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MINIMUM BIT ERROR RATE ASSISTED QPSK FOR PRE-FFT BEAMFORMING IN LTE OFDM COMMUNICATION SYSTEMS

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ABSTRACT

In this paper, a diamond shape pilot arrangement for OFDM channel estimation is investigated. Such arrangement will decrease the number of pilots transmitted over the communication channel, which will in turn increase data throughput while maintaining acceptable accuracy of the channel estimation. The adaptive antenna array (AAA) is combined with orthogonal frequency division multiplexing (OFDM) to combat the intersymbol interference (ISI) and the directional interferences.

In this paper, The optimum beamformer weight set is obtained based on minimum bit error rate (MBER) criteria in diamond-type pilot-assisted in 3GPP long term evolution (LTE) OFDM systems under multipath fading channel. The simulation results show that the quadrature phase shift keying signaling based on MBER technique utilizes the antenna array elements more intelligently than the standard minimum mean square error (MMSE) technique.

KEYWORDS

MBER beamforming, OFDM systems, Pre-FFT, Diamond shape pilot, 3GPP, LTE.

1. INTRODUCTION

Orthogonal Frequency Division Multiplexing (OFDM) is considered an efficient technique for high speed digital transmission over severe multipath fading channels, where the delay spread is larger than the symbol duration. When the inserting guard time is longer than the delay spread of the channel, this makes the system robust against inter-symbol interference (ISI). In addition to that, channel estimation and compensation can be achieved by inserting known pilot symbols between data symbols [1]-[3].

Over the past few years, Adaptive Antenna Array (AAA) has gained much attention due to its ability to increase the performance of wireless communication systems, in terms of spectrum efficiency, network scalability and operation reliability. Antenna arrays can mitigate the effect of ISI and relax the design of channel equalizer [2].

Adaptive beamforming can separate transmitted signals on the same carrier frequency, provided that they are separated in the spatial domain. The beamforming processing combines the signals received by the different elements of an antenna array to form a single output. The adapted weight set of each element of the antenna array is obtained by the processor, achieving certain criteria to suppress the co-channel interference; thus improving coverage quality.

For a communication system, it is the achievable bit-error rate (BER), not the MSE performance, that really matters. Ideally, the system design should be based directly on minimizing the BER, rather than the MSE. It is demonstrated in Ref. [3] that the MBER solution utilizes the array weights more intelligently than the MMSE approach.

One of the two main techniques which are used in OFDM systems is called Pre-FFT, where an optimum beamformer weight set is obtained in time domain before Fast Fourier Transform (Pre-FFT). The main motivation behind Pre-FFT scheme is reducing the cost due to FFT processing [1]-[7]. The weight obtained for each pilot subcarrier can be identically applied on all data subcarriers in the same OFDM symbol; thereby reducing the number of frequency domain narrow-band beamformers. Post FFT is not always better in performance than pre-FFT [2]-[3]. In [4], a pre-FFT least mean square (LMS) beamforming for OFDM systems was analyzed in additive Gaussian noise channel. An adaptive MBER beamforming was analyzed in [4] for single carrier modulation and in [2] for OFDM systems in additive Gaussian noise channel. A class of MBER algorithms were studied in [8] and combined with space time coding in [9]. Eigenvector combining was considered in [8]. MIMO MBER beamforming for OFDM was studied in [5]. A block by block post-FFT multistage beamforming was considered in [4].

In [1]-[2], the MMSE and MBER beamformers for Pre-FFT OFDM are presented, respectively, without investigating several factors affecting performance. The channel is assumed to be non-dispersive with additive Gaussian noise, which is not a practical channel.
Since new wireless standards, such as IEEE 802.11 and 802.16, use the pilot subcarriers in their structures, our focus in this paper will be given to suppress co-channel interference and mitigate the multipath interference in pilot-assisted OFDM systems.

In [3], the MMSE beamforming algorithm for Pre-FFT OFDM system is applied on a channel assumed to be frequency selective fading. A recent work [5]-[6] has suggested an adaptive MBER beamforming assisted receiver for binary phase shift keying OFDM communication systems. This paper first presents a novel beamforming technique based directly on minimizing the system's BER for broadband OFDM wireless systems with quadrature phase shift keying (QPSK) modulation. The main contribution in this paper is to show that the diamond type pilot aided channel estimation has better performance when the channel is time-variant and the reduced number of pilot increases efficiency. The paper is an extension to [2] with an improved channel estimator and the performance results are more applicable to 3GPP-LTE.

This paper is organized as follows: Section 2 describes the LTE pilot structure. Section 3 describes the Pre-FFT adaptive beamforming based on MMSE criteria. In Section 4, Pre-FFT adaptive beamforming based on MBER criteria is introduced. Sections 5 and 6 clarify the computational complexity and the convergence rate for the simulated system, respectively. System specification is shown in section 7. In section 8, simulation results are provided. Finally, conclusions and possible directions for future work are presented in section 9.

2. LTE PILOT STRUCTURES

As indicated earlier, wireless standards, such as IEEE 802.11 and 802.16, use the pilot subcarriers in their structures. This pilot signal is used to measure the channel quality and perform channel estimation at the end-user side. There are several types of pilot structures: Block-type pilot, Comb-Type pilot and Diamond-type pilot. In Table 1, we present a quick comparison between the system specifications for Block-type pilot [14] and Diamond-type pilot simulated in this paper:

In this paper, we research further the time domain MBER and LMS channel estimation based on diamond-type pilot structure. The aim is to achieve better performance and fewer computations and at the same time increase spectral efficiency and throughput by using less number of pilot signals.

Table 1. Simulation system specifications for Block and Diamond type Pilot.

	Block-type Pilot	Diamond-type Pilot
Cluster Size	4	6 (As shown in Figure 1)
Subcarrier	64 , 128 , 256 , 512	60 , 120 , 300 , 600
Number Pilot = (Subcarrier ÷ Cluster Size)	16 , 32 , 64 , 128	10 , 20 , 50 , 100

Figure 1 shows an example of the pilot pattern in every RB of a subframe at the first antenna port, but the location may be shifted in frequency domain for different subframes [10].

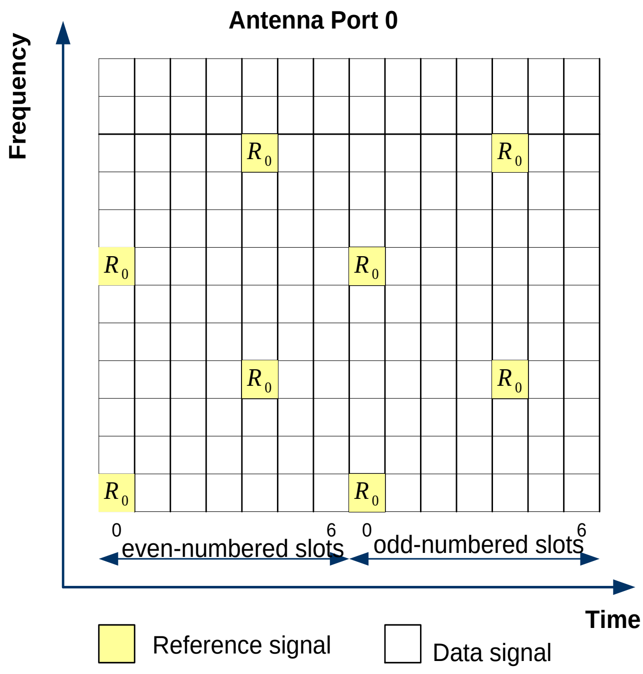

Figure 1. Pilot locations for the first transmit antenna in the 3GPP-LTE system [10].

Consider M-users, where each user transmits a QPSK signal and the OFDM system uses K subcarriers for parallel transmission [2]. The sample modulated by the kth subcarrier of the mth user is given by:

$$x_m(k) = b_m(k) \qquad 1 \le m \le M \qquad 1 \le k \le K \tag{1}$$

where $b_m(k) \in \{\pm 1 \pm j\}$ are QPSK symbols. Source 1 is assumed to be the desired user and the rest of the sources are the interfering users. This data can be interpreted to be a frequency-domain data and subsequently converted into a time-domain signal by an IFFT operation. This process can be written as:

$$\bar{y}_m = \frac{1}{K} F^H \bar{x}_m \qquad 1 \le m \le M \tag{2}$$

where,

$$\bar{y}_m = \left[y_m(1), y_m(2), \ldots, y_m(K)\right]^T \tag{3}$$

$$F = \begin{bmatrix} 1 & 1 & \cdots & 1 \\ 1 & e^{\frac{-j2\pi(1)(1)}{K}} & \cdots & e^{-\frac{j2\pi(1)(K-1)}{K}} \\ \vdots & \vdots & \ddots & \vdots \\ 1 & e^{\frac{-j2\pi(K-1)(1)}{K}} & \cdots & e^{-\frac{j2\pi(K-1)(K-1)}{K}} \end{bmatrix} \tag{4}$$

$$\bar{x}_m = \left[x_m(1), x_m(2), \ldots, x_m(K)\right]^T \tag{5}$$

F is representing the FFT operation matrix and H denotes the Hermitian transpose of a matrix. To add the CP, $\bar{y}_m$ is cyclically extended to generate $\tilde{y}_m$ by inserting the last v element of y_m at its beginning; i.e.,

$$\tilde{y}_m = \begin{bmatrix} J_v \\ I_K \end{bmatrix} \bar{y}_m \tag{6}$$

where J_v contains the last v rows of a size K identity matrix I_K .

Finally, the OFDM time signals are transformed to the analog format by a D/A converter prior to transmission. A multipath channel with a maximum of L paths exists between the m^{th} source (desired or interference) and the array in the form of:

$$h_m(k) = \sum_{l=0}^{L-1} \alpha_{m,l} \delta(k-l) \qquad m = 1, \cdots, M \tag{7}$$

where $\alpha_{m,l}$ denotes a complex random number representing the l^{th} channel coefficient for the m^{th} source and δ(.) is the delta function.

Figure 2 illustrates the architecture of Pre-FFT beamforming at the receiver of an OFDM system, where *CP* is assumed to be longer than the channel length ($v > L$); thus, received signal on the p^{th} antenna of a Uniform Linear Array (ULA) for one OFDM symbol can be written as:

$$r_p(k) = \sum_{m=1}^{M} \sum_{l=0}^{L-1} \alpha_{m,l} \tilde{y}_m(k+v-l) e^{-j\frac{2\pi}{\lambda}(p-1)d\cos(\theta_{m,l})} + \eta_p(k) \tag{8}$$
$$1 \le p \le P \quad , \qquad 1 \le k \le K$$

where, $\eta_p(k)$ represents the channel noise entering the p^{th} antenna. $\theta_{m,l}$ denotes the direction of arrival (DOA) of the l^{th} path and m^{th} source. Without loss of generality, we have assumed that the channels of all sources have the same length L. At the receiver, the converted digital signal with a spatial phase for each array element is multiplied by the weight (w_p) of adaptive beamformer and then transformed back into frequency-domain (data and pilot) symbols by the FFT. This process can be written as:

$$Z(k) = W^H \cdot \bar{R} \tag{9}$$

$$W = [w_1 \, w_2 \quad \cdots w_P]^T \tag{10}$$

$$\bar{R}(k) = [\bar{r}_1(k) \bar{r}_2(k) \quad \cdots \bar{r}_P(k)]^T \tag{11}$$

$$\bar{Z} = [z(1) \, z(2) \quad \cdots z(K)] \tag{12}$$

$$\hat{Z} = \bar{Z} \cdot F \tag{13}$$

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where $\hat{Z}$ is the frequency-domain data, which is given by:

$$\hat{Z} = [\hat{z}(1), \hat{z}(2), \ldots, \hat{z}(K)]^T \tag{14}$$

and $\hat{z}(k)$ denotes the corresponding received sample at the k^{th} subcarrier .

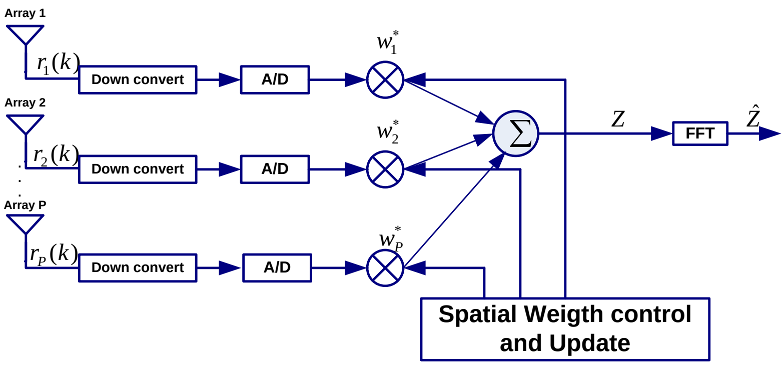

Figure 2. Block diagram of the Pre-FFT OFDM adaptive receiver [1].

The estimate of the transmitted bit $b_i(k)$ is given by:

$$\hat{b}_i(k) = \begin{cases} b^{[1]} = +1 + j, & (\hat{z}_{\mathrm{Re}}(k)) \geq \text{and } (\hat{z}_{\mathrm{Im}}(k)) \geq 0, \\ b^{[2]} = -1 + j, & (\hat{z}_{\mathrm{Re}}(k)) < 0 \text{ and } (\hat{z}_{\mathrm{Im}}(k)) \geq 0, \\ b^{[3]} = -1 - j, & (\hat{z}_{\mathrm{Re}}(k)) < 0 \text{ and } (\hat{z}_{\mathrm{Im}}(k)) < 0, \\ b^{[4]} = +1 - j, & (\hat{z}_{\mathrm{Re}}(k)) \geq 0 \text{ and } (\hat{z}_{\mathrm{Im}}(k)) < 0, \end{cases} \tag{15}$$

where $\hat{b}_i(k)$ is the kth symbol of user i, which takes values from a QPSK symbol set shown in equation (15), $(\hat{z}_{\mathrm{Re}}(k))$ denotes the real part of $\hat{z}(k)$ and $(\hat{z}_{\mathrm{Im}}(k))$ denotes the imaginary part of $\hat{z}(k)$.

3. Adaptive Beamforming for Pre-FFT OFDM System

Implementing MMSE Beamforming using LMS adaptive algorithm is done by comparing the received pilot symbols with their known values, so that an error signal is generated at the receiver [1]-[2]. Since this error signal is in frequency domain while Pre-FFT weights are updated in time domain, the frequency-domain error signal must be converted into time domain.

If there are a total of Q pilot symbols in every OFDM symbol, then we define two $K \times 1$ vectors d_q and Z_q, such that the k^{th} element of d_q is zero if k is a data subcarrier and is the known pilot value if k is a pilot subcarrier.

Similarly, the k^{th} element of Z_q is zero if the k is a data subcarrier and is the received pilot value if k is a pilot subcarrier. Therefore, the error signal in frequency domain is given by:

$$E_q = d_q - Z_q . \tag{16}$$

This error signal must be converted into time domain for the Pre-FFT weight adjustment algorithm. Therefore,

$$\bar{e} = \frac{1}{K} F^H E_q \tag{17}$$

where e is the vector of error samples in time domain.

$$\bar{e} = [e(1)\, e(2) \quad \cdots e(K)]^T \; . \tag{18}$$

Consequently, the Pre-FFT weights are updated using the following Least Mean Squares (LMS) algorithm [1]-[2].

$$\begin{gathered} W(k) = W(k-1) + 2\mu \cdot r(k) \cdot e^*(k) \\ 1 \le k \le K \end{gathered} \tag{19}$$

where μ is the step size parameter, and $*$ represents the complex conjugate. The last update at the end of each OFDM block (W (K)) is used as the initial value of the next block.

4. MBER-BASED BEAMFORMING ALGORITHMS

In this section, Pre-FFT adaptive beamforming based on MBER criteria is introduced to obtain the optimum weight set. The theoretical MBER solution for the Pre-FFT OFDM beamformer is obtained in [1]-[2], [17] where the channel is assumed to be non-dispersive with additive white Gaussian noise. The error probability (BER cost function) of the frequency domain signal of the beamformer is given by:

$$P_E(W) = \text{Prob}\{\text{sgn}(b_1(k)\text{Real}(\hat{z}(k)) < 0\} \tag{20}$$

where $\text{sgn}(\cdot)$ is the sign function.

From equation (20), define the signed decision variable

$$\begin{aligned} \hat{z}_s(k) &= \text{sgn}(b_1(k))\text{Re}(\hat{z}(k)) \\ &= \text{sgn}(b_1(k))\text{Re}(\hat{z}'(k)) + \eta'(k) \end{aligned} \tag{21}$$

where,

$$\hat{z}'(k) = W^H[\bar{r}(k) - \eta(k)]F(k) \tag{22}$$

and

$$\eta'(k) = \text{sgn}(b_1(k))\text{Re}(W^H)\eta(k)F(k)) \tag{23}$$

$\hat{z}_s(k)$ is a very good error indicator for the binary decision; i.e., if it is positive, then the decision is correct, else if it is negative, then an error occurred. $F(k)$ is the $\eta'(k)$ k^{th} column of F. Notice that F is a unitary matrix, so is still Gaussian with zero mean and variance $\sigma_n^2 \cdot W^H W$.

Obviously, the two marginal conditional p.d.f.s for $z_{\text{Re}}(k)$ and $z_{\text{Im}}(k)$ are Gaussian mixtures. Define

$P_{E,\text{Im}}(W\ ) = \text{prob}(\hat{b}_{\text{Im},1}(k)) \ne b_{\text{Im},1}(k))$, $\hat{b}_1(k) = \hat{b}_{\text{Re},1}(k) + j\hat{b}_{\text{Im},1}(k)$ and

$b_1(k) = b_{\text{Re},1}(k) + jb_{\text{Im},1}(k)$. Obviously, the two marginal conditional p.d.f.s are for $z_{\text{Re}}(k)$ and $z_{\text{Im}}(k)$.

The BER of the beanformer for equation (20) is:

$$P_E(W) = \frac{1}{2}(P_{E,\text{Re}}(W\ ) + P_{E,\text{Im}}(W)). \tag{24}$$

The MBER beamforming solution is defined as:

$$W = \arg\min_W P_E(W) \; . \tag{25}$$

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The gradient of $P_E(w)$ with respect to w is:

$$\nabla P_E\ (W) = \frac{1}{2}(\nabla P_{E,\text{Re}}(W\ ) + \nabla P_{E,\text{Im}}(W))\,. \tag{26}$$

Given the gradient, the optimization problem (26) can be solved for interactively using the simplified conjugated gradient algorithm, which is detailed in [4], [14].

The conditional probability density function (pdf), given the channel coefficients $\alpha_{m,l}$ of the error indicator, $\hat{z}_s(k)$, is a mixed of Gaussian distributions [3]; i.e.,

$$p_z(\hat{z}_s) = \frac{1}{K\sqrt{2\pi}\sigma_n\sqrt{W^H W}}\,. \\ \sum_{k=1}^{K}\exp(-\frac{(\hat{z}_s - \text{sgn}(b_1(k))\text{Re}(\hat{z}'_s(k)))^2}{2\sigma_n^2 W^H \cdot W}) \tag{27}$$

And it is the best indicator of a beamformer's BER performance deriving a closed form for the average error probability is not easy. Therefore, we use the gradient conditional error probability to update the weight vector. The conditional error probability, given the channel coefficients $\alpha_{m,l}$ of the beamformer $P_E(W\ )$, is given in [3].

$$p_E(w) = \frac{1}{K\sqrt{2\pi}\sigma_n\sqrt{W^H W}}\,. \\ \sum_{k=1}^{K}\int_{q_k(w)}^{\infty}\ \exp(-\frac{u^2}{2})du \tag{28}$$

$$\hat{P}_E(W) = \frac{1}{2K}\sum_{k=1}^{K}(Q(q_{\text{Re}}^{(k)}(W\ )) + Q(q_{\text{Im}}^{(k)}(W\ )) \tag{29}$$

where $Q(\cdot)$ is the Gaussain error function and is given by:

$$Q(u) = \frac{1}{\sqrt{2\pi}}\int_{u}^{\infty}\exp(-\frac{y^2}{2})dy \tag{30}$$

$$q_{\text{Re}}^{(k)}(W\ ) = \frac{\text{sgn}(b_1(k))(\hat{z}'_{\text{Re}}(k))}{\sigma_\eta\sqrt{W^H W}} \tag{31}$$

$$q_{\text{Im}}^{(k)}(W\ ) = \frac{\text{sgn}(b_1(k))(\hat{z}'_{\text{Im}}(k))}{\sigma_\eta\sqrt{W^H W}}\ . \tag{32}$$

Based on the definition, the gradient of $P_E(W\ )$ with respect to W is:

$$\nabla P_E(W\ ) = \frac{1}{K_p\sqrt{2\pi}\sigma_n\sqrt{W^H W}} \cdot \sum_{n=1}^{N_b}\exp(-\frac{\hat{z}'^2(k)}{2\sigma_n^2 W^H \cdot W}) \\ \cdot\text{sgn}(b_1(k))(\frac{\text{Re}(\hat{z}'(k))W}{W^H W} - A\bar{r}(k)) \tag{33}$$

In an OFDM system, it is assumed that there are pilot signals in every symbol to do channel estimation [1]-[2]. The pilot signals are also used to adaptive update of the weight vector of the beamformer. The transmitted pilot signal vector of desired user $\bar{x}_{1p}$ and the received pilot signal vector $\bar{\hat{z}}_p$ in frequency domain can be written as follows:

$$\bar{x}_{1p} = [x_1(1),0..,x_1(\Delta p+1),0,..,x_1((K_p-1)\Delta p+1)),0,..] \tag{34}$$

$$\bar{\hat{z}}_p = [\hat{z}(1),0..,\hat{z}(\Delta p+1),0,..,\hat{z}((K_p-1)\Delta p+1),0,..] \\ = W^H \bar{R} F_p$$

where

$$Fp=\begin{bmatrix} 1 & 0 & \cdots & 1 & \cdots & 1 & 0\cdots \\ 1 & 0 & \cdots & e^{-j2\pi(1)(\Delta p)/K} & \cdots & e^{-j2\pi(1)(K_p-1)\Delta p/K} & 0\cdots \\ \vdots & 0 & & \vdots & \ddots & \vdots & 0\cdots \\ 1 & 0 & \cdots & e^{-j2\pi(K-1)(\Delta p)/K} & \cdots & e^{-j2\pi(K-1)(K_p-1)\Delta p/K} & 0\cdots \end{bmatrix} \tag{35}$$

$\bar{R} = [\bar{r}(1)\,\bar{r}(2) \quad \cdots \bar{r}(K)]$, Δp and K_p represents the frequency spacing between consecutive pilot symbols and the number of pilot symbols inserted in OFDM symbol, respectively. We assume that the first pilot symbol is positioned at the first sub-channel.

The method of approximating a conditional pdf, known as a kernel density or Parzen window-based estimate [7]-[8], is used to estimate the conditional error probability given that the channel coefficients $\alpha_{m,l}$ is used on OFDM systems. Given a symbol of K_p training samples $\{\bar{r}(k), b_1(k)\}$, a kernel density estimate of the conditional pdf given the channel coefficients $\alpha_{m,l}$ at pilot locations, is given by:

$$\hat{p}(\hat{z}) = \frac{1}{2\pi K_p \rho_\eta^2 W^H W} \sum_{K_p=1}^{K_p} \exp\left(\frac{|\hat{z} - \hat{z}(k)|^2}{2\rho_\eta^2 W^H W} \right) \tag{36}$$

where the kernel width ρ_η is related to the noise standard deviation σ_η .From this estimated p.d.f., the estimated BER is given by:

$$\hat{P}_E(W) = \frac{1}{2K_p} \sum_{k=1}^{K_p} (Q(\hat{g}_{\mathrm{Re}}^{(k)}(W)) + Q(\hat{g}_{\mathrm{Im}}^{(k)}(W)) \tag{37}$$

where

$$\hat{g}_{\mathrm{Re}}^{(k)}(W) = \frac{\mathrm{sgn}(b_{\mathrm{Re},1}(k \times \Delta p+1))\mathrm{Re}(W^H \bar{R} F_p(k \times \Delta p+1))}{\rho_\eta \sqrt{W^H W}} \tag{38}$$

and

$$\hat{g}_{\mathrm{Im}}^{(k)}(W) = \frac{\mathrm{sgn}(b_{\mathrm{Im},1}(k \times \Delta p+1))\mathrm{Im}(W^H \bar{R} F_p(k \times \Delta p+1))}{\rho_\eta \sqrt{W^H W}} \tag{39}$$

$F_p(k \times \Delta p+1)$ is the $(k \times \Delta p+1)^{th}$ column of F_p. From this estimated conditional pdf, given the channel coefficients $\alpha_{m,l}$, the gradient of the estimated BER is given by [3]:

$$\nabla \hat{P}_{E,\mathrm{Re}}(W) = \frac{1}{2K_p \sqrt{2\pi}\rho_\eta \sqrt{W^H W}} \sum_{k=1}^{K_p} \exp\left(\frac{-z_{\mathrm{Re}}^2(k)}{2\rho_\eta^2 W^H W}\right) \\ \times \mathrm{sgn}(b_{\mathrm{Re},1}(k))\left(\frac{z_{\mathrm{Re}}^2 W}{W^H W} - \bar{R}F(k)\right) \tag{40}$$

ρ_n is related to the standard deviation σ_n of the channel noise. From this estimated pdf, the gradient of the estimated BER is given by [1]:

$$\nabla \hat{P}_E(W) = -\frac{1}{\sqrt{2\pi}\rho_n\sqrt{W^H W}} \sum_{k=0}^{K_p-1} \exp\left(-\frac{(\mathrm{Re}(\hat{z}(k\times\Delta p+1)))^2}{2\rho_n^2 W^H W}\right) \times \mathrm{sgn}(b_1(k\times\Delta p+1))\bar{R}F_p(k\times\Delta p+1) \tag{41}$$

$$\nabla \hat{P}_{E,\mathrm{Im}}(W) = \frac{1}{2K_p\sqrt{2\pi}\rho_\eta\sqrt{W^H W}} \sum_{k=1}^{K_p} \exp\left(\frac{-z_{\mathrm{Im}}^2(k)}{2\rho_\eta^2 W^H W}\right) \times \mathrm{sgn}(b_{\mathrm{Im},1}(k))\left(\frac{z_{\mathrm{Im}}^2 W}{W^H W} - \bar{R}F(k)\right) \tag{42}$$

For each OFDM symbol, we can find the optimum weight vector W by the steepest-descent gradient algorithm [3]:

$$\nabla \tilde{P}_E(W,k) = \frac{1}{4\sqrt{2\pi}\rho_\eta}\left[-\mathrm{sgn}(b_{\mathrm{Re},1}(k\times\Delta p))\exp\left(\frac{-(\hat{z}_{\mathrm{Re}}(k\times\Delta p+1))^2}{2\rho_\eta^2}\right) + j\,\mathrm{sgn}(b_{\mathrm{Im},1}(k\times\Delta p))\exp\left(\frac{-(\hat{z}_{\mathrm{Im}}(k\times\Delta p+1))^2}{2\rho_\eta^2}\right)\right]\times \bar{R}F_p(k\times\Delta p+1)) \tag{43}$$

That is to say, W weight vector can be updated K_P times in one OFDM symbol. Thus, complexity is reduced and consequently, the update equation is given by:

$$W(k+1) = W + \mu(-\nabla\tilde{P}_E(W(k),k) \tag{44}$$

$$c_1 = W^H(k+1)P_1 \tag{45}$$

$$P_1 = A_1 s_1 \tag{46}$$

where s_1 is the steering vector and A_1 is power of desired user.

$$W\ (k+1) = \frac{c_1}{|c_1|}W(k+1)\ . \tag{47}$$

We assume a perfect P1 at receiver and known arrival direction of the desired user. It is shown in ref. [15] that the steering vector s1 should be known at the receiver, because the beamformer's output consists of: (the desired signal + residual interference) , the steering vector s1 determines the desired signal. Also it is indicated that QPSK case is different from BPSK case in which the receiver does not require the steering vector of the desired user to make a decision.

The proposed MBER algorithm is summarized in Table 2, this algorithm is composed of two main loops. The outer loop is for each symbol of data and the inner loop is repeated over the same symbol of data until certain number of iterations is reached. In the main loop, we formulate a symbol of data (300 bits) from the output of the antenna array. In the inner loop the gradient vector is determine from (43) at pilot locations (Np =50). Then, we compute the weight update vector from (44). After the end of the inner loop, we determine the detected signal by multiplying the computed optimized weight vector with the received signal in order to use it in calculating the BER the last update at the end of each OFDM block W (k) which used as an initial value in the next symbol. Then, we get back to the main loop and form another symbol of data and so on. These processes iterate until we finish all the incoming data.

5. COMPUTATIONAL COMPLEXITY

In this section, we compare the two algorithms in terms of computational complexity [4]. Table 3 illustrates the computational complexity of pre-weight update to complete a single iteration; i.e., detecting one bit. The proposed MBER maintains the linearity in complexity.

Table 2. MBER algorithm summary.

Initialization
$i=1, \mu=.01,$ size $K=300, K_p=50$. • Calculate variance of noise σ_η , ρ_η . • Initial weight vector $W=.01\times ones(N_t,1)$.
Outer loop (1: floor (all bits/symbol))
• Form a symbol of data from the received signals.
Inner loop (while $k<K_p$)
• Calculate the gradient matrix over the symbol in eqn. (43). • Update the weight matrix as: $W(k+1)=W(k)+\mu(-\nabla\hat{P}_E(W(k),w))$. • $P_1=A_1 s_1$ $c_1=W^H(k+1)P_1$, where c_1 is real and positive. $W(k+1)=\frac{c_1}{\lvert c_1\rvert}W(k+1)$, the rotating operation. • Normalize the solution $W(k+1)=W(k+1)/\lVert W(k+1)\rVert$.
• ***end of inner loop*** • Determine the detected signals in order to be used for calculating the BER. • Increment the symbol number.
• ***end of outer loop***

Table 3. Comparison of computational complexity pre-weight update.

	Multiplications	additions	exp(•) evaluation
MBER	$4\times P+4$	$4\times P-1$	1
MMSE	$8\times P+2$	$8\times P-1$	–

6. CONVERGENCE RATE

In this section, we run the algorithm of the MBER for 400 samples limited to 1 and 11 iterations. The results are shown in Figure 11, where we can see that the proposed algorithm converges very fast to the optimal solution (after one iteration only).

Figure 9 and Figure 10 illustrate the convergence performance of LMS based on MMSE Pre-FFT beamformer. The optimum weights and steady state MSE performance, respectively, can be obtained under the same conditions after about 200 OFDM symbols.

The figures resulting from the simulation show that shorter training symbols and less computational complexity OFDM symbols are required in the MBER Pre-FFT beamformer algorithm and the MBER algorithm converges faster than MMSE.

7. SYSTEM SPECIFICATION

In this paper, we follow the specification of 3GPP-LTE Release 8 [10]-[11] to perform the analysis, simulation and implementation. The 3GPP-LTE supports a maximum 512-point FFT size at 10-MHz bandwidth and has six antenna elements and half-wavelength spacing. The general shape of the diamond shape pilot signal was shown in Figure 1.

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8. Simulation Results

In this section, the simulation is performed to illustrate and compare the performances of Pre-FFT beamformer using different MBER-based algorithms. 300 subcarriers (*50 + 250*) are used. The OFDM system is perfectly synchronized with a CP length larger than the channel length ($v=16$). QPSK modulation is used in the system with six antenna elements and half-wavelength spacing. The example used in our computer simulation study considers one desired user with DOA at 90° and two interferers with SIR= -3dB and 0dB, respectively and DOA 50° and 140°. We further assumed channels with different lengths and with real coefficients *0.8935,0.0957,0.0107* and *0* for all sources and with an angle spread of $\pm 15^o$ (for all sources) [12].

Figure 3 and Figure 4 compare the BER performance of the MBER beamformer with that of the MMSE beamformer for SIR= -3dB and 0dB, respectively with AWGN. Figure 5 and Figure 6 compare the BER performance against SNR for the LMS and MBER beamformers for the case that the number of elements is 6 under selective fading channel. It is observed that the BER performance of the MBER beamformer is better than that of the LMS. Figure 7 and Figure 8 for different channel real coefficients illustrates the beam pattern of the LMS, MBER and beamformers for Pre-FFT OFDM adaptive antenna array when the number of antenna elements is 6, respectively. It shows that the MBER Pre-FFT beamformer has lower sidelobe levels and deeper nulls.

It is observed that the beam pattern performance of the MBER beamformer is better than that of the LMS beamformer. Note that the LMS beamformer appears to have a better amplitude response than the MBER beamformer. If the amplitude response alone would constitute the ultimate performance criterion of a beamformer, the MMSE beamformer would appear to be more beneficial. However, considering the magnitude alone can be misleading. It is shown in [3] in detail that the MBER solution has a better ability to cancel interfering signals.

Figure 9 shows that regarding the convergence performance of LMS based on MMSE Pre-FFT beamformer, we can see that after about 400 OFDM symbols we still don’t have convergence. Figure 10 shows that we still have mean square errors even after 400 OFDM symbols. As we note from Figure 11, the optimum weights and steady state MSE performance, respectively, can be obtained under the same conditions after about 5 OFDM symbols which results in much fewer calculations to reach convergence. For the simulation, shorter training OFDM symbols are required in the MBER Pre-FFT beamformer algorithm.

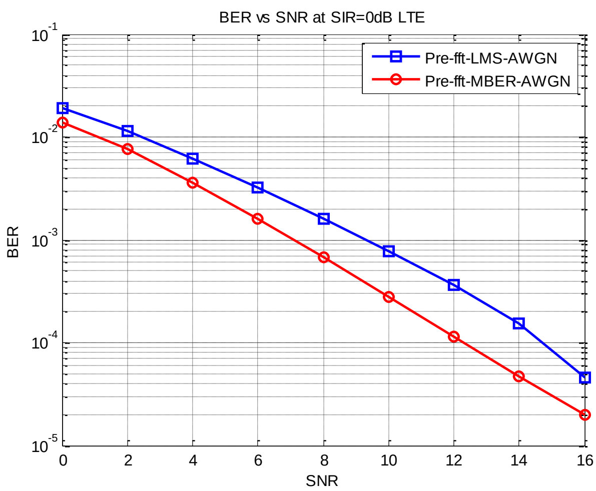

Figure 3. Comparison of the bit error performance.

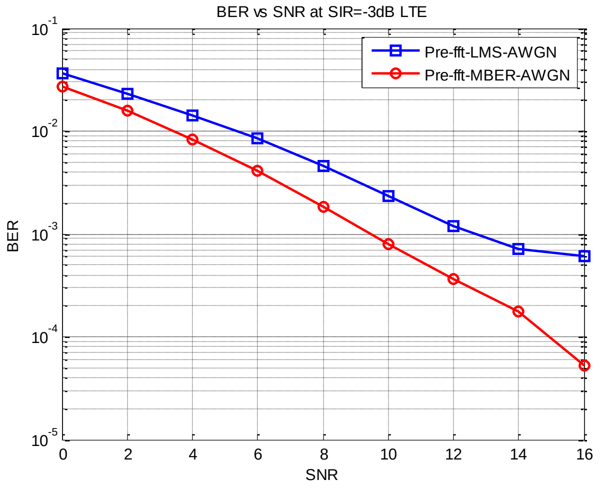

Figure 4. Comparison of the bit error performance.

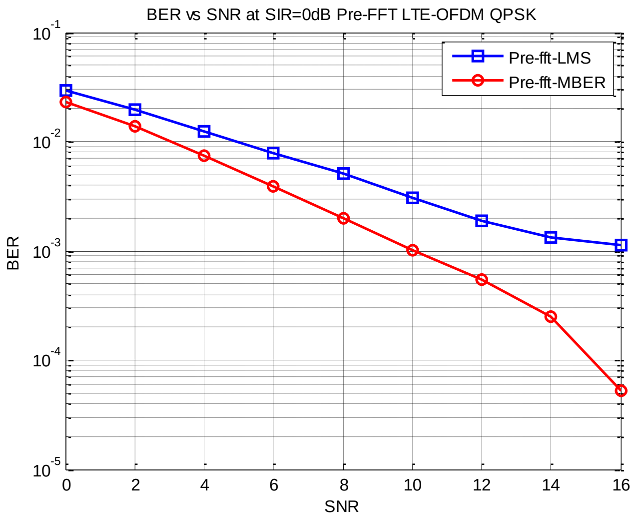

Figure 5. Comparison of the bit error performance.

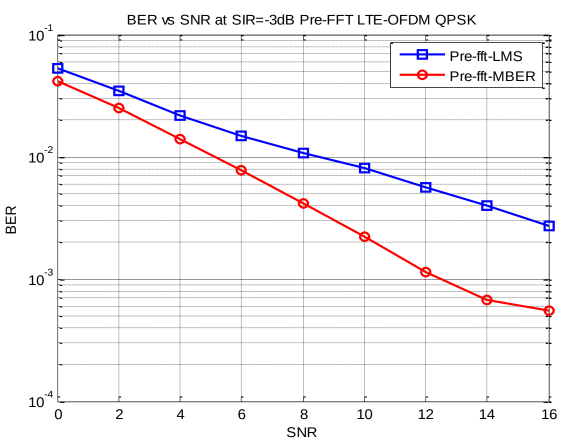

Figure 6. Comparison of the bit error performance.

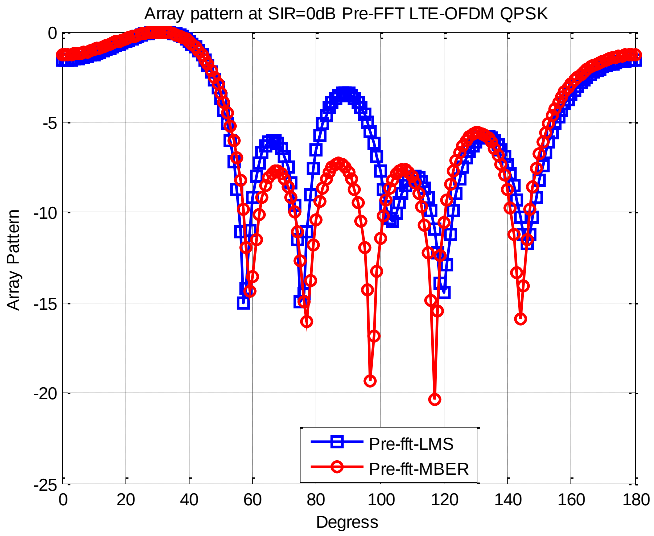

Figure 7. Beam pattern of the MBER and LMS.

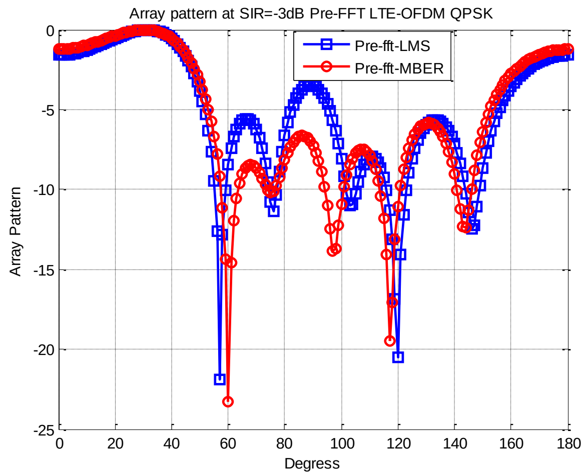

Figure 8. Beam pattern of the MBER and LMS.

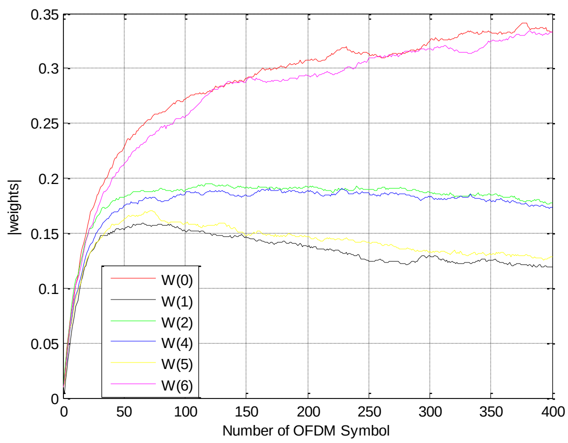

Figure 9. Convergence of the MMSE beamforming to obtain the optimum weights on the Pre-FFT performance.

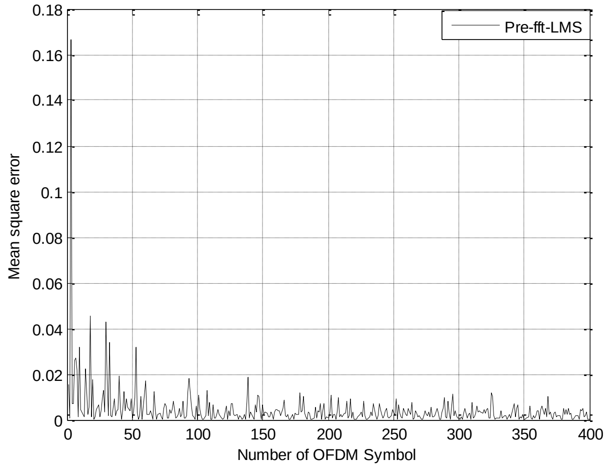

Figure 10. Mean square error of the LMS beamformer.

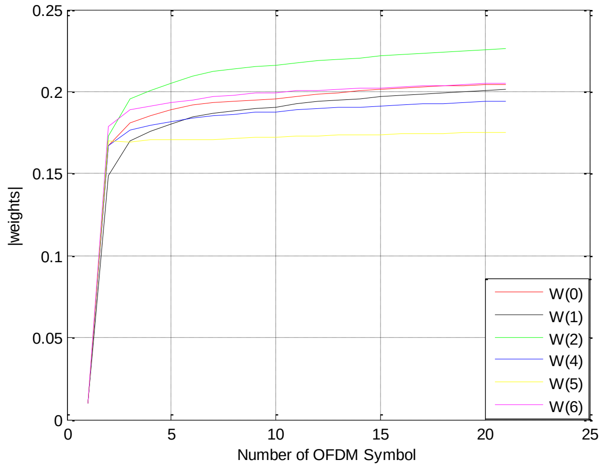

Figure 11. Convergence of the MBER beamforming to obtain the optimum weights on the Pre-FFT performance.

9. CONCLUSIONS AND FUTURE WORK

In this paper, BER performance is compared against the iteration index during adaptive implementation, SNR, beam pattern. When applying on LTE, simulation results show that better efficiency is achieved, because fewer pilots are used and as a result more data is transmitted. Also, the use of MBER resulted in quick conversion compared to MMSE. The proposed MBER yields a much better convergence speed while maintaining quadratic complexity. A proposed extension to this work would be to investigate the use of other modulation techniques, such as 16 QAM or 32 QAM, with a diamond shape pilot for different types of communication channels.

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ملخص العمل:

في هذا البحث، تمت دراسة أثر استخدام إشارة مرجعية على شكل معين (Diamond Shape Pilot Signal) لتقدير الإشارات المرسلة عبر قناة تستخدم نظام تقسيم الترددات المتعامدة (OFDM). ومن شأن هذا الترتيب أن يقلل من عدد الإشارات المرجعية المرسلة عبر قناة الاتصال ، وهو ما يؤدي بدوره إلى زيادة تدفق البيانات عبر القناة مع المحافظة على دقة مقبولة في تخمين القناة. وهنا يتم الجمع بين مصفوفة الهوائيات المتكيفة (Adaptive Antenna Array) ونظام تقسيم الترددات المتعامدة (OFDM) بهدف التغلب على التداخل بين الرموز (ISI) والتداخلات الاتجاهية.

في هذا البحث يتم الحصول على مجموعة الأوزان المثالية لتشكيل الشعاع بناءً على خوارزمية أدنى معدل الخطأ (MBER) مع استخدام إشارة مرجعية على شكل معين في أنظمة إرسال تقسيم الترددات المتعامدة (OFDM) لشبكات الاتصال للجيل الرابع (LTE)، حيث تم افتراض أن قناة الإرسال هي قناة خبو متعددة المسارات (Multipath Fading Channel). تبين نتائج المحاكاة أن إرسال الإشارات باستخدام الترميز التربيعي المعتمد على إزاحة الطور (QPSK) بناءً على خوارزمية أدنى معدل الخطأ (MBER) يستفيد من عناصر مصفوفة الهوائي على نحو أكثر فعالية من التقنية المعيارية القائمة على الحد الأدنى للخطأ في المتوسط التربيعي(MMSE).