$(6.2)

Where z is an additive Gaussian noise vector. Then, the statistical covariance matrix of r is written as

$$R_r = E\{rr^H\} = \mathcal{g}E\{XX^H\}\mathcal{g}^H + \sigma_z^2 I_{NN_R} = \mathcal{g}R_x\mathcal{g}^H + \sigma_z^2 I_{NN_R}$$

Where N_R is number of received antennas. If $\mathcal{g}$ is a full column rank, the rank of $\mathcal{g}R_x\mathcal{g}$ is N. since the rank of R_r has full rank, the covariance matrix R_r has two mutually orthogonal subspaces, i.e., a signal subspace with dimension N and a noise subspace with dimension d=NN_R-N. The noise eigenvectors are denoted by V_i where$i = 1,2, \dots, d$. Using the techniques in the standard subspace method is given by

$V_i^H \mathcal{g} = 0, i = 1,2, \dots, d$ (6.3)

Where V_i spans the left null space of $\mathcal{g}$. However, R_r is estimated over N_b OFDM blocks as $\widehat{R_r} = \frac{1}{N_b}\sum_{i=1}^{N_b} r^{(i)}[r^{(i)}]^H$. Thus, when a channel is estimated by the orthogonal relationship is estimated $\widehat{V_i}$ of the eigenvectors spanning the noise subspace are obtained by the eigenvalue decomposition of $\widehat{R_r}$is available in practice. $\widehat{V_i}$ Is divided into blocks as $\widehat{V_i} = [\hat{v}_i^T(0), \dots, \hat{v}_i^T(N-1)]^T$ where $\hat{v}_i(m)$ is $N_R \times 1$ vector and $h = [h^T(0), \dots, h^T(L)]^T$ as the $N_R(L+1) \times 1$ channel impulse response vector. Then, the eq. (6.3) can be expressed equivalently as $\hat{V}_i^H h = 0_{N\times 1}$ (6.4)

Where $\hat{V}_i = \begin{bmatrix} \hat{v}_i(0) & \cdots & \hat{v}_i(N-1) \\ \vdots & \ddots & \vdots \\ \hat{v}_i(L) & \cdots & \hat{v}_i(N-1+L)_N \end{bmatrix}$

By stacking across d vectors from the noise subspace, then the eq. (6.4) is generalized as $\hat{V}_i h = 0_{dN\times 1}$ (6.5)

Where $\hat{V} = [\hat{V}_1, \dots, \hat{V}_d]$. Then, the weighting factor c (i) is the orthogonality between vectors $b \epsilon V_{\mathbb{C}}$. Using the weighting factor c (i) a blind channel estimation algorithm based on the weighted vector of an initial decision matrix D through the existing channel estimation method. $D_{ii} = \sum_{i=0}^{L} H(l)\, W^{-l(i-1)}$. Where i=1, 2… M, W=$e^{\frac{j2\pi}{M}}$.

The weighting factor c (i) between $\hat{V}_i$and D as follow $c(i) \coloneqq 1 - \|cos\theta_c(\hat{v}(i), \mathcal{D})\|_2$, (6.6)

$cos\theta_c(\hat{v}(i), \mathcal{D} = \frac{(\hat{v}(i).\mathcal{D})_c}{\|\hat{v}_i\|_2 \cdot \|\mathcal{D}\|_2}$ (6.7)

Where $\hat{V}_i$ denotes the i^th column vector of the matrix $\hat{V}$ as

$\bar{V} = \{\bar{V}(i) | \bar{V}(i) = c(i)\hat{V}(i), 1 \le i \le dN\}$ (6.8)

Then reconstruct the eq. (6.5) as follow $\bar{V}^H h = 0_{dN\times 1}$ (6.9)

The devising semi blind channel estimation consists of the pilot-based least-square and the subspace based blind channel estimators is given as cost function:

$C_{semi} = \lambda\|Y - Wh\|^2 + (1-\lambda)h^H\bar{V}\bar{V}^H h$ (6.10)

Where λ is a scalar weight coefficient determines the contribution of the pilot and the subspace-based blind estimators of the cost function C_{semi} and λ Є [0, 1]. Taking the derivative of the cost function with respect toh^H , then the semi-blind channel estimator is obtained.

7. Simulation Results

MIMO-OFDM system with 2 transmits and 2 receive antenna simulations are provided to demonstrate the performance of the channel estimation is measured in terms of the Mean Square Error (MSE). The transmitted symbols are drawn from the QAM 16 constellation through simulations of an OFDM system, where the number of subcarrier is N=128 and a cyclic prefix length is 16. AWGN and Rayleigh fading channels are considered through the whole evaluation.

$$MSE = \left[\frac{(h - \overline{h})^2}{|h|^2} \right]$$

A.Training Sequence Method:

1) MSE by using LSE method:

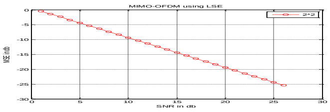

Figure 10. Channel Estimation Using LSE

2) MSE by using MMSE method:

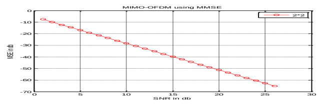

Figure 11. Channel Estimation Using MMSE

3) MSE by using LMS method:

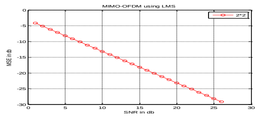

Figure 12. Channel Estimation Using LMS

B. Blind Channel Estimation Method:

1) MSE by using SVD method

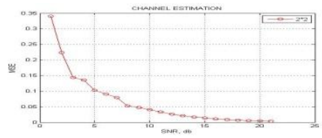

Figure 13. Channel Estimation Using SVD

2) MSE by using SVD with Iteration method

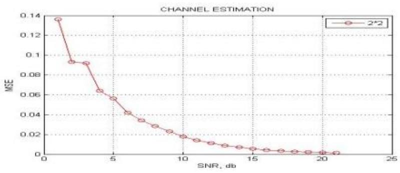

Figure 14. Channel Estimation Using SVD-Iteration

C. Semi-Blind Channel Estimation Method:

1) MSE by using Semi-Blind method

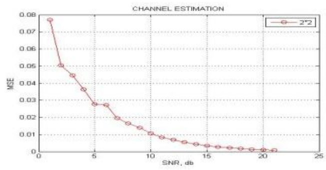

Figure 15. Channel Estimation Using Semi Blind

Semi-blind channel estimation method has the relatively low estimation error comparing with the Blind channel estimation method. It is able to achieve the better performance as compared with other algorithms.

8. Conclusion

We evaluated the performance of OFDM system on the basis of MIMO technique and focused on the two Transmit and two Receive antennas configurations under both AWGN and Rayleigh fading channel. STBC coding is applied in MIMO OFDM system in which carrier allocation are separately processed by simple IFFT type operation. From the simulation results, it is found that Semi Blind channel estimation method reduces error when comparing with other algorithms. In Semi Blind channel estimation error performance is better than Blind channel and Training Based channel estimations. Overall MIMO SFBC CI-OFDM system outperforms MIMO SFBC OFDM significantly when system is interrupted by the HPA nonlinearity or NBI under both AWGN and Rayleigh fading channel. Therefore, the MIMO SFBC CI OFDM method can be further applicable to the any kinds of MIMO type multi-carrier communication systems with many sub-carriers. From the computation procedure above, we can note that Semi blind methods can be used to identify the channel, no matter whether the MIMO-OFDM system is homogeneous or not.

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