# A Low-Complexity CFO Compensation Technique for Interleaved OFDMA System Uplink

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*Abstract*—**In interleaved orthogonal frequency division multiple access (OFDMA) systems uplink, multiple access interference (MAI) occurs due to different carrier frequency offsets (CFOs) from different users. In this paper, a new low complexity scheme is proposed to compensate CFO at the OFDMA receiver. The proposed scheme maximizes signal to interference ratio (SIR). Simulation results show considerable improvement in comparison to the existing methods in the literature.**

***Keywords- carrier frequency offset; interleaved subcarrier allocation; inter-carrier interference; multiple access interference; signal to interference ratio;***

## I. INTRODUCTION

Orthogonal frequency division multiple access (OFDMA) is an accepted technology for wireless communication networks, e.g., IEEE 802.16, and Long Term Evolution (LTE), due to its spectral efficiency and immunity against multipath fading. In addition, OFDMA inherits the simple frequency domain equalization and implementation from OFDM. In OFDMA subcarriers are divided into several mutually exclusive sets assigned to different users. Orthogonality among subcarriers protects system against inter carrier interference (ICI) and multiple access interference (MAI).

However, similar to OFDM, OFDMA is sensitive to carrier frequency offsets (CFOs) [1]. The Doppler effect and oscillator instabilities cause CFO that disrupts orthogonality among subcarriers. CFO introduces ICI and MAI in OFDMA systems [2]. CFO in uplink OFDMA is a more challenging problem, since it is not possible to compensate different CFOs at the base station (BS) from different users simultaneously [2]. Therefore, the performance of OFDMA system in uplink is limited by MAI.

In interleaved OFDMA, the assigned subcarriers to desired user are equally spaced over the whole transmission bandwidth, this results in largest frequency diversity [3]. Also for each subcarrier, adjacent subcarriers are occupied by different users, therefore interleaved OFDMA is more sensitive to CFO compared to other allocation schemes [3].

There are many schemes proposed to compensate the CFO effects. These schemes are classified in two categories, feedback adjustment and user detection [4]. Feedback schemes need to send feedback information that reduces data rate and spectral efficiency. In user detection techniques, synchronization errors are compensated directly in base station using advanced signal processing techniques. For example, single user detection (SUD) is a time domain technique that is proposed to compensate user's CFO [5]. The computational complexity of SUD increases by increasing the number of users. To reduce the complexity, its modified version in frequency domain has been proposed by Choi-Lee-Jung-Lee's (CLJL) [6]. As residual MAI limits performance of SUD and CLJL, some Interference cancelation algorithms have been proposed to reduce residual MAI. These algorithms are inherently iterative and improve performance at the cost of increased complexity. Briefly, conventional methods maximize desired signal power and mitigate ICI, and then use high-complexity interference cancelation algorithms to reduce MAI.

In this paper a new low-complexity interference cancellation method has been proposed to maximize average signal to interference power ratio (SIR). Conventional schemes use estimated CFO value to compensate CFO for each user directly. Unlike other methods, our proposed scheme uses an optimum correction value to compensate for CFO. It is proved that this optimum value maximizes SIR for each user. For calculating the optimum value, an iterative scheme has been proposed in [7], and attempts to maximize SIR using steepest decent (SD) method which results in more implementation complexity. In our approach there is no additional complexity compare to SUD, since the correction value is obtained directly. It has been shown that optimum correction value for a certain user is approximately the average of CFOs of two nearest adjacent users. Simulation results show improvement over conventional schemes with comparable complexity.

The rest of this paper is organized as follows. In Section II the uplink interleaved OFDMA system in the presence of CFO is modeled. The proposed scheme is presented in Section III. Section IV provides simulation results and performance comparison. Finally this paper is concluded in Section V.

## II. SYSTEM MODEL

Consider an uplink OFDMA system with $M$ active users. Each user communicates with BS through an independent multipath channel (Fig. 1). There are $N$ subcarriers on each OFDMA symbol assigned to active users with interleaved

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allocation. The information symbol of the $m$th user on the $k$th subcarrier is denoted by $X^{(m)}(k)$ that $k \in \mathcal{I}_m$ and $\mathcal{I}_m$ is the set of subcarrier indices that are assigned to the $m$th user, where $\mathcal{I}_n \cap \mathcal{I}_m = \emptyset$, for $m \neq n$ and $\bigcup_{m=1}^{M} \mathcal{I}_m = \{1,2,\dots,N\}$. The length of guard interval is $N_g$. After IDFT processing and adding guard period, the time domain transmitted signal of the $m$th user is

$$x^{(m)}(n) = \sum_{k \in \mathcal{I}_m} X^{(m)}(k)\, e^{\frac{j2\pi kn}{N}}, -Ng < n < N-1. \qquad (1)$$

The received signal at the BS from the $m$th user is

$$y^{(m)}(n) = x^{(m)}(n) * h^{(m)}(n) \qquad (2)$$

where $h^{(m)}(n)$ is channel impulse response (CIR) between the $m$th user and BS. For channel, $h^{(m)}(n), m = 0,\dots,M-1$ are statistically independent and i.i.d. complex Gaussian random variables with zero mean and $E\left\{\left|h^{(m)}(n)\right|^2\right\} = \beta_m e^{-\frac{n}{L}}$ for $n=0,1,\dots,L-1$, where $L$ is maximum delay spread and $\beta_m$ is scaling factor for the average energy of CIR, as $\sum_{n=0}^{L-1} E\left\{\left|h^{(m)}(n)\right|^2\right\} = 1$.

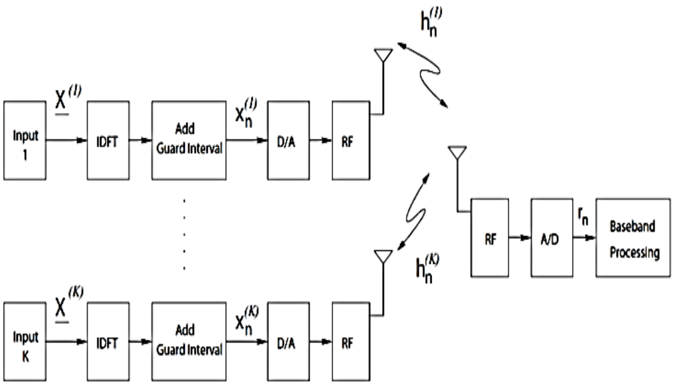


Figure 1: Uplink OFDMA system

The received signal at BS after coarse frequency synchronization is

$$r(n) = \sum_{i=1}^{M} y^{(i)}(n).\, e^{\frac{j2\pi n\epsilon_i}{N}} + w(n), -N_g < n < N-1 \qquad (3)$$

where $w(n)$ is AWGN with zero mean and variance $\sigma^2$, $\epsilon_m$ is the CFO between the $m$th user and BS normalized to subcarrier spacing ($\Delta f$). At the receiver, we have an estimation of $\epsilon_m$ denoted by $\hat{\epsilon}_m$. SUD is a direct method to detect the $m$th user's signal by multiplying $r(n)$ to $e^{-\frac{j2\pi n\hat{\epsilon}_m}{N}}$. If the received signal is compensated by $\epsilon$, the compensated signal after DFT is

$$R(k) = DFT\{r(n)\} = \sum_{i=1}^{M} \sum_{u \in \mathcal{I}_i} X^{(i)}(u) H^{(i)}(u)\, D(u,k,\epsilon_i - \epsilon) + W(k), \qquad (4)$$

where $X^{(m)}(u)$, $H^{(m)}(u)$ and $W(k)$ are information symbol, DFT of CIR of the $m$th user and DFT of noise respectively. The leakage power of the $u$th user on the $k$th subcarrier is $D(k,u,\epsilon_i - \epsilon)$. It can be shown that [1] $D(k,u,\epsilon_i - \epsilon) = f_N(u - k + \epsilon_i - \epsilon)$ and $f_N(x)$ is given by

$$f_N(x) = \frac{\sin\pi x}{N\sin\left(\frac{\pi x}{N}\right)} .\, e^{j\pi x\left(1-\frac{1}{N}\right)}. \qquad (5)$$

If $k \in \mathcal{I}_m$, we can categorize $R(k)$ in four parts as fallows

$$R(k) = \underbrace{X^{(m)}(k)H^{(m)}(k)}_{desired\ signal} D(k,k,\epsilon_m - \epsilon) + \underbrace{\sum_{\substack{u \in \mathcal{I}_m \\ u \neq k}} X^{(m)}(k)H^{(m)}(k)D(k,u,\epsilon_m - \epsilon)}_{ICI} + \underbrace{\sum_{\substack{i=1 \\ i \neq m}}^{M} \underbrace{\sum_{u \in \mathcal{I}_i} X^{(n)}(u)H^{(n)}(u)\, D(k,u,\epsilon_n - \epsilon)}_{MAI^i}}_{MAI} + W(k). \qquad (6)$$

The first part is the desired signal on the $k$th subcarrier of the $m$th user, the second is interference caused by other subcarrier of this user or ICI and the third is interference caused by other users known as MAI. The average SIR on the $k$th subcarrier, $k \in \mathcal{I}_m$, is

$$\overline{SIR}(m,k) = \frac{SP(m,k)}{IP\,(m,k) + \sum_{\substack{n=1 \\ n\neq m}}^{M} MP\,(n,m,k)} \qquad (7)$$

where $SP(m,k)$ is the desired signal power on the $k$th subcarrier. Also $IP\,(m,k)$ and $MP\,(n,m,k)$ are ICI and MAI powers respectively. In the interleaved OFDMA system uplink with perfect power control, we have [1].

$$SP = |f_N(\epsilon_m - \epsilon)|^2 \qquad (8)$$

$$IP(m,k) = |f_M(\epsilon_m - \epsilon)|^2 - |f_N(\epsilon_m - \epsilon)|^2 \qquad (9)$$

$$MP(n,m,k) = |f_M(i_n - i_m + \epsilon_n - \epsilon)|^2 \qquad (10)$$

A simple mathematical description for SIR in compensated interleaved OFDMA system can be obtained by substituting (8), (9) and (10) in (7).

## III. Proposed CFO compensation technique

In the conventional SUD method the received signal from each user is compensated by $\hat{\epsilon}_m$ and if this is an accurate estimate of $\epsilon_m$ then, as it is shown in Fig. 2, the FFT points would be aligned on the peak power of each carrier spectrum. However, at this point there could be a considerable amount of MAI from neighboring carriers. Our proposed method is to find another compensation value so that the FFT points are taken where MAI is minimum.

The difference between the proposed and conventional schemes is illustrated in Fig. 2. As shown in this figure conventional schemes maximize signal power but it will suffer from MAI. The proposed scheme maximizes SIR of the $m$th user by trying to find an optimum compensation value for $\epsilon$ called $\epsilon_{m,opt}$ that would suppress MAI and improve system performance. We assume that $\epsilon_{m,opt}$ exists and SIR can be written as an analytical function of $\epsilon$ in the vicinity of $\epsilon_{m,opt}$. With this assumption we can write

$$\frac{\partial \overline{SIR}(m,k)}{\partial \epsilon}\bigg|_{\epsilon\,=\,\epsilon_{m,opt}} = 0. \tag{11}$$

In the following, an approximation of $\epsilon_{m,opt}$ is calculated. Then an SINR analysis and complexity comparisions are made.

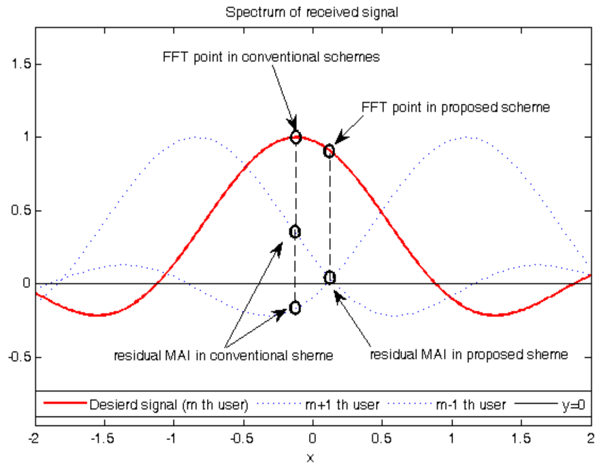


Figure 2: Proposed scheme using SIR maximization technique

### A. Calculation of aproximate value of $\epsilon_{m,opt}$

To calculate $\epsilon_{m,opt}$, (11) must be solved. Solution of (11) is very complicated. However, it should be noted that a major part of the interference is caused by two nearest adjacent subcarriers and contribution of other carriers is negligible. Also, attending to (5) and (9) can easily conclude that ICI is very small in comparison with MAI and it can be ignored, too.

If we assume that the difference between normalized CFOs of the two neighboring users are less than 0.2, it can be shown that optimum correction value for a certain user is approximately the average of CFOs of two neighbors (see Appendix A), i.e.

$$if\ |\epsilon_i - \epsilon_j| < 0.2\ \ \forall\, i,j \in \{1,\dots,M\}$$
$$\Rightarrow\quad \epsilon_{m,opt} = \frac{\epsilon_{m-1}+\epsilon_{m+1}}{2}. \tag{12}$$

It is interesting to note that $\epsilon_{m,opt}$ does not depend on $\epsilon_m$ of the desired user. Also it should be noted that BS has to use an estimation of $\epsilon_{m-1}$ and $\epsilon_{m+1}$.

In [7], in order to obtain an accurate value of $\epsilon_{m,opt}$ to maximize SIR, an iterative algorithm obtaining $\epsilon_{m,opt}$ by the steepest descent method has been employed as

$$\hat{\epsilon}_{m,opt}^{(n+1)} = \hat{\epsilon}_{m,opt}^{(n)} + \mu.\nabla\overline{SIR}(\hat{\epsilon}_{m,opt}^{(n)}) \tag{13}$$

where $\mu$ is a step size which controls the convergence rate and the steady-state error and $n$ is iteration number. Also

$$\nabla\overline{SIR}(\hat{\epsilon}_{m,opt}^{(n)}) = \lim_{\Delta_\epsilon\to 0}\frac{\overline{SIR}(\hat{\epsilon}_{m,opt}^{(n)}+\Delta_\epsilon) - \overline{SIR}(\hat{\epsilon}_{m,opt}^{(n)})}{\Delta_\epsilon} \tag{14}$$

This method has high computational complexity due to the iterative nature of steepest decent algorithm. As a further step to improve the results we propose to use the initial value obtained by (12) in (13) which would give us a better approximation of $\epsilon_{m,opt}$. As shown in the simulation results after such initialization only one step of this algorithm can be sufficient to obtain the desired accuracy. By substituting (12) to (13) and $n = 1$ iteration we have

$$\epsilon_{m,opt} = \frac{(\epsilon_{m-1}+\epsilon_{m+1})}{2} - \mu.\nabla SIR\left(\frac{(\epsilon_{m-1}+\epsilon_{m+1})}{2}\right). \tag{15}$$

According to (6), the desired signal after CFO compensation with $\epsilon_{m,opt}$ is

$$R_m(k) = X^{(m)}(k)H^{(m)}(k)f_N(\epsilon_m - \epsilon_{m,opt}),k = 1,\dots,N \tag{16}$$

where $R_m(k)$ is the $m$th user desired signal and $k$ is subcarrier index. According to (6) the power and phase of desired signal are affected by $f_N(\epsilon_m - \epsilon_{m,opt})$ and this has to be compensated, too, to restore the constellation points. To do so the signal, $R_m(k)$, must be multiplied by $\alpha$ defined as

$$\alpha = \frac{1}{f_N(\epsilon_m - \epsilon_{m,opt})} > 1. \tag{17}$$

Therefore, the required steps for the proposed scheme are listed in Tab. I

TABLE I. PROPOSED CFO COMPENSATION SCHEME

| |
|---|
| Proposed scheme is presented as follow.<br><br>To detect the $m$th user signal<br>1. Calculate $\epsilon_{m,opt}$ for each user using (15).<br>2. Compensate received signal $r(n)$ by $\epsilon_{m,opt}$<br>3. Multiplying signal with $\alpha$ given by (17) to correct phase and amplitude. |

### B. SINR analysis for proposed method

It is clear that the proposed method decreases MAI and improves SIR. Also as mentioned above, since the amplitude of the desired signal is affected by $f_N(\epsilon_m - \epsilon_{m,opt})$ this will reduce the SNR and compensation of this factor by amplifying both signal and noise with $\alpha$ will not improve SNR any more. As a result our proposed algorithm will outperform SUD as long as the the reduced SNR is less than the improvement in

MAI. Signal to interference plus noise (SINR) analysis shows these effects more clearly.

The performance of the proposed method improve over SUD when $SINR_{SUD} < SINR_p$, where p denotes the proposed method. This condition can be written as

$$(IP_p + MAI_p) - (IP_{SUD} + MAI_{SUD}) > N_p - N_{SUD}$$
$$N_{SUD} = N_0, N_p = \alpha^2 N_0. \quad (18)$$

where $N_0$ is noise power before compensation and $N_p$ is noise power after compensation by proposed method. Equation (18) leads to a threshold for signal to noise ratio (SNR), $\mathrm{SNR_t}$. can be computed as

$$SNR_t \triangleq \left| \frac{(\alpha^2-1)}{\frac{1}{SIR_p} - \frac{1}{SIR_{SUD}}} \right|. \quad (19)$$

If $SNR > SNR_t$, proposed scheme is better than SUD and vice versa.

### C. *Complexity comparision*

In this part, a comparison between computational complexity among different detectors is performed. The complexities of various detectors are listed in Tab. II. The complexities of CLJL and SUD also HLCC schemes are the same as those given in [8].

Compared to SUD scheme, the SIR maximization scheme has an additional complexity of $\frac{M(2N+1)}{4}$ per iteration [7]. The complexity of proposed scheme has an additional complexity of $N$, due to the third step in Tab I., compared to SUD.

TABLE II. COMPLEXITY COMPARISON

| Detector | Complexity (complex multiplications) |
|---|---|
| SUD [5] | $\frac{MN}{2}logN - \left[\frac{MN}{2}logM - \frac{3}{2}(M-1)N\right]$ |
| CLJL [6] | $\frac{N^2}{M} + \frac{N}{2}logN$ |
| Proposed scheme in [7] | $\frac{MN}{2}logN - \left[\frac{MN}{2}logM - \frac{3}{2}(M-1)N\right] + \frac{UM(2N+1)}{4}, U\ is\ number\ of\ iteration\ of\ SD$ |
| HLCC [4] | $\frac{N^2}{M} + \frac{N}{2}logN + (m-1)\left[N^2 + \frac{N^2}{M}\right]$ $, m\ is\ number\ of\ stage$ |
| Proposed | $\frac{MN}{2}logN - \left[\frac{MN}{2}logM - \frac{3}{2}(M-1)N\right] + N$ |

## IV. SIMULATION RESULTS

Computer simulation compares the performance of different CFO compensation techniques. Consider an interleaved OFDMA uplink system with $N = 64$ subcarriers, quadrature phase shift keying (QPSK) modulation and $M = 4$ active users that communicate with BS through a multipath Rayleigh channel. For each user, the coding scheme is a rate-1/2 convolution code with constraint length 5. Maximum delay spread is $L = 4$ and guard period is $N_g = 8 > L$. CFO values are fixed at [0.15, 0.12, 0.16, 0.08]. The proposed scheme is used in two case: 1) using (12) to calculate $\epsilon_{m,opt}$. 2) using (15).

Figure 3 illustrates SIR on each subcarrier for different compensation schemes. In case 1, SIR of proposed scheme shows considerable improvement over SUD and CLJL. Also it is close to proposed scheme in [7] , named “Maximize SIR”, in case 2.

Average SINR and SIR of subcarriers of the 4th user for different SNR are shown in Fig. 4. In this case CFO values are fixed at [0.25, -0.15, 0.2, -0.1]. As shown in this figure SIR does not change by changing SNR. But SINR increase with increasing SNR. Also the threshold value for SNR is clear in this figure. Using (19) $\mathrm{SNR_t}$ is obtained as $\mathrm{SNR_t}$=6 dB.

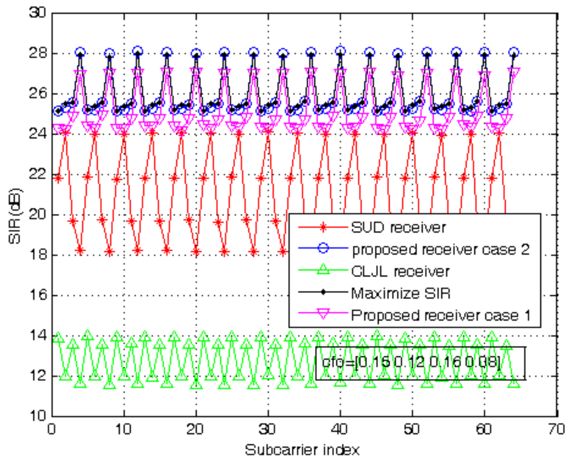


Figure 3: SIR performance, CFO=[0.15, 0.12, 0.16, 0.08]

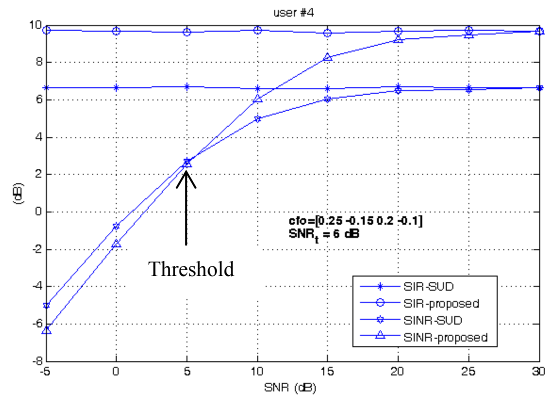


Figure 4: AWGN noise’s effect on performance of proposed scheme.

The bit error rate (BER) performance is illustrated in Fig. 5. In high SNR, this figure shows significant improvement for proposed method over SUD and CLJL. Also this shows that the performance gain cannot be achieved on a low SNR, because proposed scheme reduces SNR. In comparison with proposed scheme in [7] (maximize SIR), our proposed method has almost same performance.

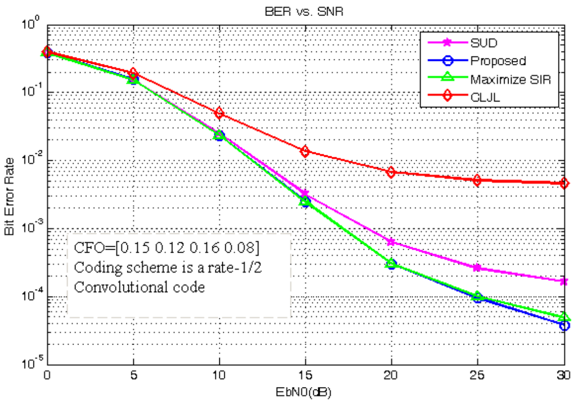


Figure 5: comparison of the BER performance

## V. CONCLUSIONS

In this paper, a new low complexity scheme was presented to compensate CFO effect for the interleaved OFDMA receiver. The proposed scheme is based on calculating optimum CFO correction value which maximizes SIR for each user. Proposed scheme imposes no additional complexity compare with SUD. Simulation results show significant improvement for proposed scheme over conventional schemes with comparable complexity e.g. SUD and CLJL.

## APPENDIX A

It should be noted that major part of interference is caused by two nearest adjacent subcarriers and others are negligible. Also, attending to (5) and (9) can easily conclude that ICI is very small compared with MAI and it can be ignored, too.

By substituting (8), (9) and (10) in (7) and ignoring ICI and far user, we have

$$\overline{SIR}(\epsilon_c) \approx$$

$$\frac{|f_N(\epsilon_m-\epsilon)|^2}{|f_M(i_{m-1}-i_m+\epsilon_{m-1}-\epsilon)|^2+|f_M(i_{m+1}-i_m+\epsilon_{m+1}-\epsilon)|^2}$$

$$=\frac{M^2}{N^2}\frac{\dfrac{\sin^2\big(\pi(\epsilon_m-\epsilon)\big)}{\sin^2\left(\frac{\pi(\epsilon_m-\epsilon)}{N}\right)}}{\dfrac{\sin^2\big(\pi(\epsilon_{m-1}-\epsilon)\big)}{\sin^2\left(\frac{\pi(-1+\epsilon_{m-1}-\epsilon)}{M}\right)}+\dfrac{\sin^2\big(\pi(\epsilon_{m+1}-\epsilon)\big)}{\sin^2\left(\frac{\pi(1+\epsilon_{m+1}-\epsilon)}{M}\right)}}. \qquad \text{(A- 1)}$$

We assume that the difference between CFO of different users is small enough

$$|\epsilon_i-\epsilon_j|=\delta<0.2 \;\; , i,j\in(1,\ldots,M) \qquad \text{(A- 2)}$$

Now we are ready to find $\epsilon_{m,opt}$. The goal is solve (11) and find $\epsilon_{m,opt}$ that

$$\begin{cases}\epsilon_{m,opt}=arg\max_{\epsilon} SIR^{(m)}(\epsilon\;\;) \\ s.t \;\; \min\{\epsilon_1,\ldots,\epsilon_M\}\le\epsilon \;\; \le\max\{\epsilon_1,\ldots,\epsilon_M\}\end{cases} \qquad \text{(A- 3)}$$

Before applying differential operator, sinusoidal phrases with small arc are replaced with first sentence of their Taylor expansion. After applying differential operator and some simplifications, the following equation is obtained as

$$\frac{\partial\overline{SIR}(m,k)}{\partial\epsilon}=0 \;\;\Rightarrow$$

$$\pi(\epsilon_m-\epsilon)\sin\big(2\pi(\epsilon_m-\epsilon)\big)\;(\pi^2(\epsilon_{m-1}-\epsilon)^2+\pi^2(\epsilon_{m+1}-\epsilon)^2)$$
$$=(1-\cos(2\pi(\epsilon\_m-\epsilon))\,)(\pi^2(\epsilon_{m-1}-\epsilon)^2+\pi^2(\epsilon_{m+1}-\epsilon)^2\;+(\epsilon_m-\epsilon)(\pi\,(\epsilon_{m-1}-\epsilon)+\pi\,(\epsilon_{m+1}-\epsilon))) \qquad \text{(A- 4)}$$

To solve this equation, other approximation is used. Sinusoidal phrases replaced with two first sentences of their Taylor expansion. After simplification we have

$$\left(2\pi-\frac{4}{3}\pi^3(\epsilon_m-\epsilon)^2\right)(\pi^2(\epsilon_{m-1}-\epsilon)^2+\pi^2(\epsilon_{m+1}-\epsilon)^2)$$
$$=2\pi\,(\epsilon_{m-1}-\epsilon)^2+2\pi\,(\epsilon_{m+1}-\epsilon)^2+\;2\,(\epsilon_m-\epsilon)(\epsilon_{m-1}-2\epsilon+\epsilon_{m+1}) \qquad \text{(A- 5)}$$

Solution of (A-5) leads to very simple answer as follow

$$\epsilon\approx(\epsilon_{m-1}+\epsilon_{m+1})/2\;\pm j(\epsilon_{m-1}-\epsilon_{m+1})/2$$
$$\xRightarrow{with\ attend\ to\ ((A-2)}$$
$$\epsilon_{m,opt}\approx(\epsilon_{m-1}+\epsilon_{m+1})/2 \qquad \text{(A- 6)}$$

## REFERENCES


[1] S.K.Hashemizadeh, H.Saeedi-Sourck, M.J.Omidi, Sensitivity analysis of interleaved OFDMA system uplink to carrier frequency offset, PIMRC, 2011, pp. 1631-1635.

[2] B.M. Morelli, C.C.J. Kuo, M.N. Pun, Synchronization techniques for orthogonal frequency division multiple access (OFDMA): A tutorial reveiew, proc.of the IEEE 95 (7) (2007) 1394-1427.

[3] S. H. Song, G. L. Chen, and K. B. Letaief, Localized or Interleaved? A Tradeoff between Diversity and CFO Interference in Multipath Channels, IEEE Trans. on Wireless Commun, vol. 10, no. 9, pp. 2829-2834, 2011.

[4] D. Huang and K. B. Letaief, "An interference cancellation scheme for carrier frequency offsets correction in OFDMA systems," IEEE Trans. Commun., vol. 53, no. 7, pp. 1155-1165, jul 2005.

[5] A. Tonello and S. Pupolin, “Performance of single user detectors in multitone multiple access asynchronous communications,” in Proc. IEEE VTC, pp. 199–203,May 2002.

[6] J. Choi, C. Lee, H.W. lung, and Y.H. Lee, "Carrier frequency offset compensation for uplink of OFDM-FDMA systems," IEEE commun. Lett., vol. 4, no. 12, pp. 414-416, Dec. 2000.

[7] W. Gang, L. Xia, X. Yue, and L. Shaoqian, Multiple carrier frequency offsets compensation for the uplink of interleaved OFDMA, in Proc. WCSP,pp. 21-23 Oct. 2010.

[8] S. Manohar, V. Tikiya, D. Sreedhar, and A. Chockalingam, “Cancellation of multiuser interference due to carrier frequency offsets in uplink OFDMA,” IEEE Trans. Wireless Commun., vol. 6, no. 7, pp. 2560–2571, Jul. 2007.