# JOINT PILOT-AIDED AND BLIND DECISION-DIRECTED CHANNEL ESTIMATION FOR MIMO-OFDM SYSTEM

Lin Tang
Shanghai Research Center
for Wireless Communications
Shanghai, China

Mosa Ali Abu-Rgheff
Mobile Communications Research
University of Plymouth
Plymouth, uk

## ABSTRACT

In the conventional pilot-symbol-aided channel estimation for the MIMO-OFDM system, the pilots should be orthogonal in frequency, time, code, or spatial domain to mitigate the inter-antenna interference. In the frequency domain multiplexed (FDMed) pattern, the training pilots transmitted from different antennas conventionally occupy different sub-carriers exclusively within the same OFDM symbol duration. However, this method dramatically increases the pilot overhead as the number of the transmit antennas increases. In this paper a joint pilot-aided and blind decision-directed channel estimation (PA-DDCE) scheme is proposed to achieve a good tradeoff between the pilots overhead and the system performance. The proposed scheme utilizes the initial estimated channel obtained by the conventional channel estimation to detect the transmitted data symbol in the middle sub-carrier between any two adjacent pilot sub-carriers. Then these symbols are regarded as the "extra pilot symbols" and are applied to estimate the channel on the whole frequency band. From our simulation results the proposed PA-DDCE scheme can outperform the conventional channel estimation significantly in both the BER and the throughput performances.

## I INTRODUCTION

By using the multiple antennas at both the transmitter and the receiver, the orthogonal frequency division multiplexing (OFDM) system can significantly increase the channel capacity, which makes it to be an attractive candidate for the high-data-rate wireless applications. Such systems are called the multiple-input multiple-output (MIMO) OFDM systems. To correctly demodulate the received data symbols, the channel estimation plays an important role to the system performance greatly. In practice the channel state information (CSI) is estimated from known pilot symbols and subsequently used for the data detection. Therefore the system performance depends on the quality of channel estimation and hence the number of pilot symbols. Estimation of the MIMO fading channel is a major challenge for the multiple-antenna wireless communication systems. When the number of antennas increases, the accurate channel estimation becomes more difficult due to the increase parameters to be estimated. In the conventional channel estimation, the pilots should be orthogonal to avoid the inter-antenna interference.

For the MIMO-OFDM systems with Frequency Domain Multiplexed (FDMed) pilots, the training pilots transmitted from different antennas occupy different sub-carriers exclusively within the same OFDM symbol duration. However, this method radically increases the pilot overhead when the number of the transmit antennas increase, which diminishes the system spectrum efficiency. The semi-blind and blind channel estimators [1] [2] have been proposed to reduce the pilot overhead; however, these methods are not practical due to their poor performance and high complexity. Motivated by this fact, a joint pilot-aided and blind decision-directed channel estimation (PA-DDCE) scheme is proposed in this paper to achieve a good tradeoff between the pilots overhead and the system performance. In the previous literatures [3] [4], the decision-directed channel estimation is proposed to support the high terminal velocities. The contribution of our proposed scheme is to design a pilot structure for the MIMO-OFDM system which has a relative small pilot overhead but still guarantees the accuracy of the channel estimation. At first the data symbols with the orthogonal format are transmitted in the middle of any two adjacent pilot sub-carriers. After that with the knowledge of the modulation alphabet of the data symbols, the blind hard decision on the symbols can be made and then the CSI estimate performance can be improved with slight increase of the pilot overhead versus the conventional orthogonal pilot structure. The pilot structure of the PA-DDCE scheme with two alternative orthogonal data formats have also been proposed in this paper, they are named as null-data format and space frequency block code (SFBC)-data [5] format.

The rest of the paper is organized as follows. Section II introduces the MIMO-OFDM system model and presents the conventional pilots-aided channel estimation in the MIMO-OFDM system. Then section III proposes our PA-DDCE scheme with two orthogonal data formats. The simulation results are given in section IV and finally we conclude our paper in section V.

## II THE MIMO-OFDM SYSTEM

An OFDM system with two transmitters and two receivers is shown in Fig.1.

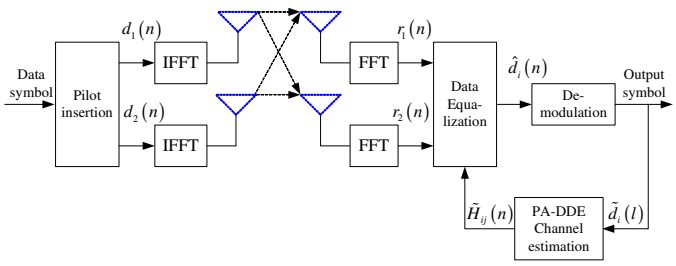


Figure 1: MIMO-OFDM system with PA-DDCE scheme

1-4244-1144-0/07/$25.00©2007 IEEE

Authorized licensed use limited to: Anthony Mcgrath. Downloaded on March 26,2024 at 18:30:04 UTC from IEEE Xplore. Restrictions apply.
Petitioner Samsung Ex-1008, 0001

At one OFDM symbol duration, the transmitted symbols $\{d_i(n) : n = 0, 1, \ldots N-1\}$, for transmit antenna $i = 1, 2$, inserted with pilots on each antenna form an OFDM block, where $N$ is the total number of OFDM sub-carriers. The two transmit antennas simultaneously transmit the OFDM signals modulated by inverse fast Fourier transform (IFFT). At the receiver side, the signal demodulated by fast Fourier transform (FFT) can be expressed as

$$r_j(n) = \sum_{i=1}^{2} H_{ij}(n) d_i(n) + w_j(n), \tag{1}$$

where $H_{ij}(n)$ is the channel frequency response for the $n$th sub-carrier of the OFDM symbol, corresponding to the $i$th transmit antenna and the $j$th receive antenna. $w_j(n)$ denotes the complex additive white Gaussian noise (AWGN) with zeros mean and variance $\sigma_n^2$, on the $j$th receive antenna.

## III CONVENTIONAL CHANNEL ESTIMATION

In the conventional channel estimation with FDMed pilots of the MIMO-OFDM system, to keep the pilot orthogonality, a training pilot from one antenna is transmitted at a sub-carrier while the remaining transmit antennas are idle at the same sub-carrier, which is shown in Fig.2, where $p_1$ and $p_2$ are pilot symbols at the first and the second antenna respectively.

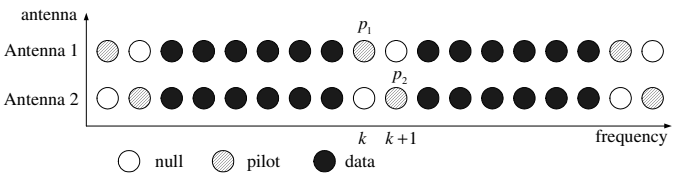


Figure 2: Pilot structure of the conventional channel estimation

For the MIMO-OFDM system with two transmit antennas and two receive antennas, by using Least Square (LS) method [6] the estimated channel frequency response on the pilot sub-carrier $k$ and $k+1$ are given by

$$\begin{aligned} \hat{H}_{11}(k) &= r_1(k)/p_1(k) \\ \hat{H}_{12}(k) &= r_2(k)/p_1(k) \\ \hat{H}_{21}(k+1) &= r_1(k+1)/p_2(k+1) \\ \hat{H}_{22}(k+1) &= r_2(k+1)/p_2(k+1). \end{aligned} \tag{2}$$

The final estimated channel frequency response on the whole band sub-carriers is given by

$$\hat{H}_{ij}(n) = interp\left(\hat{H}_{ij}(k)\right), \tag{3}$$

where

| | |
|---|---|
| $k$ | pilot frequency subscript (sub-carrier indices) |
| $p_i$ | transmitted pilot on $i$th transmitter antenna |
| $H$ | estimated channel frequency response |
| interp() | interpolation. |

The performance of the conventional channel estimation for the MIMO-OFDM system greatly depends on the pilot number. With more pilot symbols, the pilot spacing could be smaller thus implies less interpolation error. However, the system spectrum efficiency is diminished and the system throughput is decreased due to the larger pilot overhead. To overcome this problem, we propose a novel PA-DDCE scheme with two alternative pilot structures in the following section.

## IV PROPOSED PA-DDCE SCHEME

In this section, the basic principle of the PA-DDCE scheme is introduced at first. Then we study the proposed scheme for two kinds of the orthogonal data symbol formats, namely null-data format and SFBC-data format.

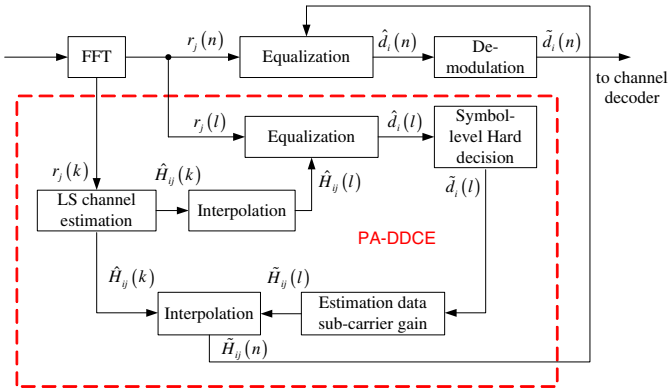


Figure 3: Principle of PA-DDCE scheme

The principle of the PA-DDCE scheme is illustrated in Fig.3, and the algorithm procedure is described as followed. Step 1: LS channel estimation to calculate the estimated channel frequency response $\hat{H}_{ij}(k)$ on the pilot sub-carriers. Step 2: Linear or other interpolation methods to obtain the estimated channel frequency response on the non-pilot sub-carriers $\hat{H}_{ij}(l)$ , where $l$ is defined as the middle sub-carriers between any two adjacent pilot sub-carriers. Step 3: Equalize the received data symbols $r_i(l)$ with the previous estimated channel frequency response $\hat{H}_{ij}(l)$ in step 2 and get the equalized data symbol $\hat{d}_i(l)$. Step 4: Hard decision of the equalized data symbol by demodulation and re-modulation, finally obtain the constellation points $\tilde{d}_i(l)$. Step 5: Calculate the new data sub-carrier channel frequency response $\tilde{H}_{ij}(l)$ by using the detected data symbols. Step 6: Linear or other interpolation with both $\hat{H}_{ij}(k)$ and new $\tilde{H}_{ij}(l)$ updated in step 5 to obtain the final estimated channel frequency response $\tilde{H}_{ij}(n)$.

In the followed paragraphs, the detailed descriptions of the step 3 and step 5 in the PA-DDCE scheme will be presented with regard to the null-data format and the SFBC-data format.

### A *PA-DDCE Scheme with Null-Data Format*

The pilot structure of the PA-DDCE scheme with null-data format is shown in Fig.4. The difference to the conventional one is that in the middle of any two adjacent pilot symbols, a null-data format is inserted. When the data symbol $d_1$ on the sub-carrier $l$ is transmitted at the first antenna, the second antenna is idle.

Authorized licensed use limited to: Anthony Mcgrath. Downloaded on March 26,2024 at 18:30:04 UTC from IEEE Xplore. Restrictions apply.
Petitioner Samsung Ex-1008, 0002

Vice versa, when the data symbol $d_2$ on the sub-carrier $l+1$ is transmitted at the second antenna, the first antenna is idle.

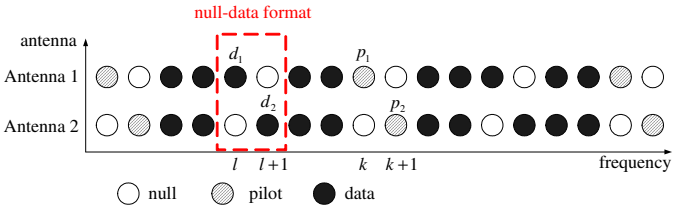


Figure 4: Pilot structure of PA-DDCE scheme with null-data format

Both the received data symbols on the sub-carrier $l$ and $l+1$ can be shown by

$$\mathbf{r} = \mathbf{H}\mathbf{d}_n + \mathbf{w}, \tag{4}$$

where we have

$$\mathbf{r} = \begin{bmatrix} r_1(l) & r_1(l+1) \\ r_2(l) & r_2(l+1) \end{bmatrix}, \tag{5}$$

$$\mathbf{H} = \begin{bmatrix} H_{11}(l) & H_{21}(l+1) \\ H_{12}(l) & H_{22}(l+1) \end{bmatrix}, \tag{6}$$

and

$$\mathbf{d}_n = \begin{bmatrix} d_1(l) & 0 \\ 0 & d_2(l+1) \end{bmatrix}. \tag{7}$$

The receiver adopts the two-branch maximal-ratio receiver combining (MRRC) and the transmitted data symbol $\hat{d}_1(l)$ $\hat{d}_2(l+1)$ on the sub-carrier $l$ and $l+1$ are computed by zero forcing (ZF) equalizer, i.e.,

$$\hat{d}_1(l) = \frac{\hat{H}_{11}^*(l)\, r_1(l) + \hat{H}_{12}^*(l)\, r_2(l)}{\left|\hat{H}_{11}(l)\right|^2 + \left|\hat{H}_{12}(l)\right|^2}, \tag{8}$$

and

$$\hat{d}_2(l+1) = \frac{\hat{H}_{21}^*(l+1)\, r_1(l+1) + \hat{H}_{22}^*(l+1)\, r_2(l+1)}{\left|\hat{H}_{21}(l+1)\right|^2 + \left|\hat{H}_{22}(l+1)\right|^2}, \tag{9}$$

where $()^*$ denotes the complex conjugation and $|()|$ is the absolute value. $\hat{H}_{ij}(l)$ and $\hat{H}_{ij}(l+1)$ are the estimated channel frequency response by the conventional channel estimation with the linear or other interpolation methods, which are obtained in step 2.

With the pre-known modulation alphabets, the symbol-level hard-decision is made on the transmitted data symbols $\hat{d}_1(l)$ and $\hat{d}_2(l+1)$ calculated in step 3. Assuming the error-free of the symbol hard decision, the re-modulated data symbol $\tilde{d}_1(l)$ and $\tilde{d}_1(l+1)$ on the constellation points can be regarded as extra "pilot symbols", which means the pilot spacing is reduced to 1/2 of that in the conventional channel estimation. A new channel estimation on the $l$th and $l+1$th sub-carriers can thus be performed with these new "pilot symbols", which leads to

$$\tilde{\mathbf{H}}_n = \left(\tilde{\mathbf{d}}_n^H \tilde{\mathbf{d}}_n\right)^{-1} \tilde{\mathbf{d}}_n^H \mathbf{r}, \tag{10}$$

for the new channel estimate, where we have

$$\tilde{\mathbf{H}}_n = \begin{bmatrix} \tilde{H}_{11}(l) & \tilde{H}_{21}(l+1) \\ \tilde{H}_{12}(l) & \tilde{H}_{22}(l+1) \end{bmatrix}, \tag{11}$$

and

$$\tilde{\mathbf{d}}_n = \begin{bmatrix} \tilde{d}_1(l) & 0 \\ 0 & \tilde{d}_2(l+1) \end{bmatrix}. \tag{12}$$

The update of the estimated channel frequency response is carried out by processing the previous estimated channel frequency response and the new ones through the interpolation. Thus the final enhanced estimated channel frequency response is given by

$$\hat{H}_{ij}(n) = interp\left(\hat{H}_{ij}(k), \tilde{H}_{ij}(l)\right). \tag{13}$$

### *B PA-DDCE Scheme with SFBC-Data Format*

The pilot structure of the PA-DDCE scheme with the SFBC-data format is shown in Fig.5.

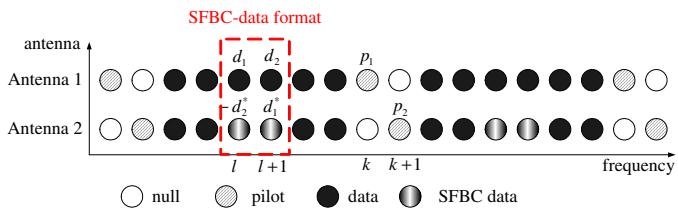


Figure 5: Pilot structure of PA-DDCE scheme with SFBC-data format

In the middle of any two adjacent pilot sub-carriers, the data symbol $d_1$ and $d_2$ on the sub-carriers $l$ and $l+1$ are encoded by the SFBC. Then the received signals on the sub-carrier $l$ and $l+1$ can be written as

$$\mathbf{r} = \mathbf{H}\mathbf{d}_s + \mathbf{w}, \tag{14}$$

where

$$\mathbf{d}_s = \begin{bmatrix} d_1(l) & d_2(l+1) \\ -d_2^*(l) & d_1^*(l+1) \end{bmatrix}. \tag{15}$$

Assuming the channel fluctuation on the adjacent sub-carriers is constant, i.e., $H_{ij}(l) = H_{ij}(l+1)$. The combined received signals are equivalent to that of the four-branch MRRC, thus the transmitted data symbols $\hat{d}_1(l)$ and $\hat{d}_2(l+1)$ on the sub-carrier $l$ and $l+1$ are computed by ZF equalizer, i.e.,

$$\begin{bmatrix} \hat{d}_1(l) \\ \hat{d}_2(l+1) \end{bmatrix} = \alpha \begin{bmatrix} \hat{H}_{11}^*(l) & \hat{H}_{21}(l) & \hat{H}_{12}^*(l) & \hat{H}_{22}(l) \\ \hat{H}_{21}^*(l) & -\hat{H}_{11}(l) & \hat{H}_{22}^*(l) & -\hat{H}_{12}(l) \end{bmatrix} \times \begin{bmatrix} r_1(l) \\ r_1^*(l+1) \\ r_2(l) \\ r_2^*(l+1) \end{bmatrix}, \tag{16}$$

Authorized licensed use limited to: Anthony Mcgrath. Downloaded on March 26,2024 at 18:30:04 UTC from IEEE Xplore. Restrictions apply.

where

$$\alpha = \frac{1}{\left|\hat{H}_{11}(l)\right|^2 + \left|\hat{H}_{21}(l)\right|^2 + \left|\hat{H}_{12}(l)\right|^2 + \left|\hat{H}_{22}(l)\right|^2}, \tag{17}$$

$\hat{H}_{ij}(l)$ are the estimated channel frequency response by the conventional channel estimation with the linear interpolation or other interpolation method, which are obtained in step 2.

The equalized symbol $\hat{d}_1(l)$ and $\hat{d}_2(l+1)$ are then demodulated and re-modulated to $\tilde{d}_1(l)$ and $\tilde{d}_1(l+1)$ respectively by the symbol-level hard-decision which are regarded as extra "pilot symbols" and can be used to compute the channel frequency response at the sub-carriers $l$ and $l+1$:

$$\tilde{\mathbf{H}}_s = \left(\tilde{\mathbf{d}}_s^H \tilde{\mathbf{d}}_s\right)^{-1} \tilde{\mathbf{d}}_s^H \mathbf{r}, \tag{18}$$

where

$$\tilde{\mathbf{H}}_s = \begin{bmatrix} \tilde{H}_{11}(l) & \tilde{H}_{21}(l) \\ \tilde{H}_{12}(l) & \tilde{H}_{22}(l) \end{bmatrix}, \tag{19}$$

and

$$\tilde{\mathbf{d}}_s = \begin{bmatrix} \tilde{d}_1(l) & \tilde{d}_2(l+1) \\ -\tilde{d}_2^*(l) & \tilde{d}_1^*(l+1) \end{bmatrix}. \tag{20}$$

The final enhanced estimated channel frequency response can be written as

$$\hat{H}_{ij}(n) = interp\left(\hat{H}_{ij}(k), \tilde{H}_{ij}(l)\right). \tag{21}$$

## V NUMERICAL RESULTS

Table 1: simulation parameters

| | |
|---|---|
| Bandwidth | 5 MHz |
| Number of Sub-carrier | 1024 |
| Channel Model | ITU-Pedestrian B |
| Channel Coding | Turbo Coding, Rate 1/2 |
| Data Modulation | QPSK |
| Antenna Configuration | 2Tx, 2Rx |
| Pilot Modulation | QPSK |
| Data Detection | MMSE |
| Repeat Factor (RPF) | 24, 36 |
| Channel Estimation | Ideal,<br>Conventional (conv),<br>PA-DDCE with null data (dd-null),<br>PA-DDCE with SFBC data (dd-SFBC). |
| Pilot Overhead | 8.4 % |
| Interpolation | Linear |

In this section, the performance of the proposed PA-DDCE scheme is measured by the bit error rate (BER) and the system throughput, which are compared with those of the conventional channel estimation. The simulation specifications are presented in Table I. The repeat factor (RPF) means the sub-carrier spacing between the adjacent pilots. For fair comparison, the pilot overhead is assumed to be the same, thus for conventional channel estimation the RPF is 24 while for PA-DDCE scheme it is 36. In other words, we constrain the useful data sub-carriers the same. The channels are quasi-static Rayleigh faded over a coding frame and independent among different frames. We also assume the channels between the transmit antennas and the receive antennas are independent with each other. Here the ITU-PB channel [7] refers to a 6 taps Rayleigh fading channel. To detect the data symbols the well-known linear minimum mean square error (MMSE) [8] solution is applied for the MIMO-OFDM system in our simulation.

### A *Estimated channel accuracy comparison*

Fig.6 plots the estimated channel gain $\left|\hat{H}\right|^2$ obtained from both the conventional channel estimation and the proposed PA-DDCE scheme with SFBC-data format in comparison with the ideal channel gain $|H|^2$. As it is shown when SNR is 20dB the estimated channel gain in the middle of any two adjacent pilot sub-carriers computed from the extra "pilot symbols" by PA-DDCE scheme can improve the total estimated channel quality dramatically.

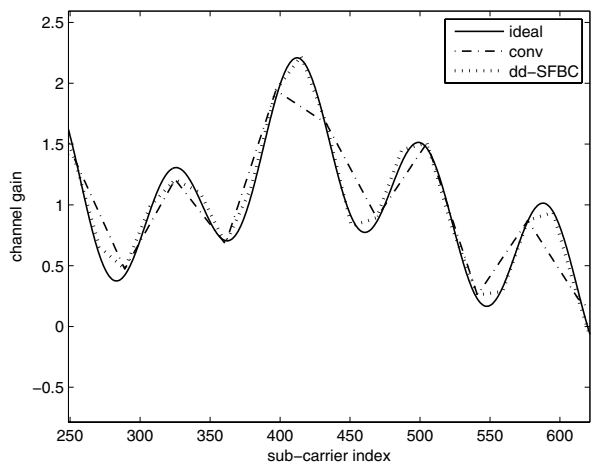


Figure 6: Estimated channel comparison under SNR = 20dB

The Mean Square Error (MSE) of the estimated channel by using different channel estimations is depicted in Fig.7. Applying the PA-DDCE scheme, it can be observed that the MSE performance is significantly improved compared with the conventional channel estimation. Under high SNR, the improvements become larger due to the higher reliability of the symbol-level hard decision. Further more, the PA-DDCE scheme with the SFBC-data format works even better than that with the null-data format. The reason is that for the SFBC-data format, the received extra "pilot symbols" are combined as the four-branch MRRC while for the null-data format they are combined as the two-branch MRRC.

### B *BER and throughput performance*

The channel estimation performance is mainly dominated by two factors: the estimation error on the pilot sub-carriers and the interpolation error on the none-pilot sub-carriers. When the

Authorized licensed use limited to: Anthony Mcgrath. Downloaded on March 26,2024 at 18:30:04 UTC from IEEE Xplore. Restrictions apply.
Petitioner Samsung Ex-1008, 0004

pilots spacing is larger than the channel coherent bandwidth, the interpolation error becomes the major problem affecting the channel estimation. Thus with fixed small pilot overhead, the conventional channel estimation can not achieve the required system throughput performance. When the channel is ideal, to guarantee the same system spectrum, we assume the pilot overhead is still 8.4%, i.e. RPF of 24. The results also show the proposed PA-DDCE scheme outperforms the conventional channel estimation significantly. For example, when the SNR is 12.5dB, the system throughput gains are 0.5 Mbps and 1 Mbps for the null-data format and the SFBC-data format respectively.

## VI CONCLUSIONS

For the high-data-rate wireless applications especially utilizing the MIMO-OFDM technology as the air interface, the number of the channel parameters used for the data detection increased dramatically. To achieve the channel estimation quality while still maintaining the high spectrum efficiency, a novel PA-DDCE scheme for the MIMO-OFDM system is presented in this paper. We also proposed two alternative pilot structures with the null-data format and the SFBC-data format respectively to mitigate the inter-antenna interference for the PA-DDCE scheme. From our simulation results the PA-DDCE scheme can outperform the conventional channel estimation significantly respect to both the BER and the throughput performances. The only disadvantage is a slight but acceptable complexity on the receiver side.

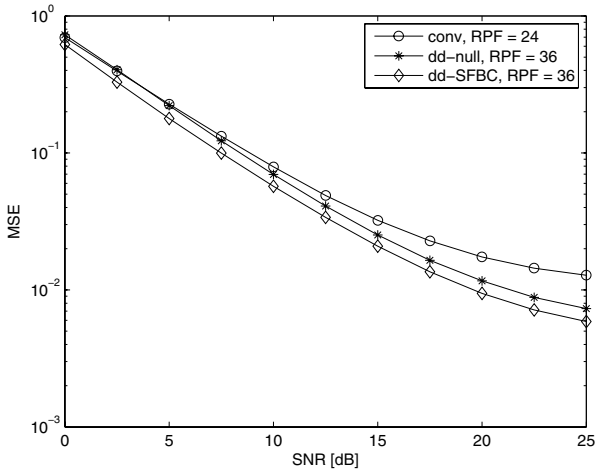


Figure 7: MSE performance of QPSK 1/2 for ITU-PB channel

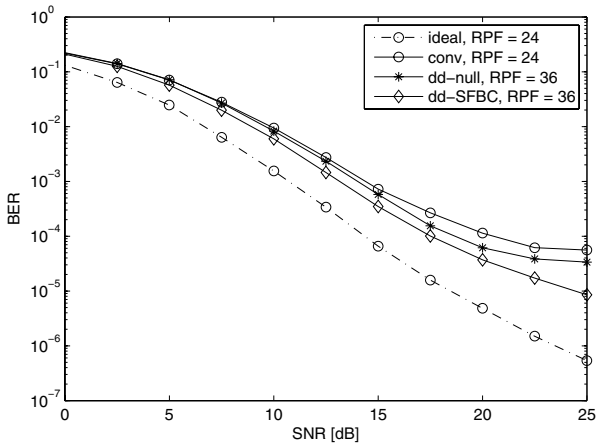


Figure 8: BER performance of QPSK 1/2 for ITU-PB channel

## REFERENCES


[1] C Shin, W Robert Jr, EJ Powers,"Blind Channel Estimation for MIMO-OFDM Systems," in *IEEE Transactions on Vehicular Technology*,vol.56, pp.670-685, 2007.

[2] S. Zhou, B. Muquet and G. B. Giannakis, "Subspace-based (Semi-) Blind Estimation of Frequency-Selective Channels for Space-Time Block Precoded Transmissions," *IEEE Trans. On Signal Processing*, pp. 975-979, Feb. 2001.

[3] Ran J, Grunheid R, "Decision-directed channel estimation method for OFDM systems with high velocities," *Vehicular Technology Conference*, pp. 2358-2361, 2003.

[4] T. Kella, "Decision-directed channel estimation for supporting higher. terminal velocities in OFDM based WLANs," *Proc. IEEE GLOBECOM*, vol.3, pp.1306C1310, Dec. 2003.

[5] S. M. Alamouti, "A simple transmitter diversity scheme for wireless communications," *IEEE J. Select. Areas Commun.*, vol. 16, pp. 1451-1458, Oct. 1998.

[6] J. J. van de Beek, O. Edfors, M. Sandell, S. K. Wilson, and P. O. Borjesson, "On channel estimation in OFDM systems," *Proc. 45th IEEE. Veh. Technol. Conf.*, Chicago, pp. 815C819, Jul. 1999.

[7] The International Telecommunications Union, Guidelines for Evaluation of Radio Transmission Technologies for IMT-2000, *Rec. ITU-R M.1225*, 1997.

[8] R. Lupas and S. Verdu, "Linear multiuser detectors for synchronous code-division multiple-access channels," *IEEE Trans.Inf.Th.*, Vol.35, no.1, pp.123-136, Jan. 1989.


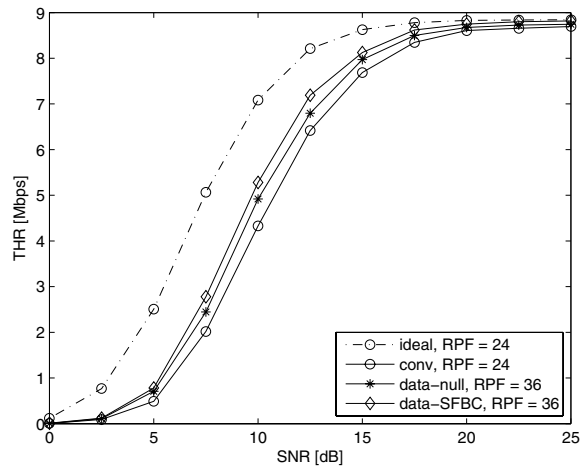


Figure 9: Throughput performance of QPSK 1/2 for ITU-PB channel

Authorized licensed use limited to: Anthony Mcgrath. Downloaded on March 26,2024 at 18:30:04 UTC from IEEE Xplore. Restrictions apply.