A Simple Equalization Technique to Minimize ICI in OFDMA-Based Femtocell Networks

Amila P. K. Tharaperiya Gamage*, Nandana Rajatheva †

Abstract—Inter carrier interference (ICI) is deemed as a critical issue in orthogonal frequency division multiple access (OFDMA) systems. Though ICI caused by Doppler spread is negligible in femtocell systems, frequent reception of signals over misaligned subcarriers will introduce severe ICI. Successive interference cancellation and decision feedback equalizers are some of the techniques proposed to mitigate ICI. However, most of these methods are highly complex, and thus, not suitable for commercial femtocell systems. In this paper, a simple equalization technique to reduce ICI in OFDMA based femtocell networks is proposed. It has been designed particularly considering the femtocell system’s operating environment and their limited computational capacity. This technique provides excellent ICI reduction performance for smaller delay spreads of the users.

I. INTRODUCTION

Femtocell networks have attracted a significant attention due to their promised ability to meet the very high indoor capacity demands in coming few years [1], [2]. Orthogonal frequency division multiplexing (OFDM) is one of the most suitable candidates for the physical layer of these systems due its high flexibility. Femtocell networks coexist with macrocell networks utilizing the same set or a subset of the OFDM subcarriers used in the macrocell network. Femtocell access points (FAP) are connected to their operators via IP backhauls. Since achieving reliable time synchronization over IP backhauls is not trivial, especially for simple low cost FAPs, FAPs may not be synchronized with macro base station (BS) and other FAPs [1]–[3]. Furthermore, the incapability of FAPs to calibrate their clock periodically, unlike 2G/3G BSs which synchronize it to the reference signal in E1’s and T1’s, may fluctuate the clock with time making the transmission instants of the uplink users to be different. In addition to that, FAPs and femto receivers receive signals from macrocell users with a wide range of delays. Thus, these mechanisms will result in misaligned subcarriers in both uplink (UL) and downlink (DL), and hence severe ICI. However, ICI caused by Doppler spread is negligible at indoor scenarios due to very slow movements of the users [4]. Therefore, femtocell systems must employ a mechanism to combat ICI caused by misaligned subcarriers.

There are several schemes that are proposed to mitigate ICI in OFDMA based systems, such as, successive interference cancellation (SIC) techniques in [5]–[8] and decision feedback equalizers (DFE) in [9], [10]. SIC methods involve two main operations: estimate the reliability of the received symbols over each subcarrier and order them, and equalize those symbols to reduce the effect of ICI using an equalizer constructed utilizing the reliability information. Although improved SIC methods in [6]–[8] use simpler reliability estimators and low complex equalizers, these SIC based methods are still very complex. DFE technique proposed in [9] is computationally expensive as it requires to invert large matrices. Sparse linear equations and least square algorithm, and LDLH factorization are used to reduce the complexity of the method proposed in [10] to a certain extent. The semi-blind ICI equalization scheme based on the joint diagonalization of cumulant matrices proposed in [11] is also very complex as it involves a large amount of matrix manipulations as described in the algorithm. Therefore, those methods are rather complex to be employed at low cost FAPs. The authors of [12] proposed to synchronize the FAP to the first arriving signal using blind time synchronization techniques. Though this is a simple method, its ICI reduction performance is not satisfactory.

To the best of our knowledge, not many ICI reduction methods are proposed specifically for femtocell systems meeting their requirements. Thus, a novel low complexity equalizer technique to minimize ICI is proposed in this paper. The main advantages of the proposed technique are the requirement of low computational complexity and the excellent ICI reduction performance at indoor environments. This paper consists of six sections. System model is presented in section II while the intuitive ICI reduction methods and the proposed ICI reduction technique are explained in section III and IV respectively. Section V describes the simulation results and conclusions are given in section VI.

II. SYSTEM MODEL

The UL of an OFDMA based femtocell network coexisting with a macrocell network is considered. Multiple users simultaneously transmit signals to their respective BSs possibly with timing misalignments. Femto BS is assumed to be capable of estimating the channel impulse responses and the delays of the received signals over all the subcarriers. Furthermore, the channel is assumed to be constant over two consecutive
OFDM symbol periods. $M$ users are in OFDMA UL with $N$ subcarriers. The durations of CP, useful OFDM symbol and total OFDM symbol are $T_{CP}$, $T_U$ and $T_S$ respectively. The transmitted signal by the $m^{th}$ user in time domain is [13], [14]

$$S_m(t) = \sum_{n=-\infty}^{\infty} S_m^n(t-nT_S), -\infty < t < \infty,$$

(1)

$$S_m^n(t) = \sum_{k=-k_1}^{k_2} X[n,k] \exp \{j2\pi k\Delta f t\}, -T_{CP}\leq t \leq T_U$$

(2)

where $k_1$ and $k_2$ are the subcarrier indexes of the boundary of the sub-band occupied by the $m^{th}$ user, $\Delta f = 1/T_U$ is the bandwidth of a subcarrier and $X[n,k]$ is the transmitted symbol over the $k^{th}$ subcarrier during the $n^{th}$ OFDM symbol period.

The channel impulse response of time-varying multipath fading channel for the $m^{th}$ user is

$$h_m(t,\gamma) = \sum_{l=1}^{L} \alpha_{m,l}(t) \delta(\gamma - \gamma_{m,l})$$

(3)

where $\alpha_{m,l}(t)$ and $\gamma_{m,l}$ are the complex gain and the delay of the $l^{th}$ path respectively. Therefore, the OFDM signals from different users arrive at the BS with different delays. At the receiver, the sampling window is aligned with the signal with minimum delay and the relative delays (RD) of all the other arriving signals are measured with respect to that signal. Let these RDS also be included in the delays of each multipath component and call it $\tau_{m,l}$. Then, the sampled received signal from the $m^{th}$ user without noise can be expressed as,

$$r_m[i] = \sum_{n=-\infty}^{\infty} \sum_{k=k_1}^{k_2} \sum_{l=1}^{L} X[n,k] \alpha_{m,l}[i] \times \exp \left\{ \frac{j2\pi k(i-q_{m,l}+\theta_{m,l}-n(N+N_{CP}))}{N} \right\}$$

(4)

where $N_{CP}$ is the number of samples in CP, $q_{m,l} = \lfloor N\tau_{m,l}/T_U \rfloor$ and $\theta_{m,l} = q_{m,l} - \tau_{m,l}$. Therefore, the total OFDM signal received by the BS is

$$r[i] = \sum_{m=1}^{M} r_m[i] + w_i$$

(5)

where $w_i$ is the $i^{th}$ complex additive white Gaussian noise sample, i.e. $w_i \sim CN(0, N_0/2)$, and $N_0$ is the single sided power spectral density of noise at the channel. The samples of the CP removed total received signal during the $n^{th}$ OFDM symbol period can be rewritten as

$$r_n[i] = \sum_{m=1}^{M} \sum_{l=1}^{L} r_{m,n,l}[i] + w_i$$  \quad 0 \leq i \leq N - 1$$

(6)

$$r_{m,n,l}[i] = \begin{cases} \sum_{k=k_1}^{k_2} X[n-1,k] \alpha_{m,l}[i] \times \exp \left\{ \frac{j2\pi k(i-q_{m,l}+\theta_{m,l}+N_{CP})}{N} \right\}, & \text{if } i<q_{m,l} - N_{CP} \\ \sum_{k=k_1}^{k_2} X[n,k] \alpha_{m,l}[i] \exp \left\{ \frac{j2\pi k(i-q_{m,l}+\theta_{m,l})}{N} \right\}, & \text{Otherwise} \end{cases}$$

(7)

More generally, we assume that each subcarrier is occupied by a different user and denote the RD of the $l^{th}$ multipath component of the signal arriving over the $k^{th}$ subcarrier as $\tau_{k,l}$. Since the channel is assumed to be constant over two OFDM symbol periods, the fast Fourier transform (FFT) of $r_n[i]$ samples can be expressed in a compact matrix form as follows

$$R = Q_n X_n + Q_{n-1} X_{n-1} + F w$$

(8)

where $R$ is a $N \times 1$ matrix with FFT of $r_n[i]$ samples as the elements, $X_n$ and $X_{n-1}$ are $N \times 1$ matrices consisting of the QAM symbols transmitted during $n^{th}$ and $(n-1)^{th}$ OFDM symbol periods respectively, $w$ is a $N \times 1$ matrix with $w_i$ as the elements and, $Q_{n-1}$, $Q_n$ and $F$ are $N \times N$ matrices with $\chi_{p,k}$, $\psi_{p,k}$ and $f_{p,k}$ as the $(p,k)^{th}$ element of each matrix respectively. For the clarity of notations, row and column indexes are marked from $0$ to $N - 1$. Then,

$$f_{p,k} = \exp \left\{ -\frac{2\pi p k}{N} \right\}, \quad \forall p, k,$$

(9)

$$Q_{n-1} = \sum_{l=1}^{L} Q_{n-1,l} h_{n-1,l}$$

(10)

where $h_{n-1,l}$ is a $N \times N$ diagonal matrix with the complex gain of the $l^{th}$ multipath component of the channel over the $k^{th}$ subcarrier during the $(n-1)^{th}$ OFDM symbol period as the $k^{th}$ diagonal element and the $(p,k)^{th}$ element of the $N \times N$ matrix $Q_{n-1,l}$ is given by

$$\zeta_{l,p,k} = \begin{cases} 0, \forall p \text{ if } q_{k,l} \leq N_{CP} \\ \left\{ \begin{array}{l} \left(1 - \exp \left\{ \frac{2\pi (p-k) q_{k,l}}{N(N+N_{CP})} \right\} \right) \times \exp \left\{ \frac{j2\pi (p-k) q_{k,l}}{N(N+N_{CP})} \right\}, \\ \left(1 - \exp \left\{ \frac{2\pi q_{k,l}}{N} \right\} \right) \times \exp \left\{ \frac{j2\pi q_{k,l}}{N} \right\} \end{array} \right\}, \quad \text{for } p\neq k \text{ if } q_{k,l} > N_{CP} \end{cases}$$

(11)

$$Q_n = \sum_{l=1}^{L} Q_{n,l} h_{n,l}$$

(12)

where $h_{n,l}$ is also similar matrix to $h_{n-1,l}$ except that the multipath components of the channels during the $n^{th}$ OFDM symbol period are used to form it, and $Q_{n,l}$ is a $N \times N$ matrix with its $(p,k)^{th}$ element computed as

$$\xi_{l,p,k} = \begin{cases} N \exp \left\{ -\frac{2\pi p (q_{k,l} - \theta_{k,l})}{N} \right\}, \quad \text{for } p = k \text{ if } q_{k,l} \leq N_{CP} \\ \left\{ \begin{array}{l} \left(1 - \exp \left\{ \frac{2\pi (p-k) q_{k,l}}{N(N+N_{CP})} \right\} \right) \times \exp \left\{ \frac{j2\pi (p-k) q_{k,l}}{N(N+N_{CP})} \right\}, \\ \left(1 - \exp \left\{ \frac{2\pi q_{k,l}}{N} \right\} \right) \times \exp \left\{ \frac{j2\pi q_{k,l}}{N} \right\} \end{array} \right\}, \quad \text{for } p\neq k \text{ if } q_{k,l} > N_{CP} \end{cases}$$

(13)

According to (10)–(13), the received signal is ICI-free when the RDS of all the signals are less than $T_{CP}$. 

$$R = Q_n X_n + Q_{n-1} X_{n-1} + F w$$

(8)
III. INTUITIVE ICI REDUCTION TECHNIQUES

Based on (8), the QAM symbols in $X_n$ can be easily estimated by a simple DFE as follows:

$$X_n = Q_n^{-1}(R - Q_{n-1}X_{n-1})$$

(14)

where $X_{n-1}$ is a $N \times 1$ matrix which consists of the estimated QAM symbols during $(n-1)^{th}$ OFDM symbol period. Also,

$$X_n = X_n + Q_n^{-1}(X_{n-1} - X_{n-1}) + Q_n^{-1}Fw$$

(15)

Since $Q_n^{-1}F$ represents the FFT of the rows of $Q_n^{-1}$ and the noise samples can be treated as uncorrelated zero mean random variables, the average noise power at the equalizer output calculated by Parseval’s theorem is

$$\sigma_E^2 = \sigma^2N \sum_{p=0}^{N-1} \sum_{k=0}^{N-1} |\psi_{p,k}|^2 = \sigma^2 \sum_{p=0}^{N-1} \sum_{k=0}^{N-1} |\psi_{p,k}|^2$$

(16)

where $\psi_{p,k}$ is the $(p,k)^{th}$ element of $Q_n^{-1}$. Although this method is very straightforward, it significantly increases the noise power level, because in most of the cases, the determinant of $Q_n$ is very close to zero making the coefficients of $Q_n^{-1}$ to become very large values. The increments in average noise power due to this technique employed OFDM UL system with $N = 128$, $N_{CP}$ = 8 and uniformly distributed RDs within 0 and a maximum RD (MRD) for a few cases are shown in Table I. Thus, this equalizer technique reduces the signal to interference plus noise ratio (SINR) level significantly due to overshoot of the noise power level making it impractical to use in any OFDMA system.

The form in (8) suggests that Viterbi decoding can also be used to decode the ICI affected signal. However, the application of Viterbi decoding in this setup is highly complex since the number of nodes in the trellis is very high as a state is represented by the number of subcarriers and the number of constellation points used in the modulation scheme.

IV. PROPOSED EQUALIZATION TECHNIQUE

In order to avoid the noise power overshooting and higher complexity problems faced by the techniques discussed under section III, we propose the following equalization technique, termed as Equalizer-II, to reduce ICI. Let

$$Q_n = Q_{n+} + Q_I$$

(17)

where $Q_{n+}$ is a $N \times N$ matrix and its $(p,k)^{th}$ element is

$$\mu_{p,k} = \begin{cases} -\lambda\psi_{p,k} & , \text{for } k \neq p \\ \psi_{p,k} & , \text{for } k = p \end{cases}$$

(18)

and $Q_I$ also a $N \times N$ matrix with its $(p,k)^{th}$ element is

$$\nu_{p,k} = \begin{cases} (1 + \lambda)\psi_{p,k} & , \text{for } k \neq p \\ 0 & , \text{for } k = p \end{cases}$$

(19)

By substituting (17) to (8) and after some simplifications

$$X_n = Q_{n+}^{-1}R - Q_{n+}^{-1}Q_I X_n - Q_{n+}^{-1}Q_{n-1}X_{n-1} - Q_{n+}^{-1}Fw$$

(20)

The rationale behind proposing the substitution in (17) is to avoid having very large elements in the inverted matrix, i.e. $Q_{n+}^{-1}$, as it would prevent noise power overshooting. According to (20), the receiver has to perform a recursive decoding method to recover the QAM symbols and the proposed receiver structure employing Equalizer-II is shown in Fig.1.

The optimum value for $\lambda$ is determined based on the SINR of the equalized signal and it is highest when $\lambda \approx 0$. In addition to that, when $\lambda = 0$, $Q_{n+}$ is a diagonal matrix, and thus, calculating the inverse of it is simply reciprocating its diagonal elements. Thus, we choose $\lambda = 0$ for Equalizer-II.

A. Decoding Algorithm

Once the signals from different users arrive at the receiver, the receiver estimates the delays of each signal and aligns the sampling window with the signal having the least delay in order to avoid any interference from the $(n+1)^{th}$ OFDM symbol. Then the estimated channel impulse responses and the RDs are used for calculating $Q_n$ and $Q_{n-1}$ matrices based on (10)–(13). $Q_{n+}$ consists of all the diagonal elements of $Q_n$ while $Q_I$ consists of all the off-diagonal elements of $Q_n$. All the other elements of $Q_{n+}$ and $Q_I$ are zeros. Once the FFT of the sampled signal, i.e. $R$, is calculated, the recursive decoder performs $V(V \geq 2)$ number of recursions to estimate the QAM symbols in the $n^{th}$ OFDM symbol.

Recursion 1 : Decode the symbols ignoring $Q_{n+}^{-1}Q_I X_n$ term,

$$X_{n-1} = Q_{n+}^{-1}R - Q_{n+}^{-1}Q_{n-1}X_{n-1}$$

(21)

where $X_{n-1,v}$ is a $N \times 1$ matrix consisting of the QAM symbols estimated in $v^{th}$ recursion for the $(n-1)^{th}$ OFDM symbol period.

Recursion $v$ ($2 \leq v \leq V$) : Decode the symbols using the decoded QAM symbols in the previous recursion.

$$X_{n,v} = Q_{n+}^{-1}R - Q_{n+}^{-1}Q_{n-1}X_{n-1,v} - Q_{n+}^{-1}Q_I X_{n,v-1}$$

(22)

$$X_{n,v} = X_{n-1} - Q_{n+}^{-1}Q_I X_{n,v-1}$$

(23)
As (23) suggests, during each iteration, it is only required to subtract \( Q_{n+1}^{-1}Q_n\) from the estimated symbols in the first iteration. Since \( Q^{-1}_{n+1}Q_n \) is fixed for all the iterations, the complexity of Equalizer-II is not significantly increased as \( V \) increases. Furthermore, calculating the required matrix products, i.e., \( Q_{n+1}^{-1}Q_n \) and \( Q_{n+1}^{-1}Q_{n-1} \), are rather simple operations since \( Q_{n+1}^{-1} \) and \( Q_{n-1}^{-1} \) are diagonal matrices. In the recursive decoder, the symbols can be either estimated by a simple decoder which uses the Euclidean distance between the received symbols and the QAM constellation points as the metric and fed back to the next recursion, or the symbols can be directly fed back as soft values. In the latter, symbols are estimated only at the end of the last recursion. It should be noted that the choice of \( \lambda \) and the utilization of a recursive decoder with a simple symbol estimator have simplified the design of Equalizer-II.

V. SIMULATION RESULTS

OFDMA baseband system parameters of the simulation model are shown in Table II. All the simulations are carried out over a 6-tap multipath Rayleigh faded channel with the power delay profile of the ITU Pedestrian-B channel model [4]. Each UL user is allocated contiguous 8 subcarriers and the RDs of the femtocell users are uniformly distributed between 0 and MRD specified in the simulation, i.e., \( U(0,\text{MRD}) \). MRD is specified in terms of the number of samples of the signal with no over sampling, i.e., one sample in MRD equivalent to RD of 0.8\( \mu \)s. Performance of the Equalizer-II is analyzed with soft decision feedback (SDF) and hard decision feedback (HDF), and it is compared with the performance of the ICI reduction technique which synchronizes the receiver to the first arriving user signal (SFAS). Different numbers of recursions (\( V \)) are used in the recursive decoder while all the simulations are carried out allowing the errors to propagate.

According to the simulation results shown in Figures 2–4, Equalizer-II with HDF provides very good ICI reduction performance for smaller delay spreads, such as, MRD values up to 30 samples, whereas the performance of the SFAS method is not satisfactory even for smaller MRDs. As shown in Fig. 2, for MRD of 15 samples, this technique with HDF improves the bit error rate (BER) of the ICI affected signal to a level which is almost equal to the BER of the case when all the signals are synchronized and for MRD=20, it is only about 2dB less than the BER of the synchronized case at BER of \( 10^{-3} \) using only 2 recursions. As shown in Fig. 3, when \( V \) increases to 5, the performance of this method becomes better. With HDF, BER of the equalized signal is almost same as the case when all the subcarriers are synchronized for MRD of 20, and the BER loss when MRD of 30 is only about 2dB at BER of \( 10^{-3} \). Performance of the Equalizer-II with HDF is further improved as \( V \) is increased to 10.

Propagation delays within femtocell networks are very small and these RDs are dominated by the misalignments of transmission instants. However, these transmission instants could easily be confined to a range of 20 samples using initial-ranging procedures. Furthermore, the femtocell systems are more likely to operate at low signal to noise ratio (SNR) levels as the frequently encountered walls at indoor environments introduce additional signal attenuations which is in the range of 5-12dB per wall [15]. As the Equalizer-II performs even better at low SNR values, the range of MRD in which Equalizer-II performs well is adequate for femtocell systems.

Figures 2–4 show that the decoder with HDF provides superior performance compared to the decoder with SDF. The reason for the low performance with SDF is the use of simplified decision metric in the symbol estimator. The performance of SDF is increased if a maximum likelihood (ML) decoder is used. However, in order to employ a ML decoder, the diagonal elements of \( (-Q_{n+1}^{-1}Q_n)^V \) matrix should be calculated as the useful signal component at each recursion is \( (I - (-Q_{n+1}^{-1}Q_n)^V)^{\dagger}X_n \), where \( I \) is the \( N \times N \) identity
matrix. Since it is a computational burden for FAPs, the symbols are estimated based on the Euclidean distance between the equalized symbols and the original constellation points. Furthermore, though the decoder with HDF provides better performance, it slightly increases the complexity of the equalizer as it estimates the symbols in each recursion.

The BER performance of Equalizer-II with HDF at the presence of two macrocell users with larger delays is shown in Fig. 5. Macrocell users are assigned two sets of 8 contiguous subcarriers. The RDs of the femtocell users are distributed with U(0,20). In the scenario where the macrocell user’s delay distribution is U(20,30), Equalizer-II with 5 recursions has almost eliminated the effect due to ICI. When the same equalizer operates at $E_b/N_0$ of about 20dB, the BER loss is less than 3dB for macrocell user’s delay distribution of U(20,50) and the BER loss is even lower for macrocell user’s delay distribution of U(20,70) with $E_b/N_0$ about 15dB. Thus, Equalizer-II is able to provide excellent ICI reduction performance in the presence of macrocell users with large RDs at low SNR indoor environments. Thus, approximately synchronizing the Equalizer-II employed femtocell systems with the macrocell network will provide excellent BER performance.

VI. CONCLUSIONS

A novel low complexity equalizer technique to reduce ICI in OFDMA based femtocell systems is proposed in this paper. Equalizer-II with SDF does not provide sufficient performance. However, it performs well with HDF achieving its objective. Equalizer-II provides excellent ICI reduction performance for small values of MRDs. With HDF and 5 recursions, it improves the BER of an ICI affected signal which has a MRD of 20 samples to a level which is almost equivalent to the case where all the subcarriers are synchronized, and the BER loss when MRD=30 is only about 2dB at BER of $10^{-3}$ as well. Its ICI reduction performance can be further improved by increasing the number of recursions used in the recursive decoder. This technique requires much less computational power and the ability to adjust the number of recursions used in the decoder based on the RDs of the users makes it computationally very efficient. Therefore, it is an ideal candidate for the OFDMA based femtocell systems.

REFERENCES