A Scheme to Support Concurrent Transmissions in OFDMA based Ad Hoc networks

Hongyi Xiong
School of Electronic Engineering
Queen Mary University of London
London, United Kingdom
hongyi.xiong@eecs.qmul.ac.uk

Eliane Bodanese
School of Electronic Engineering
Queen Mary University of London
London, United Kingdom
eliane.bodanese@eecs.qmul.ac.uk

***Abstract*—In this paper, we propose a novel system architecture to realize OFDMA in ad hoc networks. A partial time synchronization strategy is presented based on the proposed system model. This proposed scheme can support concurrent transmission without global clock synchronization. We also propose a null subcarrier based frequency synchronization scheme to estimate and compensate frequency offsets in a multiple user environment. The simulation results show a good performance of our proposed synchronization scheme in terms of frequency offset estimation error and variance.**

***Index Terms*—Ad hoc networks, OFDMA, time and frequency synchronization.**

I. Introduction

OFDMA have been standardized in many multiuser commercial systems, such as LTE and WiMAX. Compared to existing multiband MAC protocols [5] [6], OFDMA has many distinct advantages. For example, OFDMA can support multiple concurrent transmissions and receptions in one node while Multiband MAC protocol can only allow a node to send or receive a single transmission at a time. One of the most prominent technical challenges in OFDMA design is related to its stringent requirements for time and frequency synchronization. Many corresponding synchronization algorithms have been proposed in recent years for infrastructure based wireless systems [1-4]. However, there are only few proposals for implementing OFDMA in wireless ad hoc networks and all of them do not consider the synchronization issues. This is because in OFDMA, the signals from different transmitters need to arrive at the common receiver at the same time for FFT transformation. This normally requires global time synchronization that is very difficult to realize in an ad hoc network. So far, to the authors' knowledge, the only proposal for synchronization in ad hoc networks is [5]. However, they only consider the situation when a new user enters into the network and it needs to synchronize with the leader. Moreover, the authors do not consider the situations with multiple concurrent transmissions.

In this paper, we propose a novel system architecture and a partial time synchronization strategy to support OFDMA in ad hoc networks. We also propose a null subcarrier based frequency offset estimation and compensation algorithm to enable concurrent transmissions.

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II. Proposed System Architecture and Partial Time Synchronization

We consider an N subcarriers OFDMA system with S sub-channels. We assume that a sub-channel is the minimum data transmission unit for a user to transmit data. In OFDMA based wireless systems, the transmitters must be synchronized to transmit data packets to a common receiver so that the packets can arrive at the receiver at the same time. In traditional wireless communication systems, this is normally done by the coordinator such as the base station through pilot information exchange. However, in ad hoc networks, there is no central coordinator. Each receiver is responsible for synchronizing its transmitters simultaneously, which is a very difficult task to achieve. Moreover, the signals from different transmitters have different channel impulse responses and frequency offsets. In order to recover the signal for each transmitter, a very complex signal processing is needed. Ad hoc nodes cannot perform very complex signal processing due to their hardware limitations. In this paper, a contiguous subcarrier permutation scheme is proposed, where each sub-channel contains adjacent subcarriers. In the end of each sub-channel, there are J_0 null sub-carriers, which form a spectrum gap to separate the adjacent sub-channels. Therefore, if the frequency offset of each transmitter is smaller enough compared to this spectral gap, the received signal can be successfully separated by using a set of bandpass filters. In order to successfully separate the individual signals from different transmitters, the sub-channels used by the transmitters must be different. The number of concurrent transmissions allowed in a node will be equal to the existent number of bandpass filters in that node. Later in this paper we investigate the effect of the number of bandpass filters in a node on the system throughput.

With the help of the bandpass filters, the signals from different transmitters do not need to arrive at the common receiver at the same time. To successfully receive an OFDMA packet the receiver needs to determine the correct symbol starting position, so that it can align the FFT window for each transmitter. This task can be achieved by using a training symbol with repetitive parts [3]. In this paper, a frame transmission is proposed among the ad hoc users. Each frame consists of one training block at the beginning, which is used for time synchronization and channel estimation, followed by B data blocks. Each block contains one OFDMA symbol which is composed of $N+L$ samples where L is the length of the cyclic prefix. Each transmitter will create its own training blocks using the assigned subcarriers. Without global synchronization, each node needs to continuously monitor all the sub-channels. This excessively consumes energy. We propose to use a cross layer solution by using a separate common signaling channel. The synchronization in the physical layer will be jointly coordinated by the MAC and routing layers through a cross layer signaling. The signaling common channel consists of one sub-channel which is shared among all ad hoc nodes. Two sets of transceivers are required in each node: one for the data traffic, and the other for the signaling sub-channel. The partial time synchronization scheme is described as follow:

In an ad hoc network, each node only needs to continuously monitor the signaling channel by looking for the training symbol. In the route discovery process, when a source node wants to transmit data to a destination node, each transmitter along a multi-hop path will inform its receiver its expected transmission time and its local time reference. This information is encapsulated in the routing request signaling messages. Accordingly, each node will create a time reference table which records the time reference for each one hop transmitter. Since the transmission range for each node is limited (around 100m to 250m), the time offset due to the propagation delay can be neglected. In this case, global time synchronization becomes not necessary. Each node will only need to set proper filters and look for the training block for a specific sub-channel at a proper time according to the estimated transmission time and time reference for each transmitter.

Each node can allocate one or more sub-channels according to its QoS requirements. To simplify the synchronization task the sub-channels used by one transmitter should be adjacent to each other. Specifically, in the bth frame and pth OFDMA block, the sth sub-channel is assigned the set of frequency indexes: $(sJ+l)$, $J=J_a+J_0$, where J_a is the number of subcarriers used in one sub-channel to carry useful information and l is the index of the sub-carriers in one sub-channel which is between 0 to N/S.

Denoted by $U^k(p;b;l;s^k_m)$ the transmitted symbol modulated on the lth sub-carrier of s^k_mth sub-channel within the pth block of bth frame for the kth transmitter to a common receiver, where $s^k_m \in [0, S)$. k is the index of the transmitters for the same potential receiver and m is the index of the receiver. After N IFFT mapping, and the addition of the cyclic prefix, a pth block of length $N+L$ OFDMA symbol in bth frame generated from the kth transmitter can be expressed as follow:

$$x^k(p;b;l;n) = \sum_{\{s^k_m\}} \sum_{l=0}^{J_a-1} U^k(p;b;l;s^k_m)\, e^{j\frac{2\pi}{N}(s^k_m J+l)n}, \quad n \in [-L, N-1\,] \quad (1)$$

Where $\{s^k_m\}$ is the set of sub-channels used by the kth transmitter to send data to the mth receiver. n represents the time domain of one OFDMA sample. The total length of an OFDMA symbol is $N+L$ which consists of the duration of N data samples and the duration of L cyclic prefix. In the equation (1) the complex exponentials has periodicity of order N and $x^k(p;b;l;n)$ is the output OFDMA symbol from the IFFT transformation which transfers the signal from frequency domain to time domain. The last L samples of $x^k(p;b;l;n)$ is the same as the first L ones.

The received signal in the common receiver is a combination of all the signals from different transmitters with distinct frequency offsets and channel impulse responses. The combined received signal can be expressed as:

$$y(t) = \sum_{k=0}^{K} y_k(t) = \sum_{k=0}^{K} x^k(t) \otimes C_k(t) \times e^{j2\pi f_k t} + \sum_{k=0}^{K} v_k(t)$$
$$= \sum_{k=0}^{K} \sum_{\tau=-\infty}^{+\infty} x^k(\tau)\, C_k(t-\tau) \times e^{j2\pi f_k t} + \sum_{k=0}^{K} v_k(t) \quad (2)$$

where f_k is the frequency offset for the kth transmitter and $v_k(t)$ is the white noise. $C_k(t)$ is the channel impulse response for the kth transmitter.

III. The Effect of the Various Number of Bandpass Filters in a Node on the System Throughput

As described above, the most distinguished advantages of OFDMA over other MAC layer protocols such as multi-band MAC protocol is that with OFDMA concurrent transmission can be realized across a node. However, more concurrent transmissions can be supported in a node means more bandpass filters are needed which increases the node complexity. Unlike the uplink in the infrastructure based cellular networks, in ad hoc networks, the number of active corresponding transmitters to a node is limited. The number of concurrent transmission supported can be adaptable for a node. In this paper, we investigate the system performance for different bandpass filters equipped in a node in terms of system throughput.

We have proposed a signal strength based sub-channel allocation scheme in OFDMA based mobile ad hoc networks in [6] which fully distributed the sub-channels among users in ad hoc networks to reduce the signaling overhead and co-channel interference. By using the proposed schemes, two experiments have taken place. 30 nodes are uniformly distributed in an area of 1000 by 1000 meters. The transmission range for each node is 250 meters. Packets are generated according to a Poisson distribution with a Maximum Transmission Unit (MTU) of 1024*B*. The simulation results are the average of 10 different random seeds and open space environment is assumed. All the experiments are performed using OPNET 14.5 simulator. Some important parameters which are used for all simulations are shown in Table 1.

Table 1 Parameters used in simulation

Simulation Parameters	***Values***
Transmission Power	0.005 W
Reception SIR_{min}	20 dB [2]
Packet Size	1024 B
Number of sub-carriers	64
Number of data sub-channels	7

We compare the OFDMA scheme [6] with other multiband solutions, such as single radio multi-channel MAC protocol (SM) [7] and multi-radio multi-band MAC protocol (MM) [8]. Each active source node randomly selects a destination node and starts a session at a random time from 0 to 10 seconds. Each session lasts until the end of the simulation. The simulation time is set to 2 minutes.

First of all, we define that each node can only have a maximum of two bandpass filters. In this case, each node can only support two concurrent transmissions at the same time. Figure 1 shows the simulation results in terms of system throughput. If in OFDMA based networks, a node cannot support concurrent transmissions, the system is the same as the multi-band based networks. From the simulation results it can be seen that with the help of the concurrency, the system throughput can be largely increased compared to other MAC protocols.

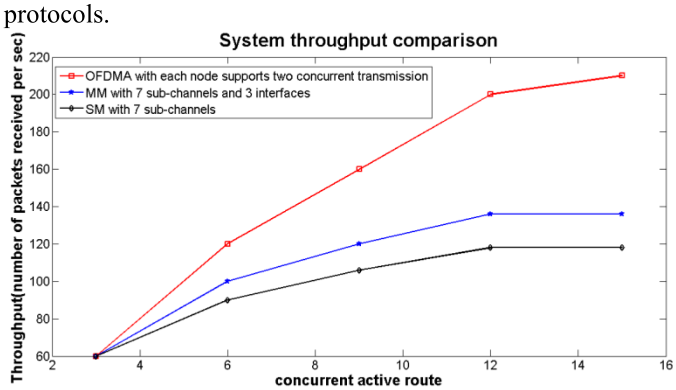

Figure 1 Comparison with two bandpass filters in each node.

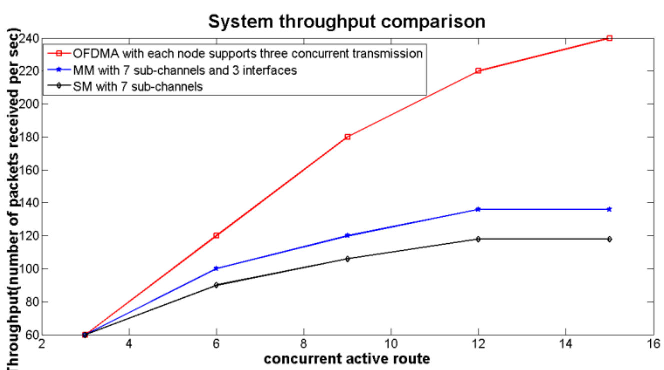

Figure 2 Comparison with three bandpass filters in each node.

Figure 2 shows the simulation results for the throughput comparison between OFDMA based system with three bandpass filters in each node and multi-band MAC systems. Compared to Figure 1, with the increased number of bandpass filters, the system throughput is increasing accordingly. If the network is not heavily loaded (the number of concurrent active routes is smaller than 15), the system throughput with three bandpass filters in each node is not significantly larger than the system with two bandpass filters in each node. Therefore, two conclusions we can be drawn from the above experiments: (1) With the help of concurrency, the performance of OFDMA system is significantly better than other MAC protocols. (2) Small number of bandpass filters is sufficient to provide substantial increase in system performance.

IV. THE NULL SUBCARRIERS BASED FREQUENCY SYNCHRONIZATION SCHEME

The received signal is passed through the bandpass filters for each transmitter and the receiver will get the time acquisition by looking for the training block. After successful time acquisition, the receiver will sample the filtered signal for FFT transfer. The sampling rate should be the same as the transmission rate of the transmitter which is set to *1/T*. The $\boldsymbol{q}$th block of $\boldsymbol{d}$th frame of the filtered received signal from the $\boldsymbol{k}$th transmitter $\boldsymbol{y_k(t)}$ collected by the receiver with sampling rate *1/T* is composed of samples taken at instant $t=(q+dB)(N+L)T+iT-\tau_k$, where τ_k is the time offset between the $\boldsymbol{k}$th transmitter and receiver. In this paper, we assume that the length of cyclic prefix is larger than the channel delay spread which means for any *t<0 or t>LT*, the channel impulse response $C_k(t)$ is equal to zero. Therefore, with the help of the training block, the time offset can be neglected and later compensated by the channel equalization [1]. Setting $r=i-n$ and removing the cyclic prefix to avoid the inter-symbol interference, the $\boldsymbol{q}$th block samples of length N in $\boldsymbol{d}$th frame from $\boldsymbol{kth}$ transmitter becomes:

$$y_k(q;d;i)=e^{j2\pi((q+dB)(N+L)T+iT)f_k} \times \sum_{\{s_m^k\}}\sum_{l=0}^{J_a-1}\widetilde{U}^k(q;d;l;s_m^k)\times e^{j\frac{2\pi}{N}(s_m^kJ+l)i} + v_k(q;d;i) \quad i\epsilon[0,\dots,N-1] \tag{3}$$

where: $\widetilde{U}^k(q;d;l;s_m^k)=U^k(q;d;l;s_m^k)\times\sum_{r=0}^{L-1}C_k(rT)\times e^{j\frac{2\pi}{N}(s_m^kJ+l)r}$ (4)

Therefore, the main task of the frequency synchronization is to find and compensate the frequency offset f_k. The proposed frequency synchronization exploits the previously investigated property of the null subcarriers inserted in each sub-channel [2] and adapts to ad hoc networks. Without frequency offset and noise, there will be no energy falling in these null sub-carriers. Therefore, we create an estimated frequency offset $\widetilde{f_k}$ and apply it to the filtered $\boldsymbol{k}$th transmitter signal by multiplying the filtered signal by the sequence$e^{-j2\pi((q+dB)(N+L)T+iT)\widetilde{f_k}}$before passing this compensated signal through FFT transfer window. By looking for the proper value of $\widetilde{f_k}(n)$ which makes the energy of the null subcarriers in the compensated signal minimum we can find the real frequency offset f_k. The compensated $\boldsymbol{k}$th transmitter signal can be expressed as:

$$\begin{aligned}\tilde{y}_k(q;d;i)&=y_k(q;d;i)\times e^{-j2\pi((q+dB)(N+L)T+iT)\widetilde{f_k}}\\&=\sum_{\{s_m^k\}}\sum_{l=0}^{J_a-1}\widetilde{U}^k(q;l;s_m^k)\times e^{j\frac{2\pi}{N}(s_m^k+l)i}\times e^{j2\pi f_k(q(N+L)T+iT)}\times e^{-j2\pi(iT+q(N+L)T)\hat{f}_k}+\tilde{v}_k(q;i)\\&=\sum_{\{s_m^k\}}\sum_{l=0}^{J_a-1}\widetilde{U}^k(q;l;s_m^k)\times e^{j\frac{2\pi}{N}(s_m^k+l)i}\times e^{j2\pi(q(N+L)T+iT)\Delta f_k}+\tilde{v}_k(q;i)\end{aligned} \tag{5}$$

After passing through the FFT window the signal becomes:

$$Y^k(q;d;l';{s_m^k}')=\frac{1}{N}\sum_{i=0}^{N-1}\tilde{y}_k(q;d;i)e^{-j\frac{2\pi}{N}({s_m^k}'J+l')i} \quad {s_m^k}'\in[0,S-1],\ l'\in[0,J-1] \tag{6}$$

Substituting (3) (5) into (6) we can get

$$Y^k(q;d;l';s_m^{k'}) = \frac{1}{N} e^{2\pi(q+dB)(N+L)T\Delta f_k} \sum_{\{s_m^k\}} \sum_{l=0}^{J_a-1} \tilde{U}^k(q;d;l;s_m^k) \times \frac{\sin(\pi\varpi N)}{\pi\varpi} e^{j\pi\varpi N} + V_{s_m^k}^k{}'(q;d;l') \quad (7)$$

where:

$$V_{s_m^k}^k{}'(q;d;l') = (\frac{1}{N}) \sum_{i=0}^{N-1} v_k(q;d;i) \times e^{-j2\pi\tilde{f}_k((q+dB)(N+L)T+iT)} \times e^{-j\frac{2\pi}{N}(s_m^{k'}J+l')i}$$

, $\varpi = [(s_m^k - s_m^{k'})J + (l - l') + TN\Delta f_k]/N$ and $\Delta f_k = f_k - \tilde{f}_k$.

In equation (7), if there is no noise and there is no frequency offset, $Y^k(q;d;l';s_m^{k'})$ is different from zero only when $s_m^k = s_m^{k'}$ and $l = l'$. For perfect synchronization and no noise, for any null sub-carriers where $U^k(q;d;l';s_m^{k'})$ is equal to 0 there is no energy detected. However, for non perfect synchronization, there is inter-symbol interference among OFDM symbols. Therefore, there will be a non null energy falling in the band of null sub-carriers. Based on this, we make use of an energy detection function ב which is the summation of the energy of the null sub-carriers for the ***k***th transmitter. The energy detection function calculates the energy in null sub-carriers for each OFDMA block and average over N_b blocks. The ב can be expressed as:

$$\beth_{N_b}^k(\Delta f_k) = \frac{1}{N_b} \sum_{q=0}^{N_b} \sum_{s_m^{k'} \in \{s_m^k\}} \sum_{l'=J_a}^{J-1} |Y^k(q;l';s_m^{k'})|^2 \quad (8)$$

Many methods can be used here to find the value of $\tilde{f}_k$ to minimize $\beth_{N_b}^k(\Delta f_k)$ such as a steepest-gradient-descent algorithm, we selected a method similar to [2]:

$$\tilde{f}_k = \underset{\Delta f_k}{argmin}\ \beth_{N_b}^k(\Delta f_k) \quad (9)$$

We average the energy detection function over successive OFDMA blocks. When the number of the OFDMA blocks is calculated long enough, the detection function becomes:

$$\beth_N^k(\Delta f_k) = \sum_{s_m^{k'} \in \{s_m^k\}} \sum_{l'=J_a}^{J-1} E\{|Y^k(q;l';s_m^{k'})|^2\}$$

$$= \sum_{s_m^{k'} \in \{s_m^k\}} \sum_{l'=J_a}^{J-1} E\{|\frac{1}{N} e^{2\pi(q+dB)(N+L)T\Delta f_k} \sum_{\{s_m^k\}} \sum_{l=0}^{J_a-1} \tilde{U}^k(q;d;l;s_m^k) \times \frac{\sin(\pi\varpi N)}{\pi\varpi} e^{j\pi\varpi N}|^2\} + \sum_{s_m^{k'} \in \{s_m^k\}} \sum_{l'=J_a}^{J-1} E\{|V_{s_m^{k'}}^k(q;l')|^2\} \quad (10)$$

where E{∗} is the expected value operator. According to the system model described in the beginning, equal energy is allocated for each data sub-carriers. Therefore $E\{|U^k(q;l;S_m^k)|^2\}$ can be represented by the signal power spectra density which is equal to σ_u^2. Since $|e^{j\theta}|^2$ is equal to 1 for ∀θ, the equation (10) can be rewritten as:

$$\beth^k(\Delta f_k) = N\sigma_u^2 \sum_{s_m^{k'} \in \{s_m^k\}} \sum_{l'=J_a}^{J-1} \sum_{\{s_m^k\}} \sum_{l=0}^{J_a-1} |C_k(S_m^k;l)|^2 \times sinc^2(\pi\varpi N) + \sigma_V^2 \frac{J_0}{J}$$

$$= N\sigma_u^2 \sum_{s_m^{k'} \in \{s_m^k\}} \sum_{l'=J_a}^{J-1} \sum_{\{s_m^k\}} \sum_{l=0}^{J_a-1} |C_k(S_m^k;l)|^2 \times sinc^2\left(\pi[(s_m^k - s_m^{k'})J + (l - l') + TN\Delta f_k]\right) + \sigma_V^2 \frac{J_0}{J} \quad (11)$$

Where $C_k(S_m^k, l) = \sum_{r=0}^{L-1} C_k(rT) \times e^{-j\frac{2\pi}{N}(S_m^k J + l)r}$ and $sinc(x) := \frac{\sin(x)}{x}$. From equation (11) it can be seen that, $\beth^k(\Delta f_k)$ is composed of all positive terms and it reaches its absolute minimum value when all the sinc(∗) terms are zero. This will happen only when Δf_k is equal to zero. Moreover, the additive white noise is not going to affect the global minimum position of $\beth^k(\Delta f_k)$. It only add a pedestal to the energy detection function $\beth^k(\Delta f_k)$.

One thing has to be noted from equation (11) that $\beth^k(\Delta f_k)$ has several minimum points when Δf_k is integral multiple of the sub-carrier spacing *1/NT*. Therefore, any iterative carrier offset estimation can guarantee to find the global minimum point as long as the frequency offset is less than *1/2NT*. If the frequency offset is greater than *1/2NT*, then one has to try multiple initializations of the initial frequency offset guess differing by integer multiples of *1/NT*, and choose the frequency offset which gets the minimum value of $\beth^k(\Delta f_k)$.

The main task is to find out the value of $\tilde{f}_k$ that makes the energy detection function reaches its global minimum value. In this paper a conventional steepest-gradient-descent algorithm [2] is used to find the frequency offset. Specifically, at the nth step, the received signal is multiplied by $e^{-j2\pi(iT+q(N+L)T)\tilde{f}_k(n)}$, where $\tilde{f}_k(n)$ is the estimated frequency offset at step n. This algorithm starts with $\tilde{f}_k(0) = 0$. Each round the algorithm calculates the energy level of the detection function and its gradient $\partial\beth^k(\tilde{f}_k(n))/\partial\tilde{f}_k(n)$ averaging over a finite number of blocks N_b. If the modulus of the gradient exceeds a predefined threshold φ_{TH}, $\tilde{f}_k(n)$ is upgraded as follows:

$$\tilde{f}_k(n+1) = \tilde{f}_k(n) - \mu \frac{\partial\beth^k(\tilde{f}_k(n))}{\partial\tilde{f}_k(n)} \quad (12)$$

Otherwise the algorithm exits from the loop. The step size μ is selected as a compromise between convergence speed and tracking capability. Larger μ will converge rapidly but with large variance for the frequency offset. Small μ will increase the number of repetitional iteration searching for the f_k and therefore increasing the computational complexity. The effect of different values of μ to the estimation performance will be evaluated through the simulation in the following session. In the next section, the performance of the proposed estimation algorithm is evaluated through the simulation.

V. Performance Evaluation

We performed experiments to test the performance of the proposed frequency offset estimation algorithm. In here, a receiver is simultaneously receiving signals from two transmitters. There are 64 sub-carriers in total and the sub-carriers are divided into $S = 8$ sub-channels with $J = 8$ sub-carriers for each of them. Therefore, $N = S \times J = 64$. The system parameters are the length of the cyclic prefix $L = 2, J_a = 6,\ J_0 = 2$. The frequency offset for these two transmitters are generated as independent random variables, distributed uniformly in (*-1/2NT, 1/2NT*) based on [2]. A Rayleigh multipath fading model is used in this simulation with overall channel delay spread less than the length of cyclic prefix *LT*. A steepest-gradient-descent algorithm is used to find the frequency offsets.

Figure 3 shows the simulation results of the frequency estimation errors of these two transmitters as a function of the iteration index for a received signal with SNR=20dB. The step size is set to 0.05. From Figure 3, it can be seen that the frequency estimation errors are converged around 30^{th} step for both transmitters. The final estimated error is slightly different for these two transmitters because the initial different frequency offsets induce different interference levels for each of them. From the simulation results it can be seen that the proposed frequency offset estimation algorithm has good performance. The frequency estimation errors for both transmitters are less than 0.01 times the sub-carrier spacing *1/NT*.

Figure 4 shows the simulation results of the frequency estimation error variance for different received SNRs. From the simulation results it can be seen that the error variance largely decreases with the increase of the received SNR. This is because a higher SNR will reduce the effect of the interference between these two transmitters due to imperfect filtering.

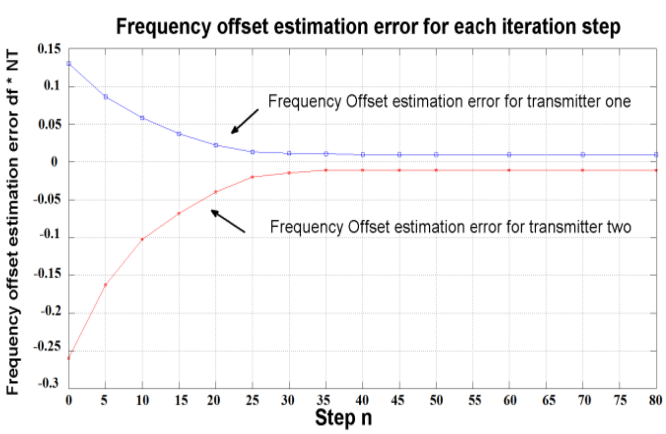

Figure 3 Frequency estimation error.

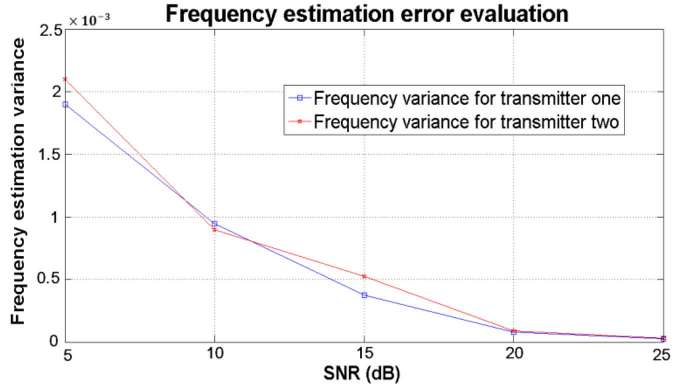

Figure 4 Frequency estimation error variance.

VI. Conclusion

We propose a novel system architecture for OFDMA based ad hoc networks. A partial time synchronization strategy is proposed to support concurrent transmissions. A null subcarrier based frequency synchronization scheme is used for the ad hoc network case. The performance of our proposed scheme is investigated through the simulations in terms of frequency offset estimation error variance.

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