SINR Maximizing Equalizer Design for OFDM Systems

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***Abstract*--A novel time domain equalizer (TEQ) design approach is proposed in this paper based on maximizing signal-to interference plus noise ratio (SINR). To increase the bandwidth efficiency of OFDM systems, the TEQ is used to mitigate the intersymbol interference (ISI) created by a channel with longer impulse response duration than that of the cyclic prefix (CP). The true ISI, ICI and noise parts of the signal at the output of equalizer are formulated based on the overall impulse response (OIR) (convolution of channel impulse response and equalizer impulse response). SINR Maximizing Time-domain Equalization (SMTE) method estimates the tap coefficients of the equalizer by maximizing the SINR of the signal at the output of equalizer. Computer simulation and analytical results show that the performance of the SMTE method is superior to the performance of the much-used method of channel impulse response shortening (IRS).**

I. INTRODUCTION

Orthogonal frequency division multiplexing (OFDM) is a multicarrier modulation scheme that partitions a broadband channel into a number of parallel and independent narrowband subchannels. A cyclic prefix (CP) is inserted between the OFDM symbols in order to mitigate intersymbol interference (ISI) and interchannel interference (ICI) effects for signal transmission over multipath channels. To avoid ISI and also ICI (and thus make the subchannels orthogonal), the CP interval should not be shorter than the channel impulse response (CIR) interval. However, appending the CP to each OFDM symbol decreases the bandwidth efficiency of the OFDM system; thus the CP interval should be a small fraction of the OFDM symbol interval in order not to lose the bandwidth efficiency.

In this paper, we propose a novel method to design the TEQ for the OFDM systems based on maximizing SINR. At the first step, we formulate the criterion by identifying the true ISI, ICI and noise powers at the output of equalizer and then define true SINR being different than those used in earlier literature [3]. Also, we point out that the contribution of each tap coefficient of the OIR to the SINR depends on its time index in addition to its power. At the next step, the equalizer tap coefficients are estimated by maximizing the SINR.

System model and criteria design of the TEQ based on impulse response shortening (IRS) and the MMSE methods are presented in Section II. Shortcoming of the IRS and the MMSE methods is highlighted in this section as well. The SINR Maximizing Time-domain Equalization (SMTE) method is proposed in Section III. Section IV provides computer simulations. Some conclusions are drawn in Section V.

II. SYSTEM MODEL AND TIME-DOMAIN EQUALIZER DESIGN CRITERION

The discrete model of an OFDM system with a TEQ is shown in Fig. 1. $\mathbf{q}(k)=[q_0(k),...,q_{N-1}(k)]^T$ is an OFDM symbol in the frequency domain such that each $q(k)$ is selected from a q-ary constellation and $(.)^T$ represents the transpose operation. $\mathbf{x}(k)=[x_0(k),...,x_{N-1}(k)]^T$ is an N-point inverse fast Fourier transform (N-IFFT) of $\mathbf{q}(k)$. $\mathbf{s}(k)=[x_{N-\nu}(k),...,x_{N-1}(k),x_0(k),...,x_{N-1}(k)]^T$ is the transmitted symbol after appending a CP with length ν to the $\mathbf{x}(k)$. $h(l)$, $l=0,...,L-1$ is the CIR with length L. $f(l)$, $l=0,...,M-1$ is the impulse response of the equalizer with length M. $n(l)$ is a zero mean additive white Gaussian noise with autocorrelation function $R_n(l)=\sigma_n^2\delta(l)$. The OIR is defined as $g(l)=h(l)*f(l)$, where '$*$'stands for the convolution operation. Duration of the OIR is $L_o=L+M-1$. Although the OIR interval becomes even longer than L, the CIR interval, the idea of using a TEQ is to compress the OIR energy within a given window with length of the CP interval plus one or, in other words, to shorten the effective length of the OIR.

If $g_{in}(l)$ indicates the part of the OIR within the CP interval and $g_{out}(l)$ indicates the remainder part, the impulse response of equalizer, $f(l)$, has been designed by the first approach in [1] such that the energy of $g_{out}(l)$, E_{out}, is minimized under the constraint of the normalized energy for $g_{in}(l)$, $E_{in}=1$. The main motivation of this criterion in the IRS method comes from the idea that $g_{out}(l)$ causes ISI and if E_{out} is minimized the ISI power is minimized as well. Based on [1], another method has been proposed in [2] such that E_{in} is maximized while satisfying the constrain $E_{out}=1$. In addition to the E_{in} and E_{out}, the power of noise at the output of equalizer has been considered in [3] for the equalizer design criterion. The sum of ISI power and noise power of all subcarriers is minimized subject to $E_{in}=1$ in [3].

The second approach considers a desired shortened CIR with the length of the CP interval plus one and designs a TEQ based on minimizing the mean square error between the OIR and the desired shortened CIR [4]. The criterion of choosing the desired shortened channel impulse response is the MMSE under unit-energy constraint or unit-tap constraint [4] or maximizing the approximated system capacity dictated by the SNRs of subcarriers [5].

In both approaches, the energy of each tap coefficient of the OIR (in the first approach) or the desired CIR (in the second

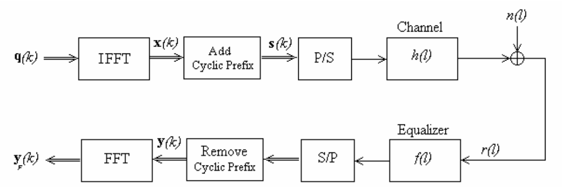

Fig. 1. A discrete model of an OFDM system with a time domain equalizer.

approach) is weighted by a rectangular window and then a criterion is used to design the equalizer. In other words, the energy of each OIR tap exceeding the CP interval is assumed to cause the same impact on the ISI power. Before analyzing the impact of the ISI and ICI on the OFDM system performance, we would like to highlight the ISI and ICI effects of each OIR tap exceeding the CP interval by giving a simple example.

Let us assume that the OIR is $g(l)=\delta(l)+a_0\delta(l-L_D)$ where L_D is larger than the CP interval. It is clear that the tap related to $a_0\delta(k-L_D)$ creates ISI and ICI. We consider that $L_D=\nu+\Delta$ where $\Delta>0$ and evaluate the ISI power based on different values of Δ. The output signal at the output of the equalizer is given by

$$z(l)=s(l)*g(l)=s(l)+a_o s(l-L_D)$$

After removing the cyclic prefix period, $z_i(k)$ (ith element of the kth OFDM symbol at the output of the equalizer, $\mathbf{z}(k)$) without considering the noise will be

$$\begin{aligned}
&z_0(k)=x_0(k)+a_0x_{N-\Delta}(k-1),\\
&z_1(k)=x_1(k)+a_0x_{N-\Delta+1}(k-1),\\
&\cdots\\
&z_{\nu+\Delta}(k)=x_{\nu+\Delta}(k)+a_0x_0(k),\\
&\cdots\\
&z_{N-1}(k)=x_{N-1}(k)+a_0x_{N-\nu-\Delta-1}(k).
\end{aligned}$$

As can be seen, only $z_i(k)$ for $0\le i\le\Delta-1$ has been contaminated by ISI and ICI. In order to clarify the part of ISI and ICI, $z_i(k)$ for $0\le i\le\Delta-1$ can be written as

$$z_i(k)=x_i(k)+a_0x_{N-\nu-\Delta+i}(k)-a_0x_{N-\nu-\Delta+i}(k)+a_0x_{N-\Delta+i}(k-1)$$

The first two terms of $z_i(k)$ represents the signal part that would happen in a scenario when the CP interval is longer than the CIR interval and consequently the third term of $z_i(k)$ represents the ICI part. Also, the fourth term of $z_i(k)$ indicates the ISI part. If we assume that the transmitted signal samples are independent and $E[|x_m(k)|^2]=\sigma_x^2$ for all "k" and "m", the total power of ISI and ICI in each OFDM symbol is a function of Δ and it is given by

$$P_{ISI+ICI}(\Delta)=2\Delta\,|a_0|^2\,\sigma_x^2 \qquad (1)$$

As (1) shows, the total power of ISI and ICI not only depends on the transmitted signal power (σ_x^2) and the channel tap coefficient power ($|a_0|^2$), but also depends on $\Delta=L_D-\nu$ which indicates the time index of the OIR tap appearing outside the CP interval. In addition, although L_D is longer than the CP interval ($L_D>\nu$), the tap coefficient "a_0" corresponding to the delay "L_D" contributes to the power of signal, $z_i(k)$, for $i\ge L_D-\nu$ without creating ISI and ICI. Thus *minimizing the total power of the OIR part that appears outside the CP interval (E_{out}) is not enough to achieve good performance.* Moreover, *an OIR with less E_{out} may cause more ISI and ICI and give poorer performance in comparison with another OIR which has a larger E_{out}*. By the same token, *the MMSE criterion does not endorse the strategy of decreasing ISI and ICI power.* Thus, regarding the part of OIR exceeding the CP interval, the TEQ design criterion should be based on maximizing the SINR that considers true ISI, ICI and noise power at the output of the equalizer.

III. SINR MAXIMIZING TIME-DOMAIN EQUALIZATION METHOD

We formulate the ISI, ICI and noise part of the signal at the output of equalizer and then estimate equalizer impulse response (tap coefficients) by maximizing the SINR in this section. Assuming that the OIR interval is less than the OFDM symbol duration, $L_o\le N$ [1], the kth OFDM symbol at the output of equalizer after removing the CP period is given by

$$\mathbf{y}(k)=\mathbf{G}\mathbf{x}(k)-\mathbf{A}\mathbf{x}(k)+\mathbf{B}\mathbf{x}(k-1)+\boldsymbol{\eta}(k) \qquad (2)$$

where $\mathbf{y}(k)=[y_0(k),y_1(k),...,y_{N-1}(k)]^T$ and $\boldsymbol{\eta}(k)=[\eta_0(k),\eta_1(k),...,\eta_{N-1}(k)]^T$ which is the colored noise vector at the output of equalizer. $\mathbf{G}$ is an $N\times N$ circular matrix and indicates the part of signal which is free of ISI and ICI (assuming that the CP interval is enough long). $\mathbf{G}$ is defined as

$$\mathbf{G}=\begin{bmatrix}
g_0 & 0 & \cdots & 0 & g_{L_o-1} & g_{L_o-2} & \cdots & g_1\\
g_1 & g_0 & 0 & \cdots & 0 & g_{L_o-1} & \cdots & g_2\\
\vdots & \ddots & \ddots & \ddots & \ddots & \ddots & \ddots & \vdots\\
g_{L_o-1} & g_{L_o-2} & \cdots & g_0 & 0 & \cdots & \cdots & 0\\
0 & g_{L_o-1} & g_{L_o-2} & \cdots & g_0 & 0 & \cdots & 0\\
\vdots & \ddots & \ddots & \ddots & \ddots & \ddots & \ddots & \vdots\\
0 & \cdots & 0 & g_{L_o-1} & g_{L_o-2} & \cdots & \cdots & g_0
\end{bmatrix}$$

where $g_l=g(l)$. $\mathbf{A}$ is an $N\times N$ matrix that indicates the ICI part and includes the coefficients of the OIR which have been added to the matrix $\mathbf{G}$ in order to avoid the ICI due to $\nu<L_o-1$ and $\mathbf{B}$ is also an $N\times N$ matrix that shows the ISI part due to interference of previous OFDM symbol $\mathbf{x}(k-1)$.

$$\mathbf{A}=\begin{bmatrix}
0 & \cdots & g_{L_o-1} & \cdots & \cdots & g_{\nu+1} & 0 & \cdots & 0\\
0 & \cdots & 0 & g_{L_o-1} & \cdots & g_{\nu+2} & 0 & \cdots & 0\\
\vdots & \ddots & \ddots & \ddots & \ddots & \vdots & \vdots & \vdots & \vdots\\
0 & \cdots & 0 & \cdots & \cdots & g_{L_o-1} & 0 & \cdots & 0\\
0 & \cdots & 0 & \cdots & \cdots & 0 & 0 & \cdots & 0\\
\vdots & \ddots & \ddots & \ddots & \ddots & \ddots & \ddots & \vdots & \vdots\\
0 & \cdots & 0 & \cdots & \cdots & 0 & 0 & \cdots & 0
\end{bmatrix}$$

$$\mathbf{B}=\begin{bmatrix}
0 & \cdots & 0 & g_{L_o-1} & \cdots & \cdots & g_{\nu+1}\\
0 & \cdots & 0 & 0 & g_{L_o-1} & \cdots & g_{\nu+2}\\
\vdots & \ddots & \ddots & \ddots & \ddots & \ddots & \vdots\\
0 & \cdots & 0 & \cdots & \cdots & 0 & g_{L_o-1}\\
0 & \cdots & 0 & \cdots & \cdots & 0 & 0\\
\vdots & \ddots & \ddots & \ddots & \ddots & \ddots & \vdots\\
0 & \cdots & 0 & \cdots & \cdots & 0 & 0
\end{bmatrix}$$

Assuming that $\mathbf{x}(k)$ and $\mathbf{x}(k-1)$ are independent, based on (2), the SINR becomes

[1] This is a practical assumption in the OFDM system that limits the ISI only from the previous OFDM symbol.

$$\text{SINR} = \frac{E[\mathbf{x}(k)^H \mathbf{G}^H \mathbf{G}\mathbf{x}(k)]}{E[\mathbf{x}(k)^H \mathbf{A}^H \mathbf{A}\mathbf{x}(k)] + E[\mathbf{x}(k-1)^H \mathbf{B}^H \mathbf{B}\mathbf{x}(k-1)] + E[\boldsymbol{\eta}(k)^H \boldsymbol{\eta}(k)]} \quad (3)$$

where $(.)^H$ denotes transposed complex conjugate. Please note that similar measures used earlier for TEQ design (e.g. in [3]) are not true SINR measures such as ours in (3). The signal power can be written as

$$P_S = E[\mathbf{x}(k)^H \mathbf{G}^H \mathbf{G}\mathbf{x}(k)] = \text{trace}(\mathbf{G}E[\mathbf{x}(k)\mathbf{x}(k)^H]\mathbf{G}^H)$$

Meanwhile, $\mathbf{x}(k) = \mathbf{W}^H \mathbf{q}(k)$, where $\mathbf{W}$ is the unitary FFT matrix whose element w_{lm} is

$$w_{l,m} = \frac{1}{\sqrt{N}} e^{-j\frac{2\pi}{N}lm} \qquad l = 0,\cdots,N-1 \quad m = 0,\cdots,N-1$$

Assuming that $q(k)$ is an independent and identically distributed (iid) process such that $E[\mathbf{q}(k)\mathbf{q}(k)^H] = \sigma_q^2 \mathbf{I}_N$, where $\mathbf{I}_N$ is the $N \times N$ identity matrix, P_S will be

$$\begin{aligned} P_S &= \text{trace}(\mathbf{G}E[\mathbf{W}^H \mathbf{q}(k)\mathbf{q}(k)^H \mathbf{W}]\mathbf{G}^H) = \sigma_q^2 \text{trace}(\mathbf{G}\mathbf{G}^H) \\ &= \sigma_q^2 N \sum_{l=0}^{L_o-1} |g(l)|^2 \end{aligned} \quad (4)$$

In the same way one can show that

$$P_{ICI} = \sigma_q^2 \text{trace}(\mathbf{A}\mathbf{A}^H) = \sigma_q^2 \sum_{m=\nu+1}^{L_o-1} \sum_{l=m}^{L_o-1} |g(l)|^2 \quad (5)$$

$$P_{ISI} = \sigma_q^2 \text{trace}(\mathbf{B}\mathbf{B}^H) = \sigma_q^2 \sum_{m=\nu+1}^{L_o-1} \sum_{l=m}^{L_o-1} g|(l)|^2 \quad (6)$$

Comparing (5) and (6), it can be seen that ISI power and ICI power at the output of the equalizer are equal. Defining $\mathbf{f} = [f_0, f_1, \cdots, f_{M-1}]^T$ as the equalizer impulse response vector (or equalizer tap coefficients), the $\boldsymbol{\eta}(k)$ becomes

$$\boldsymbol{\eta}(k) = \mathbf{N}(k)\mathbf{f} \quad (7)$$

where $\mathbf{N}(k)$ is an $N \times M$ matrix whose elements are the samples of the additive white Gaussian noise $n(l)$. Based on (7) the power of noise, P_{noise}, at the output of equalizer is given by

$$P_{noise} = E[\boldsymbol{\eta}(k)^H \boldsymbol{\eta}(k)] = \sigma_n^2 \text{trace}(\mathbf{f}\,\mathbf{f}^H) = \sigma_n^2 \mathbf{f}^H \mathbf{f} \quad (8)$$

Since $g(l) = h(l) * f(l)$, by defining $\mathbf{g} = [g(0), g(1), \cdots, g(L_o - 1)]^T$, it can be shown that $\mathbf{g} = \mathbf{H}\,\mathbf{f}$, where $\mathbf{H}$ is defined as

$$\mathbf{H} = \begin{bmatrix} h_0 & 0 & 0 & \cdots & 0 \\ h_1 & h_0 & 0 & \cdots & 0 \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ h_{L-1} & h_{L-2} & h_{L-3} & \cdots & h_{L-M} \\ 0 & h_{L-1} & h_{L-2} & \cdots & h_{L-M+1} \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & 0 & \cdots & h_{L-1} \end{bmatrix}$$

where $h_l = h(l)$. If we define $\mathbf{H}_m$ as an $(L_o - m) \times M$ matrix which includes the last $(L_o - m)$ rows of $\mathbf{H}$, we will have

$$\sum_{l=m}^{L_o-1} |g(l)|^2 = \mathbf{f}^H \mathbf{H}_m^H \mathbf{H}_m \mathbf{f} \quad (9)$$

By substituting (9) in (4), (5) and (6) and doing some manipulations, the SINR is given by

$$\text{SINR} = \frac{N\,\mathbf{f}^H \mathbf{H}^H \mathbf{H}\,\mathbf{f}}{\mathbf{f}^H \left(2 \sum_{m=\nu+1}^{L_o-1} \mathbf{H}_m^H \mathbf{H}_m + \frac{N\sigma_n^2}{\sigma_q^2} \mathbf{I}_M \right) \mathbf{f}} = \frac{\mathbf{f}^H \mathbf{P}\,\mathbf{f}}{\mathbf{f}^H \mathbf{Q}\,\mathbf{f}} \quad (10)$$

where $\mathbf{P} = N\mathbf{H}^H \mathbf{H}$ and $\mathbf{Q} = 2 \sum_{m=\nu+1}^{L_o-1} \mathbf{H}_m^H \mathbf{H}_m + \frac{N\sigma_n^2}{\sigma_q^2} \mathbf{I}_M$. Note that $\mathbf{H} = \mathbf{H}_0$. Equation (10) shows the SINR as a function of $\mathbf{f}$, the vector of equalizer tap coefficients. The criterion of equalizer design is to estimate $\mathbf{f}$ by maximizing the SINR. Since $\mathbf{P}$ is a Hermitian positive definite matrix, it can be performed as $P = \Gamma^H \Gamma$. By defining the vector $\mathbf{v} = \boldsymbol{\Gamma}\,\mathbf{f}$, the SINR can be represented by

$$\text{SINR} = \frac{\mathbf{v}^H \mathbf{v}}{\mathbf{v}^H \boldsymbol{\Gamma}^{-H} \mathbf{Q} \boldsymbol{\Gamma}^{-1} \mathbf{v}} = \frac{\mathbf{v}^H \mathbf{v}}{\mathbf{v}^H \mathbf{C}\,\mathbf{v}} \quad (11)$$

where $\mathbf{C} = \boldsymbol{\Gamma}^{-H} \mathbf{Q} \boldsymbol{\Gamma}^{-1}$. In order to maximize the SINR, the vector $\mathbf{v}$ should be in the direction of $\mathbf{v}_{\min}$, the eigenvector of $\mathbf{C}$ corresponding to the minimum eigenvalue ($\lambda_{\min}$) of $\mathbf{C}$. Then we have

$$\mathbf{v} = \boldsymbol{\Gamma}\,\mathbf{f} = a\,\mathbf{v}_{\min}$$

The nonzero scalar value of a is just a scaling factor and without loss of generality, we assume that $a = 1$. Thus the equalizer coefficient vector that maximizes the SINR becomes

$$\mathbf{f} = \boldsymbol{\Gamma}^{-1} \mathbf{v}_{\min} \quad (12)$$

Performance of the SMTE method is evaluated by estimating $\mathbf{f}$ based on (12) in the next section.

IV. COMPUTER SIMULATIONS AND COMPARISONS

A 16QAM-OFDM system with $N = 64$ subcarriers and a cyclic prefix interval $\nu = 16$ is considered in the simulations. A multipath channel with L taps is used based on the following impulse response model

$$h(l) = \sum_{m=0}^{L-1} \alpha_m e^{-\beta m} \delta(l - m) \quad (13)$$

where α_m is a zero mean complex, circularly symmetric, Gaussian random process such that $E[\alpha_m \alpha_j^*] = \delta(m - j)$, where $(.)^*$ denotes complex conjugate operation, and β is an exponential decay factor. In simulations, two channel types are considered; $\beta = 0$ for uniform delay spread profile and $\beta = 0.05$ for exponential delay spread profile. They are briefly called uniform channel and exponential channel, respectively. Meanwhile, the CIR interval of the uniform channel and the exponential channel are chosen as $L = 24$ and $L = 40$, respectively.

Figure 2 shows the delay spread profile of the OIR[2] when the equalizer has been designed based on the SMTE and the IRS methods for the uniform channel. As seen in the IRS method for $M = 16$, the total power of the OIR tap coefficients which are outside of the CP interval is less than that of the SMTE method, but the power of the tap coefficients is spread more in comparison with the SMTE method for $M = 16$. Also, as Fig. 2 shows, the delay spread profile of the OIR in the SMTE method is decreased by increasing M. Note that all delay spread profiles in Fig. 2 (and also in Fig. 4) have been normalized. When the noise power is zero ,the cumulative distribution functions (CDF) of the signal-to-interference ratio (SIR) of subcarriers have been shown in Fig.3 for the case of without equalization, for the IRS method

[2] The logarithmic scale has been used in order to highlight the small values.

with $M = 16$ and for the SMTE method when $M = 16, 24, 32$ and 40. As Fig. 3 shows, the SIR of some subcarriers in the IRS method are less than the SIR of subcarriers when the equalizer is not used. However, the CDF of the SIR of subcarriers in the SMTE method for $M = 16$ is higher in comparison with the IRS method and without an equalizer. These results indicate that the SMTE method achieves a better performance. Similar to the uniform channel, the simulations have been carried out for the exponential channel with a CIR interval $L = 40$. As Fig. 4 and Fig. 5 show the results are similar to those of the uniform channel and confirm the conclusions obtained from the uniform channel. As seen in Fig. 4, for $M = 16$, the CIR power shows less spread in the SMTE method in comparison with the IRS method. The CDFs of the SIR of subcarriers in Fig. 5 indicate the superiority of the SMTE method to the IRS method. These results elaborate the robustness of the SMTE method to different channel models as well.

V. Conclusions and Discussions

A novel time domain linear equalizer design for OFDM systems called SMTE has been derived in this paper based on maximizing the SINR of the signal at the output of the equalizer. We formulated the impacts of the true ISI and ICI on the OFDM systems at the output of the equalizer and highlighted the difference of ISI effect between multicarrier and single carrier systems. It has been shown that each tap of the OIR exceeding the CP interval contributes different ISI/ICI power that not only relates to the power of the tap coefficient, but depends on its time index as well. The criterion of the SMTE method minimizes the effects of the ISI and ICI, at the output of equalizer, created by the CIR whose duration is longer than the CP interval. Simulations results have been showed that the SMTE method improves the SIR distribution of subcarriers compared with the IRS method and the OFDM system without equalization.

It should be mentioned that the proposed SMTE method is optimal in the sense of maximizing the total SINR at the output of equalizer. The true SINR formulation proposed in this paper, which is being different than those defined in earlier literature, can be employed to maximize capacity, maximize throughput for a specified BER or minimize the BER for a specified throughput as well. These criteria are under investigation and results would be reported in the future.

References

[1] P. J. W. Melsa, R. C. Younce, and C. E. Rohrs, “Impulse response shortening for discrete multitone transceivers,” *IEEE Trans. Commun.*, vol. 44, pp. 1662-1672, Dec. 1996.

[2] C. Yin and G. Yue, “Optimal impulse response shortening for discrete multitone transceivers,” *IEEE Electronics Letters*, vol. 34, pp. 35-36, Jan. 1998.

[3] G. Arslan, B. L. Evans, and S. Kiaei, “Equalization for discrete multitone transceivers to maximize bit rate,” *IEEE Trans. Signal Processing*, vol. 49, pp.3123-3135, Dec. 2001.

[4] N. Al-Dhahir and J. M. Cioffi, “Efficiently computed reduced-parameter input-aided MMSE equalizers for ML detection: A unified approach,” *IEEE Trans. Information Theory*, vol. 42, pp. 903-915, May 1996.

[5] N. Al-Dhahir and J. M. Cioffi, “Optimum finite-length equalization for multicarrier transceivers,” ,” IEEE Trans. Commun., vol. 44, pp. 56-64, Jan.1996.

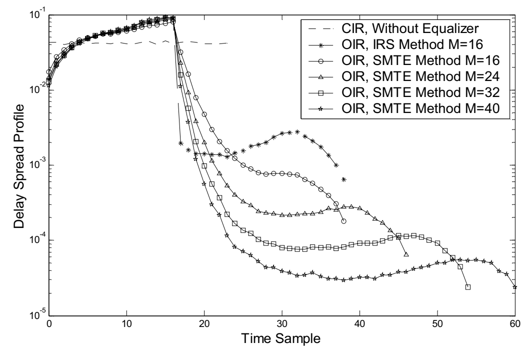

Fig. 2. Delay spread profile of the OIR for the uniform channel.

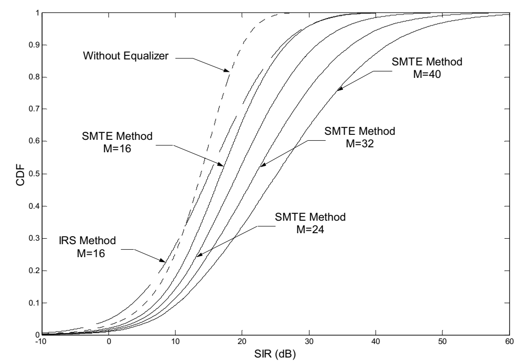

Fig. 3. The CDF of the SIR of subcarriers for the uniform channel.

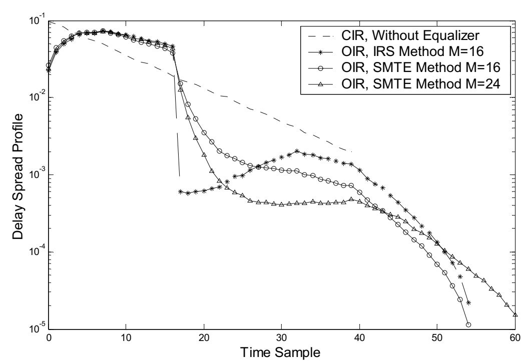

Fig. 4. Delay spread profile of the OIR for the expomential channel.

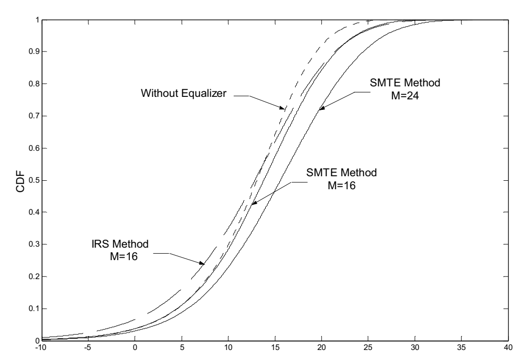

Fig. 5. The CDF of the SIR of subcarriers for the exponential channel.