Improving channel estimation using a modified root selection algorithm with model order estimation in OFDM systems

Ibrahim Adel Ghaleb
Electrical Eng. Dept., Faculty of Eng., Alexandria University, Alexandria, Egypt

In this paper a new algorithm for blind channel estimation in Orthogonal Frequency Division Multiplexing (OFDM) systems is described. The proposed algorithm uses a modification to the Root-Selection (RS) algorithm for channel estimation and a Generalized Akaike Information Criterion to estimate the channel length. The new algorithm is termed, the Modified Root-Selection (MRS) algorithm. This algorithm together with the channel order estimation effectively reduces the signal space of the estimator, and hence improves the estimation performance as demonstrated using computer simulations. The proposed MRS with channel order estimation algorithm has a 5dB lower mean square error in channel estimation when compared to the conventional approaches which translates to significant improvement in Signal-to-Noise Ratio (SNR) at receiver. In addition, the proposed modification to the RS algorithm results in a more computationally efficient algorithm. It is applicable not only to standard OFDM transmitters with cyclic prefix, but also to the recently proposed zero padded OFDM transmissions.

فى هذا البحث يتم اقتراح و تقديم خوارزم جديد ذو كفاءة حسابية عالية و اداء قوى لتحديد قناة الاتصال فى أنظمة الاتصالات ذات التقسيم الترددى المتعامد (OFDM). ويستخدم الخوارزم المقترح تعديلاً لخوارزم اختيار الجذور (RS) وتعميماً لمعيار معلومة أكيكى (AIC) لتقدير طول قناة الاتصال . وقد تم تسمية الخوارزم الجديد "الطريقة المعدلة لاختيار الجذور (MRS) " وهذه الطريقة بالإضافة إلى تحديد طول القناة تخفض فضاء الإشارة المؤثر فى خوارزم التقدير وبالتالى تحسن من الأداء . وقد تم التحقق من ذلك بعمل محاكاه باستخدام الحاسب الآلى لهذا الخوارزم . و قد تم التوصل الى تقليل متوسط الخطأ التربيعى بمقدار ٥ ديسبيل افضل من باقى الطرق التقليدية مم ينتج عنه تحسن ملحوظ فى قوة الاشارة بالنسبة الى كمية الضوضاء و التداخل عند المستقبل. بالإضافة إلى ذلك فأن الخوارزم المقترح ذو كفاءة حسابية أعلى من الطريقة التقليدية لاختيار الجذور ويمكن تطبيقه مع مختلف مواصفات الإرسال المستخدمة فى نظم ال OFDM الحالية (OFDM-ZP & OFDM-CP) .

Keywords: Orthogonal frequency division multiplexing, Wireless LANs, Blind channel estimation, High speed data

1. Introduction

Orthogonal Frequency Division Multiplexing (OFDM) has received considerable interest in the last few years for its advantages in high bit rate transmissions over frequency selective fading channels [1]. In OFDM systems, the high-rate data stream is divided into many low-rate streams that are transmitted in parallel, thereby increasing the symbol duration and reducing the inter-symbol interference (ISI). The ISI can be completely eliminated by introducing guard interval (Cyclic Prefix-CP) between adjacent OFDM symbols, given that the cyclic prefix length is greater than the length of channel impulse response [2]. The cyclically extended guard interval also converts linear convolution of signal and channel into circular convolution. As a result, a traditional complex Time-domain Equalizer (TEQ) can be replaced by a simple single tap Frequency domain EQualizer (FEQ). The channel estimation method based on a parametric channel model is proposed in [3] which uses Minimum Description Length (MDL) to estimate the channel order and Estimation of Signal Parameters via Rotational Invariance Technique (ESPRIT) to acquire mutipath time delays. Most Significant Taps (MST) approach is used in [4] to estimate the channel order. In [5], an algorithm was proposed to estimate sparse channels using the Generalized Akaike Information Criterion (GAIC). However, all of the above mentioned methods require the use of known pilot symbols at the receiver. Because they save bandwidth and are capable of tracking slow channel variations, blind channel estimation and equalization methods are well motivated as they avoid the use of training sequences [6], and [7]. A numerous blind channel estimation and blind equalization algorithms were developed in the literature (cf. [8] for a survey on these algorithms and the references therein).

In this paper, a new blind channel estimation method is developed that is applicable to both CP- and ZP-OFDM transmissions and relies on the finite alphabet property of information bearing symbols. The new Modified Root selection (MRS) method proposes a modification to the Root-Selection algorithm of [9]. The root-selection algorithm becomes very complex as the channel order increases. The MRS algorithm is highly simpler than the root-selection algorithm and can be applied for channels of higher order which can not be estimated using the root-selection algorithm. It also uses the Generalized Akaike Information Criterion (GAIC) to estimate the channel order. The MRS algorithm together with the channel order estimation yields a significant improvement in normalized mean squared error in channel estimation. The low complexity of the new algorithm allows for further improvement of the channel estimation accuracy through the use of Phase Directed (PD) steps as will be described in section (V).

2. Notations and system model

In the following, let $(.)^T$ indicate the transpose and $(\tilde{\cdot})$ indicates frequency domain samples. The MATLAB's notation **A**(*i:j,k:l*) indicates the submatrix formed from the *i*th to the *j*th rows and the *k*th to the *l*th columns of the matrix **A**, and indicates element by element matrix multiplication.

The system block diagram is shown in fig. 1. The data symbols are transmitted in blocks of size *M*: $\tilde{\mathbf{s}}_M(i) = [\tilde{s}_0(i),\ldots,\tilde{s}_{M-1}(i)]^T$, where *i* is the block index. These symbols are first precoded using the *M*x*M* IFFT matrix $\mathbf{F}_M^H = \mathbf{F}_M^{-1}$ with entry at the (m,n) element $= (1/\sqrt{M})exp(-j(2\pi/M)(m-1)(n-1))$ to yield the time domain block $\mathbf{s}_M(i)$. Then the cyclic prefix (CP) is appended between each block of symbols. The entries of the resulting redundant block $\mathbf{s}_{CP}(i)$ of length $P=M+L_{CP}$ are transmitted sequentially through the frequency selective fading channel **h.** The channel is modeled as an FIR filter with channel impulse response $h = [h_0,\ldots,h_L]^T$ where it is assmed that $L \leq L_{CP}$. We also define the *M*x1 vector $h_M = [h_0,\ldots,h_L,0,\ldots,0]^T$. Let $\mathbf{I}_{CP} = [\mathbf{I}_C, \mathbf{I}_M]^T$ be the *P*x*M* matrix representing the cyclic prefix appending where $\mathbf{I}_C$ represents the L_{CP} last columns of $\mathbf{I}_M$ (*M*x*M* identity matrix), we also define the *P*x*M* matrix $\mathbf{F}_{CP} = \mathbf{I}_{CP}\mathbf{F}_M^H$, now the transmitted block is given by:

$$\mathbf{s}_{CP}(i) = \mathbf{I}_{CP}\ \mathbf{F}_M^H\ \tilde{\mathbf{s}}_M(i) = F_{CP}\ \tilde{\mathbf{s}}_M(i)\,, \qquad (1)$$

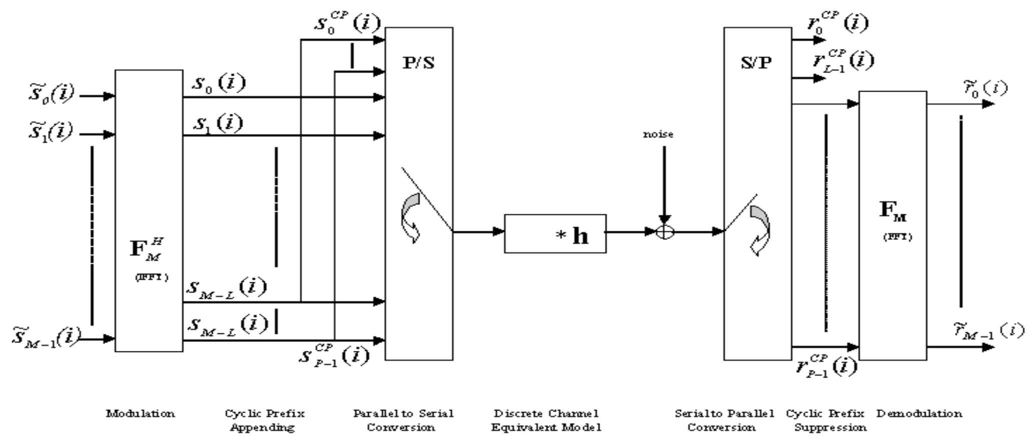

Fig. 1. The OFDM-CP baseband model.

At the receiver end, first the cyclic prefix is removed (and so is the IBI (inter block interference) assuming that the channel is shorter than the cyclic prefix). Denoting by $\mathbf{r}_M(i)$ the received Mx1 vector after suppressing the CP: $r_M(i) = [r_L^{CP}(i),\ldots,r_{P-1}^{CP}(i)]^T$ and the noise is given by : $n_M(i) = [n_L(i),\ldots,n_{P-1}(i)]^T$.

Due to the appending of the cyclic prefix the channel effect is circularized and it can be diagonalized using FFT at the receiver [10]. So we get:

$$\begin{aligned}\tilde{\mathbf{r}}_M(i) &= \mathbf{F}_M\ \mathbf{r}_M(i)\\ &= \sqrt{M}\ diag\ (F_M h_M)\ \tilde{s}_M(i)\ + \mathrm{F}_M\ \mathrm{n}_M(i)\\ &= \sqrt{M}\ diag\ \ (\mathrm{F}_M \mathrm{h}_M)\ \tilde{s}_M(i) + \tilde{\mathrm{n}}_M(i). \qquad (2)\end{aligned}$$

Where $\sqrt{M}\ diag\ (\mathbf{F}_M\mathbf{h}_M)$ is a diagonal matrix with diagonal entries (k,k) = $\mathbf{H}(e^{j\frac{2\pi}{M}(k-1)})$, where $\mathbf{H}(e^{j\frac{2\pi}{M}k})$ is the channel attenuation at carrier k. We now define the channel attenuation at the carriers corresponding to the CIR $\mathbf{h}_M$ as: $\tilde{\mathbf{h}}_M = \sqrt{M}\ \mathbf{F}_M\ \mathbf{h}_M = [\mathrm{H}(0), \mathrm{H}(e^{j\frac{2\pi}{M}}),\ldots,\mathrm{H}(e^{j\frac{2\pi}{M}(M-1)})]^T$, Eq. (2) can be re-expressed as:

$$\tilde{\mathbf{r}}_M(i) = diag(\tilde{\mathbf{h}}_M)\tilde{\mathbf{s}}_M(i) + \tilde{\mathbf{n}}_M(i) = \tilde{\mathbf{h}}_M \tilde{\mathbf{s}}_M(i) + \tilde{\mathbf{n}}_M(i). \qquad (3)$$

3. Blind channel estimation

3.1. Introduction

In this section, a new blind channel estimation algorithm is proposed. The finite alphabet property of the transmitted data symbols is used to get estimates of $\mathbf{H}^Q(e^{j\frac{2\pi}{M}k})$ and $k \in [0, M-1]$.

Based on this finite alphabet property of the transmitted symbols, a variety of algorithms were proposed in [9], and [11]. All of these algorithms are very complex and cannot be used practically for channels of high order (for example in Hiperlan/2 in which L=16). So a modification is proposed to the root-selection algorithm of [9] that results in a simpler algorithm with improvement in performance.

The new algorithm is developed under the following assumptions (also used in [9]):

1. Symbols are drawn from finite alphabet set of size Q; i.e., $\tilde{s}_k(i) \in \{\xi_q\}_{q=1}^{Q}$.
2. Noise is zero-mean complex Gaussian noise and independent of the data.

For PSK modulation we have $E\{\tilde{s}_k^Q(i)\} = 1$ deterministically. Starting from $E\{\tilde{r}_k^Q(i)\} = \mathbf{H}^Q(e^{j\frac{2\pi}{M}k})E\{\tilde{s}_k^Q(i)\}$ we deduce that (assuming noise free system):

$$\mathbf{H}^Q(e^{j\frac{2\pi}{M}k}) = E\{\tilde{r}_k^Q(i)\}, \forall k \in [0, M-1]. \qquad (4)$$

The right hand side can be obtained for each k deterministically for PSK modulation. The question now is how to get $\mathbf{H}(e^{j\frac{2\pi}{M}k})$ from $\mathbf{H}^Q(e^{j\frac{2\pi}{M}k})$. To express $\mathbf{H}^Q(e^{j\frac{2\pi}{M}k})$ in terms of $\mathbf{h}$, $\beta_Q^T = [\beta_0,\ldots,\beta_{QL}] = \mathrm{h}^T *_Q \mathrm{h}^T$ is first defined as the Q-fold convolution of the channel with itself. Since time domain convolution corresponds to multiplication in the frequency domain, we can write: $\mathbf{H}^Q(e^{j\frac{2\pi}{M}k}) = \mathbf{H}^Q(z)\big|_{z=e^{j\frac{2\pi}{M}k}} = \left(\beta_0 + \ldots + \beta_{QL} z^{-QL}\right)\big|_{z=e^{j\frac{2\pi}{M}k}}$. To determine the coefficients $\beta_0,\ldots,\beta_{QL}$ uniquely from $\mathbf{H}^Q(e^{j\frac{2\pi}{M}k})$, we need QL+1 equations that can be available if $M \geq QL+1$. In practice (noisy case), $E\{\tilde{r}_k^Q(i)\}$ is replaced by statistical averaging and thus $\mathbf{H}^Q(e^{j\frac{2\pi}{M}k})$ can be estimated as:

$$\hat{\mathbf{H}}^Q(e^{j\frac{2\pi}{M}k}) = \left(\frac{1}{N}\sum_{i=0}^{N-1} \tilde{r}_k^Q(i)\right), k \in [0, M-1], \qquad (5)$$

where N is the total number of blocks used in the averaging. $\hat{\mathbf{H}}^Q(e^{j\frac{2\pi}{M}k})$ is next collected from eq. (5) in an Mx1 vector:

$\hat{\tilde{\text{H}}}_Q = [\hat{\text{H}}^Q(0),...,\hat{\text{H}}^Q(e^{j\frac{2\pi}{M}(M-1)})]^T$. Define the matrix $\mathbf{V}_Q$ to be a scaled version of the first QL+1 columns of the FFT matrix $\mathbf{F}_M$ as follows $\mathbf{V}_Q = \sqrt{M}\mathbf{F}_M(:,1:QL+1)$ then $\boldsymbol{\beta}_Q$ can be estimated by a simple matrix inversion as:

$$\hat{\boldsymbol{\beta}}_Q = \mathbf{V}_Q^t \hat{\tilde{\mathbf{H}}}_Q = \frac{1}{M}\mathbf{V}_Q^H \hat{\tilde{\mathbf{H}}}_Q . \quad (6)$$

3.2. The root-selection algorithm

Based on the fact that $\boldsymbol{\beta}_J$ contains the L roots of $\mathbf{h}$ with multiplicity J, one can search the JL roots of $\boldsymbol{\beta}_J$ and try all the different combinations of L roots out of the JL roots to find the best L roots that minimize the following cost function:

$$\hat{\mathbf{h}} = \arg\min_{\hat{\mathbf{h}}} \left\| \hat{\mathbf{h}}^T *_J \hat{\mathbf{h}}^T - \boldsymbol{\beta}_J \right\|^2 . \quad (7)$$

The total number of trial required is given by: $\mathbf{C}_L^{JL} = \dfrac{JL!}{(JL-L)!\,L!}$.

3.3. The modified root-selection algorithm

The RS algorithm can only be applied for low order channels because the complexity increases as the order of the channel increases. For example, if BPSK modulation (Q=2) is used with L=8, the number of iterations required by the RS $= C_L^{JL} = C_8^{16} = 12{,}870$ and if QPSK (Q=4) is used the number of iterations increases to be $C_8^{32} \approx 10.5$ million iterations. In this section we propose a simple modification to the RS algorithm that results in a much simpler and computationally efficient algorithm. Starting from the fact that $\boldsymbol{\beta}_J$ contains all the L roots of $\mathbf{H}(z)$ with multiplicity J, then we propose first to make L groups each contains J roots. The grouping can be done using the following approach:

1. Select from the JL roots of $\boldsymbol{\beta}_J$ the root with the minimum absolute value.
2. Select the J-1 roots that are closest to the selected root in step 1. These roots form the first group.
3. Repeat step 1 and 2 on the remaining set of roots.

At the end we will have L groups each containing J roots. We estimate the L roots of the channel $\mathbf{h}$ as the average of the J roots within each group. So if the roots in the lth group are given as: $[R_l^1,...,R_l^J]$ then the lth channel root R_l^{ch} can be estimated as:

$$R_l^{ch} = \frac{1}{J}\sum_{j=1}^{J} R_l^j \quad \text{for} \quad l = 1,...,L . \quad (8)$$

The MRS algorithm described above is much more computational efficient than the RS algorithm, however, both algorithms are highly sensitive to the correct model order estimate. In most previous work the channel order L was assumed to be equal to its upper bound, the length of the cyclic prefix L_{CP}. The accurate knowledge of the correct channel length will help in significantly improving the performance of the MRS estimator. In this paper, the GAIC criterion is used to estimate the length of the channel. This is discussed in the next section.

4. The channel order estimation

GAIC has been a popular statistical criterion for model structure selection in system identification. It has a cost function of the form [12]

$$GAIC(L) = V_L + \gamma\, ln(ln(M))(L+1) . \quad (9)$$

Where the first term reflects the modeling error and the second term is the penalty function. Here γ is a parameter which the user can choose, and is our simulations we have chosen γ = 2. For our system, the expression for the modeling error V_L is given by

$$V_L = \frac{N}{2} ln(\hat{\sigma}_{n,L}^2) . \quad (10)$$

Where $\hat{\sigma}_{n,L}^2$ is the estimate of the noise variance for channel length L and is given by:

$$\hat{\sigma}_{n,L}^2 = \frac{1}{2L-1}(\hat{h}_L^T *_J \hat{h}_L^T - \beta_J)^H (\hat{h}_L^T *_J \hat{h}_L^T - \beta_J). \quad (11)$$

Where $\hat{h}_L$ is the MRS channel estimate of the channel $\mathbf{h}$ assuming channel length L. The GAIC estimate of the true channel length is obtained by minimizing (9) with respect to *L*. The following steps constitute the GAIC test,
Step 1. Initially set the limit $P = L_{cp}$
Step 2. Calculate the cost function GAIC(L) for *L = 1,2,…,P.*
Step 3. The GAIC estimate of *L* is then obtained as

$$\hat{L} = \arg\min_L \{GAIC(L)\}. \quad (12)$$

5. Phase directed algorithm

The channel estimation accuracy of the MRS algorithm described above can be further improved through what is termed by Phase Directed (PD) algorithm [9]. As described in [9], for each *k* we obtain $\hat{\mathbf{H}}(e^{j\frac{2\pi}{M}k}) = \lambda_k \left[\hat{\mathbf{H}}^J(e^{j\frac{2\pi}{M}k})\right]^{1/J}$ up to a scalar ambiguity $\lambda_k \in \{e^{j(2\pi/J)n}\}_{n=0}^{J-1}$. Suppose that initial estimates $\hat{\mathbf{H}}_0(e^{j\frac{2\pi}{M}k})$ are available through a the low complexity MRS algorithm. Then, for each $k \in [0, M-1]$, we can resolve the phase ambiguity by searching over the *J* candidates phase values.

$$\hat{\lambda}_k = \arg\min_{\lambda_k} \left| \hat{\mathbf{H}}_0(e^{j\frac{2\pi}{M}k}) - \lambda_k \left[\hat{\mathbf{H}}^J(e^{j\frac{2\pi}{M}k})\right]^{1/J} \right|^2 . \quad (13)$$

Therefore, we can improve channel estimation accuracy through the Directed PD steps that can be described as follows:

1. Get an initial estimate $\hat{\mathbf{h}}_0$ of the channel using the low complexity MRS method, and calculate the frequency domain response $\hat{\mathbf{H}}_0(e^{j\frac{2\pi}{M}k})$ for each $k \in [0, M-1]$.
2. Resolve phase ambiguities using eq. (13) and replace each $\hat{\mathbf{H}}_0(e^{j\frac{2\pi}{M}k})$ with $\hat{\lambda}_k \left[\hat{\mathbf{H}}^J(e^{j\frac{2\pi}{M}k})\right]^{1/J}$ each $k \in [0, M-1]$ and form the vector. $\hat{\tilde{h}}_1 = [\hat{\lambda}_0 [\hat{H}^J(0)]^{1/J}, \ldots, \hat{\lambda}_{M-1} \left[\hat{H}^J(e^{j\frac{2\pi}{M}(M-1)})\right]^{1/J}]^T$.
3. Update the channel estimates in the time domain $\hat{\mathbf{h}}_1 = \frac{1}{M}\mathbf{V}_1^H \hat{\tilde{\mathbf{h}}}_1$, their frequency domain response using $\hat{\tilde{\mathbf{h}}}_{1new} = \mathbf{V}_1 \hat{\mathbf{h}}_1$.
4. Repeat steps 1, 2 and 3 with $\hat{\mathbf{h}}_0 = \hat{\mathbf{h}}_1$ from the previous iteration, (say I_1) times or until the Euclidean norm of the variation in the channel is very small.

In the simulations described in the next section, it will be shown that using the PD algorithm with only one iteration can significantly improve the channel estimation accuracy. It should be noted that each iteration of step 3) entails one M-point inverse FFT and one M-point FFT. Thus the PD iterations have low computational complexity. It is also worth recalling that $J = 2$ for BPSK and hence $\lambda_k \in \{\pm 1\}$, while for QPSK, 16QAM and 64QAM, $J = 4$ so that $\lambda_k \in \{\pm 1, \pm j\}$. These few phase values can be resolved easily via eq. (13).

6. Simulation results

Several OFDM system were simulated using *M*=32, and 64 useful sub-carriers with cyclic prefix $L_{CP} = 8$, and 12 samples. The data was taken from BPSK and QPSK constellations. The performance of the channel estimation algorithm is evaluated for a channel model with exponential power delay profile with 4 significant taps as used in [11] and [13]. The performance is averaged over 800 independent quasi-static channel realizations. Channel order estimation was first done using the GAIC criterion as described in section 4. Then channel estimation is performed using the proposed MRS algorithm with the GAIC channel order estimation technique, and compared to the subspace algorithm of [13]. Their performances were compared in terms of the Time-Domain Normalized Mean Square Error (TD NMSE) defined as:

$$\text{TD NMSE} = \sum_{k=0}^{L}\left|h_k - \hat{h}_k\right|^2 / \sum_{k=0}^{L}\left|h_k\right|^2 .$$

Fig. 2 shows a histogram describing the accuracy of the proposed channel order estimation algorithm using a 4-tap channel as described above. For each E_b/N_o value the experiment was repeated 1000 times and in each time the channel order was estimated. Results show that for SNR above 15dB the algorithm produces the correct estimate MM=4 more than 99% of the time. Even for low SNR values, the algorithm works considerably well, with only few times over-estimating or under-estimating the channel order by 1-tap. Figs. 3-a and 3-b show the performance of the MRS algorithm for different values of the assumed channel order (MM), for the cases of BPSK and QPSK modulations, respectively. In these simulations, the channel estimation was done using the MRS algorithm assuming the channel orders MM = 4 (correct), 6, 8, 10 and 12. MM = 12, corresponds to the length of the cyclic prefix, which is usually the value used in most systems, when no channel order estimation is performed. As we can see in the figures, over-estimating the channel order significantly degrades the MRS channel estimation performance. The figures also show the performance of the proposed MRS algorithm when used together with the GAIC channel order estimation technique. The performance results in this case are almost the same as the case when the exact channel order estimate was assumed, i.e. MM = 4.

The results shown in figs. 4-a and 4-b are similar to those in fig. 3 but for the case were the MRS channel estimation algorithm is followed by PD steps as described in section 5. Only a single PD iteration was used in the simulations. The results also show the sensitivity of the algorithm estimates to the

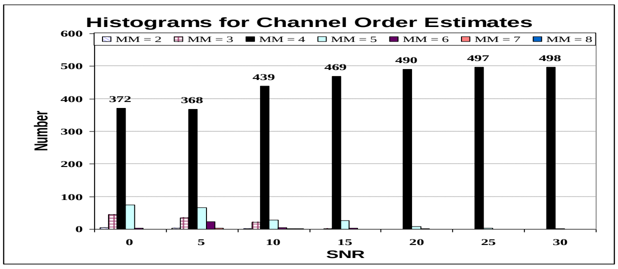

Fig. 2. Histogram for evaluating performance of the channel order estimate at different SNR.

the correct channel order, and show also a significant degradation when the channel order is over-estimated. However comparing results in fig. 4 to those in fig. 3, show the performance improvement in the channel estimation accuracy when PD steps follow the initial MRS estimate.

Figs. 5-a and 5-b compare the performance of the proposed MRS method with the GAIC model order estimation technique followed by PD iterations of [9] compared to the subspace method presented of [13] and [14] for BPSK and QPSK modulation. It should be noted that the subspace method requires a certain number of received blocks (2P for OFDM-CP and P for OFDM-ZP) to be applied. The subspace method achieves a performance that is constellation independent on the expense of higher complexity (2 SVD of large matrices) and higher memory requirement for solving the resulting system of equations. The proposed method is highly less complex. In addition, Fig. 5-a shows the improvement in the performance of the proposed algorithm to the subspace method, even with no PD iterations for BPSK modulation. In the case of QPSK, Fig. 5-b shows that the subspace method performs better than the MRS algorithm with no PD iterations. However, the improvement in the proposed MRS with PD clearly exceeds the performance of the subspace method. It should be noted that the PD iterations have low computational complexity as was explained in section 5.

7. Conclusions

A modification to the Root Selection (RS) algorithm is proposed leading to the Modified

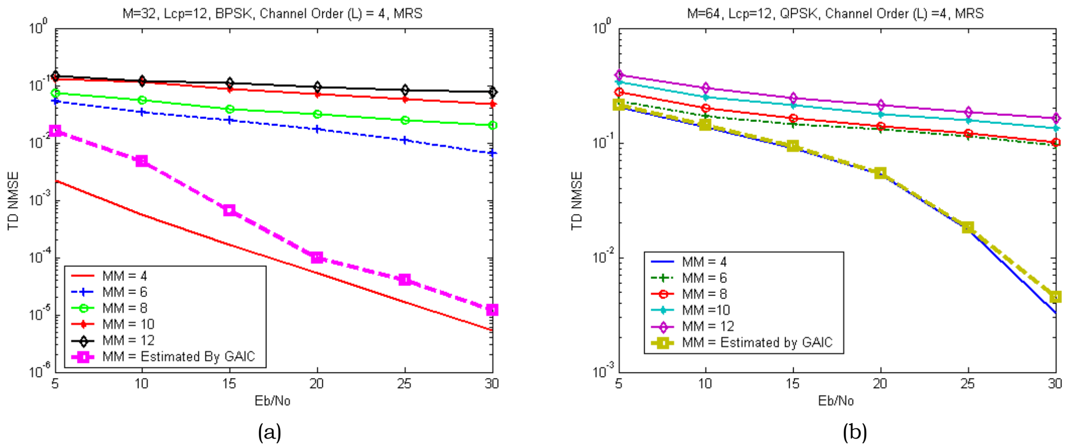

Fig. 3. Performance of the proposed MRS algorithm using different assumed channel orders, when the actual channel order is 4. (a) for BPSK, (b) for QPSK.

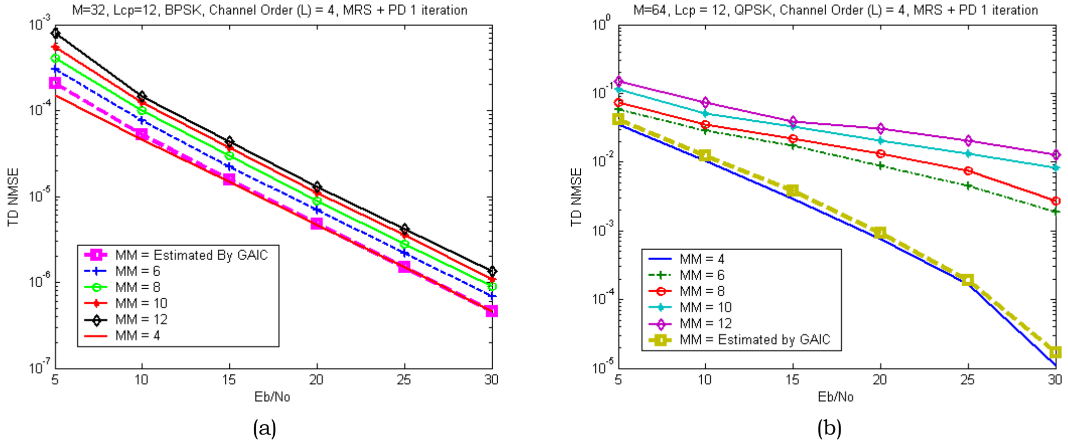

Fig. 4. Performance of the proposed MRS algorithm with a single phase directed iteration, using different assumed channel orders, when the actual channel order is 4. (a) for BPSK, (b) for QPSK.

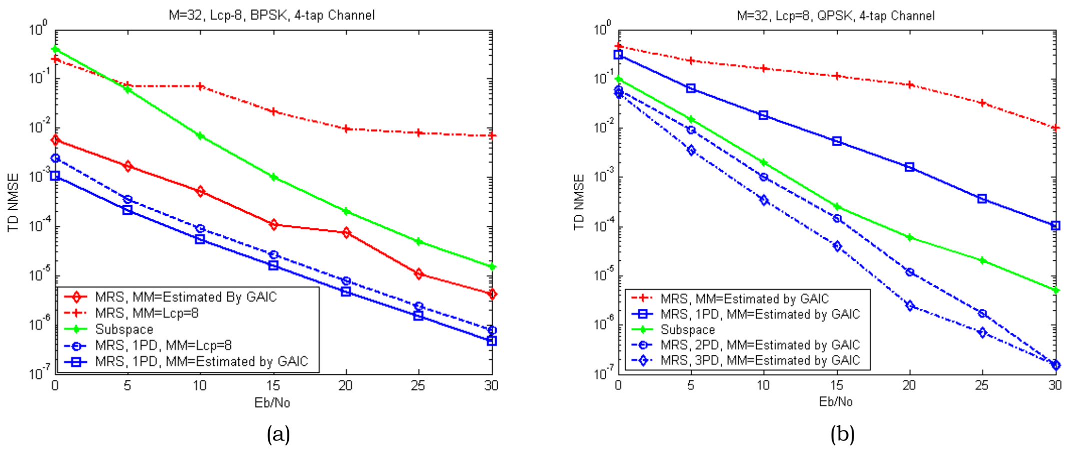

Fig. 5. Comparing the performance of the proposed MRS algorithm with channel order estimation to the subspace algorithm: (a) for BPSK, (b) for QPSK.

Root Selection (MRS) algorithm. The RS algorithm is based on finite alphabet property of the transmitted symbols. The new algorithm is highly less complex than the RS algorithm and can be applied for channels of high order. The RS becomes very complex and impractical for high order channels. The proposed MRS algorithm uses a combination of the RS algorithm and the GAIC technique to blindly estimate the channel order then estimate the channel coefficients. The low complexity of the proposed MRS algorithm allows for further improvement of the channel estimation accuracy through the use of Phase Directed (PD) steps. The performance of the proposed modified algorithm is compared with the subspace method of [13], [14]. The subspace achieves a performance that is constellation-independent but the new method is highly less complex and is independent of the channel zero locations. The subspace method fails if one (or more) channel zero is located on a sub-carrier [13].

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