Optimization of Power in Relay Path Using Sub-Carrier Resource Allocation

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Abstract- The reduction of power consumption in co-operative multiuser OFDMA systems is a great challenge in multipath environment. The power distribution among multi users in relay path has more complexity compared with direct path. To accomplish this complex problem, first we reduce Bit Error Rate (BER) using subcarrier allocation during modulation so that we can reduce the interference among multiple users. Then we can amplify the signal .After that it leads to reduce power distribution paving a good path for the improvement of resource allocation. For that purpose, we will introduce a transceiver structure to reduce the interference between adjacent subcarriers. We investigate all these things under realistic conditions. We also evaluated the performance of the signal–to-interference- noise ratio (SINR) and also bit error rate (BER) by both analysis and simulation. Finally we compare the power consumption in cooperative schemes to 16-QAM, 64-QAM, 128-QAM and 256-QAM modulation techniques.

Key words- cooperative communication, OFDMA, subcarrier resource allocation algorithm, BER, SINR, 16-QAM, 64-QAM,128-QAM,256-QAM.

1. INTRODUCTION

We always show the great concern to achieve good system throughput in wireless communication technology. The tendency of the current 3G systems is to rapidly dissipate energy in the mobile devices due to the power hungry applications which is of great concern considering the dependency of such devices on batteries for their energy. We expect 4G devices in future to be always connected supporting higher data rates and more power is required for multiple radios. We require many research issues to be addressed to advance in these directions with motivation of saving power in wireless systems. Therefore we strongly propose cooperative networks in the recent past to provide power efficient wireless systems at the expense of additional complexity and other overheads.
In this project, we present the cooperative communication techniques which are incorporated with the OFDMA systems to reduce the power distribution among multiple users. Our interest lies in the performance of cooperative OFDMA systems under subcarrier based duplexing and in particular the tradeoffs and limitations in realistic configurations. The difficulty of implementing the schemes in [8]–[12] lies in the requirement that nodes are able to transmit and receive at the same time using adjacent subcarriers, which is referred to as "subcarrier-based duplexing" in the remainder of this paper. Therefore it is important to investigate the feasibility of subcarrier based duplexing to determine the tradeoffs that may occur in the cooperative schemes in [8]–[12].
A small body of research has already proposed various sub-carrier based duplexing schemes [13]-[17].Although subcarrier based duplexing appears possible in ideal OFDM systems, the orthogonality between different subcarriers is lost in realistic communication systems due to the non-ideal characteristics of different subsystems (e.g., frequency offset of local oscillator, nonlinearity of power amplifier, etc.) and these effects need to be addressed to understand how the transmitting subcarriers will interfere with neighboring receiving subcarriers.
To perform this we make use of a transceiver structure that utilizes baseband echo cancellation to suppress the interference between the transmitting and receiving subcarriers. The performance of this transceiver is verified by analysis and computer simulation. This scheme is then incorporated into the cooperation strategy of [10] to investigate its performance under realistic conditions. It is revealed that although the performance of the cooperative network is degraded due to the residual interference imposed on the receiving subcarriers by the

transmitting subcarriers, it still performs better compared with conventional cooperation schemes.

1.1. Cooperative communication

Wireless communication which is most functional in terms of mobile access is currently a highly demanded communication technology. It has gone through several developmental phases since its inception so that it can meet to the ever changing needs of its wide range of applications [18]. The biggest challenges in the history of wireless communications which has induced considerable research for possible solutions are the multipath fading, shadowing and path loss effects of wireless channel [22]. Random variations of channel quality in time, frequency and space are caused by these effects.
The method that involves the use of a single all purpose device to deploy network services results in design complications which result in inefficient use of battery power causing short battery life [24]. Users can ease off the load on the network and in turn increase the capacity and battery life for their devices by cooperative communications in such situations.

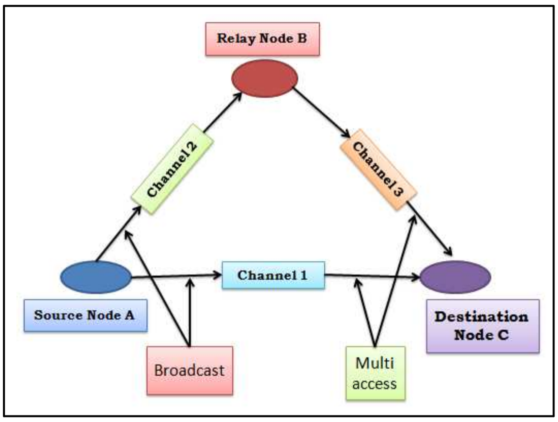

Fig.1.The relay channel

This technique which was based on the analysis of the capacity of a three- node network consisting of a source, a relay and a receiver has the assumption that all nodes operate in the same band. Therefore the system could be decomposed into a broadcast channel with respect to the source and a multipath access channel with respect to the destination. The relays whole and sole purpose is to help main channel, in the work on the relay channel but in cooperative communication, he total system resources are fixed, and users act both as information sources and as relays. In spite of indisputability of the historical importance of the first work on relay channel, recent work in cooperation has taken a somewhat different emphasis.

1.2. Cooperative communication strategies

To enable cooperation among users, different relaying techniques could be employed depending on the relative user location, channel conditions, and transceiver complexity. These are methods that define how data is processed at the relays before onward transmission to the destination. There are different types of cooperative communication strategies which would be outlined. These include the Amplify and Forward (AAF) and Decode and Forward (DAF) strategies [24].

1.2.1. Amplify-and-forward strategy (AAF)

This is a simple cooperative signaling method where each user receives a noisy version of the signal transmitted by its partner amplifies it and retransmits to the base station. The base receives two independently faded versions of the signal and combines them in order to make better decisions on information detection.

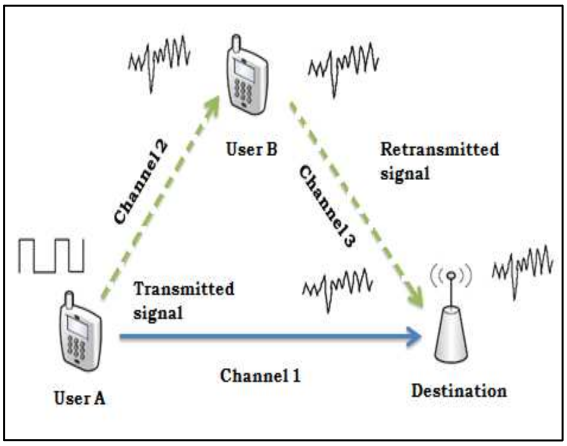

Fig.2. Amplify and forward strategy

Consider the case of a single relay. The simplest algorithm described below divides transmissions into two blocks of equal duration, one block for the source transmission and one block for the relay transmission.
For the simplest algorithm, the source transmits $X_s[k]$ for k = 1, 2 ,..., n. The relay processes its corresponding received signal $Y_r[k]$ for k = 1, 2 ,..., n, and relays the information by transmitting

$$X_r[k] = \beta_r Y_r[k-n]\,, \quad k = n+1, n+2, \ldots, 2n \tag{1}$$

To remain within its power constraint, an amplifying relay must use gain

$$\beta_r \leq \sqrt{\frac{P_s}{|A_{r,s}|^2 P_r + N_0}} \tag{2}$$

Where the gain is allowed to depend upon the fading coefficient $A_{r,s}$ between the source and relay. The destination processes its received signal $Y_d[k]$ for k =

1, 2,...,2n by some form of diversity combining of the two sub-blocks of Length n.

The main downfall of this method lies in the fact that noise contained in the signal is amplified as well and is often used when the time delay, caused by the relay to decode and encode the message has to be minimized or when there is limited computing time/power available to the relay.

1.2.2. Decode-and-forward strategy (DAF)

This strategy follows that the relay station decodes the received signal from the source node, re-encodes it and forwards it to the destination station. It is the most often preferred method to process data in the relay since there is no amplified noise in the signal sent [20]. Again, consider the case of a single relay. The simplest algorithm described below again divides transmissions into two blocks of equal duration, one block for the source transmission and one block for the relay transmission. For the simplest algorithm, the source transmits $X_s[k]$ for k = 1, 2,.. n.

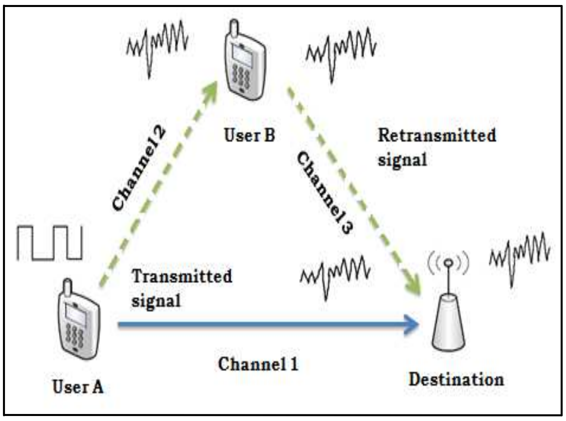

Fig.3. Decode and forward strategy

The relay forms an estimate$\hat{X}_s[k]$ by decoding its corresponding received signal $Y_r[k]$ for k = 1, 2,....,n, and relays a re-encoded version of $\hat{X}_s[k]$.For example, the relay can implement repetition coding by transmitting the signal

$$X_r[k] = \sqrt{\frac{P_r}{P_s}}\hat{X}_s[k-n]\,, \quad k = n+1, n+2, \ldots, 2n \tag{3}$$

Again, the destination processes its received signal $Y_d[k]$ for k =1, 2,..., 2n by some form of diversity combining of the two sub-blocks of length n.

Signals can be decoded by the relay completely. This takes a lot of computing time and CPU bandwidth. An error correcting code at the source makes it possible for received bit errors to be corrected at the relay station. In the absence of that, the relay can detect errors in the received signal using a checksum. Another implementation involves decoding and re-encoding the signal symbol by symbol so as to eliminate the delay caused to fully decode and process it.

2. SYSTEM MODEL

2.1. Architecture of a Transceiver

It is necessary to concentrate on the effects of the non-ideal subsystem characteristics to overcome the interference between adjacent subcarriers. To accomplish it we need to utilize special transceiver structure. So particularly in this work we introduce a transceiver structure is as shown in figure 4. Especially this transceiver is mainly based on base band echo cancellation and also used to achieve full duplex communication in wireless systems. For the sake of easy to understand, it is explained in a baseband equivalent form. The block “I/Q errors” denotes the effects of I/Q imbalance. The non- linear effects of the power amplifier and the nonlinearity introduced by other steps is denoted by “PA”.

We use the output signal of the power amplifier as a reference signal for echo cancellation in figure 4. It is obtained by attaching a coupler to the transmitter antenna and using another RF front-end to transform the acquired signal to base band samples. A finite impulse filter (FIR) is used to model the channel impulse response (CIR) between transmitter and receiver antennas. The coefficients of these antennas are determined by channel estimation using the received signal and the output signal of the power amplifier. Then we generate and subtract a replica of the received near-end signal from the received signal as shown in the figure 4. Using this approach allows the spurious signal components caused by transmit I/Q imbalance and PA nonlinearity to be suppressed.

Note that in addition to the proposed approach, nonlinear echo cancellation can also be applied in our system. In summary, the underlying principle of the proposed transceiver is to utilize digital baseband echo cancellation to suppress the nonlinear signal components generated by transmitter I/Q imbalance and PA nonlinearity. The advantage of this approach is that it can be applied to wideband systems flexibly and can explore advanced signal processing techniques to improve the performance. In addition, by using RF techniques to increase the isolation between the transmitter and receiver, the received near-end signal is attenuated such that the impact of nonlinear components resulting from quantization error and LO phase noise can also be reduced.

2.2. Sub-system imperfections

The orthogonality between subcarriers is partially lost in OFDM systems due to the non-ideal characteristics of different subsystems (e.g., nonlinearity of power amplifier, frequency offset of local oscillator etc.), resulting in signal leakage between subcarriers or inter-carrier interference (ICI). When a user is operating in subcarrier-based duplexing mode, due to the enormous difference in the power of the transmitting signal and desired signal, the effect of ICI on the receiving Sub-carriers could be significant. The subsystem imperfections that are important to consider include carrier frequency offset (CFO), time synchronization, quantization error of Analog/Digital Converter (ADC), nonlinearity of Power Amplifier (PA), I/Q imbalance and Phase Noise of Local Oscillator (LO). Their effects on the performance of subcarrier-based duplexing system are detailed as follows.

2.2.1. Carrier frequency offset

When a user is operating in subcarrier-based duplexing mode, it is natural to use the same local oscillator for the up-conversion of the transmitted signal and the down-conversion of the received signal. Therefore the received near-end signal does not experience CFO.

However, the received far-end signal suffers from CFO because the signal is transmitted from another user which uses a different oscillator, and this offset needs to be compensated in the detection of the far-end signal since OFDM systems are sensitive to frequency offset.

2.2.4. I/Q Imbalance

The effect of I/Q imbalance is modeled as

$$S' = \alpha s + \beta s^{*} \tag{5}$$

Where s represents the original signal, s' is the resulting signal and α and β are constants [21]. The term $\beta s*$ is a mirror image of the original signal and will cause ICI. Its power can be written as

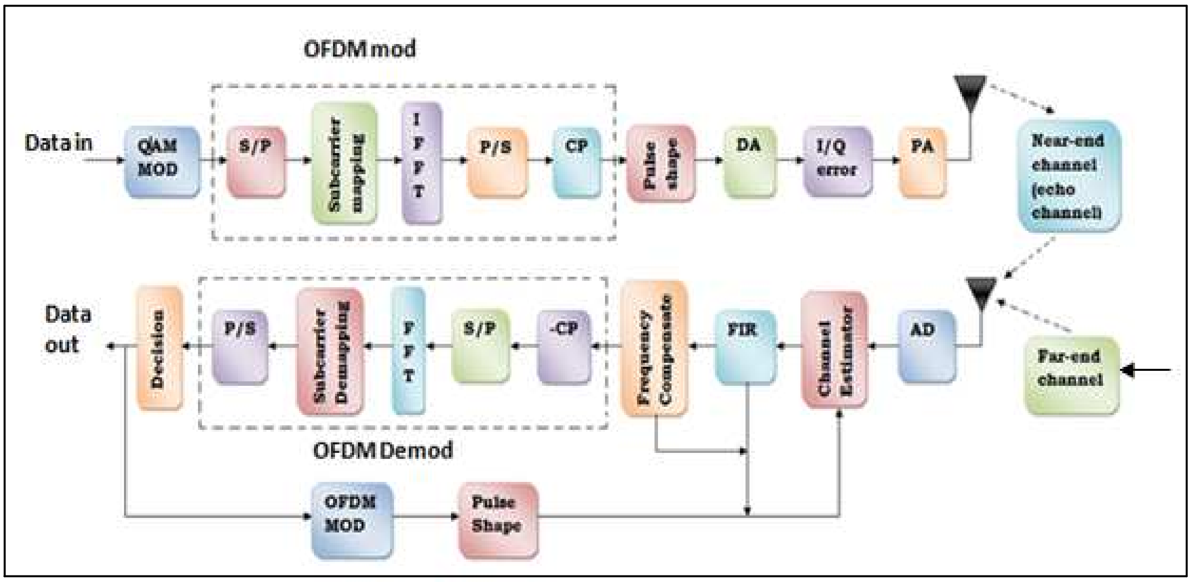

Fig.4. Transceiver structure of proposed system

2.2.2. Time synchronization

For subcarrier-based duplexing and OFDMA systems, in order to maintain orthogonality between OFDM symbols and between subcarriers, the signal received from different users should be aligned to within the cyclic prefix so that the FFT window can be re established properly. This issue can be addressed by coordinating the transmission time of each user and increasing the length of cyclic prefix.

2.2.3. Quantization error

In subcarrier-based duplexing systems, since the received near-end signal is much stronger than the far-end signal, the ADC at the receiver side needs to have sufficiently high resolution so that the far-end signal is not overwhelmed by the quantization noise. Assuming uniform, quantization noise is being used; the power of quantization noise can be approximated by [19]

$$P_{QN} = \frac{V_S^2}{3 * 4^{N_q}} \tag{4}$$

Where N_q represents the resolution of ADC in bits V_s is the peak voltage of s, which denotes the signal that is being quantized.

$$P_{S,IQ} = \gamma_{IQ} P_S \tag{6}$$

Where P_s is the power input signal,

$$\gamma_{IQ} = \left(\frac{\beta}{\alpha}\right)^2 \approx \beta^2 \tag{7}$$

Since α is close to unity.

2.2.5. Phase noise

Phase noise occurs because oscillators cannot generate pure sinusoidal waves with impulsive spectra. The power of the nonlinear signal components induced by LO phase noise can also be written as

$$P_{S,PN} = \gamma_{PN} P_S \tag{8}$$

2.2.6. PA nonlinearity

PA linearity becomes an important issue in wireless communication system design. Due to the nonlinear effects of PA, the transmitting signal consists of some nonlinear signal components, whose power can be written as

$$P_{S,NL} = \gamma_{NL} P_S \tag{9}$$

Where γ_{NL} may be obtained from the mathematical model of PA or by comparing the input and output signals of an actual PA.

3. COOPERATIVE COMMUNICATION FOR ECHO CANCELLATION

3.1. Procedure of echo cancellation

Baseband echo cancellation is carried out on a frame basis and its procedure is described as follows. Consider a node operating in subcarrier-based duplexing mode, we use S_{Tx} and S_{Rx} to denote the set of transmitting subcarriers and receiving subcarriers, respectively (S_{Tx} and S_{Rx} are mutually exclusive). Suppose there are in total W OFDM symbols in each frame. For the w-th OFDM symbol, we use $u^W = [u_1^W, u_2^W, \ldots\ldots, u_N^W]^T$. The time domain samples $X_{DT}^w = [x_1^W, x_2^W, \ldots\ldots, x_N^W]^T$ are obtained by applying IFFT to u^w, i.e., $X_{DT}^w = F^H u^w$, where the (m, n)-th entry of F is

$$F_{[m,n]} = \frac{1}{\sqrt{N}} e^{-j\frac{2\pi}{N}(m-1)(n-1)} \quad m,n = 1,2,\ldots,N \tag{10}$$

The last M samples of X_{DT}^w are then copied to the beginning of X_{DT}^w to form the Cyclic Prefix (CP), and the OFDM modulator outputs.

$$X^w = [(X_{CP}^w)^T, (X_{DT}^w)^T]^T \tag{11}$$

Where

$$X_{CP}^w = [x_{N-M+1}^w, x_{N-M+2}^w, \ldots\ldots\ldots, x_N^w]^T$$

The whole frame of OFDM symbols can thus be written as

$$X = [(X^1)^T, (X^2)^T, \ldots\ldots, (X^W)^T]^T \tag{12}$$

After passing through a pulse shaping filter and DAC, X is converted 1into an analog baseband signal $\xi(t)$, which is subsequently up-converted and amplified into $\xi_{PA}(t)$ for transmission (Fig. 4).Similarly, we can define the w-th OFDM symbol of the far-end transmit signal as

$$Y^w = [(Y_{CP}^w)^T, (Y_{DT}^w)^T]^T \tag{13}$$

The whole frame of transmit OFDM symbols is denoted as

$$Y = [(Y^1)^T, (Y^2)^T, \ldots\ldots, (Y^W)^T]^T \tag{14}$$

The near-end signal and far-end signal will be received after each passes through a multipath channel. Using $\xi_{PA}(k)$, $\Psi_{PA}(k)$, and $\rho(k)$ to denote the baseband equivalent samples of $\xi_{PA}(k)$, $\Psi_{PA}(k)$, and the received signal ($\rho(t)$ in Fig.4) respectively, we have

$$\rho(k) \approx \sum_{l=0}^{\Lambda-1} \eta_l\, \xi_{PA}(k\text{-}l) + e^{j(\omega_{0}+\omega_k)}\sum_{l=0}^{\Lambda-1} \zeta l\, \Psi_{PA}(k\text{-}l) + \zeta(k) \tag{15}$$

Where ηl and ςl ($l = 0, 1, \ldots, \Lambda - 1$) represent the Impulse Response of the near-end channel and far-end channel respectively and Λ is the order of the channels. $\zeta(k)$ represents the thermal noise. The term $e^j(\omega_0+\omega_k)$ models the CFO between the near-end user and far-end user. Considering one frame of the received samples, eq. (15) can be written in matrix-form as follows

$$\rho \approx \boldsymbol{\Xi\eta} + e^{j\omega_0}\Phi\Psi\varsigma + \zeta \tag{16}$$

Note that the frame of the near-end signal and the frame of the far-end signal need to be synchronized for eq. (16) to be valid. The alignment of OFDM symbols is achieved by coordinating the transmission time of each user and increasing the length of the cyclic prefix. The procedure of echo cancellation is made up of 2 stages. In the first stage, the channel is estimated based on the whole frame of near-end transmit samples $\xi_{PA}(k)$ and received samples $\rho(k)$ using LS algorithm, i.e.,

$$\hat{\boldsymbol{\eta}} = (\Xi^H \Xi)^{-1} \Xi^H \rho \tag{17}$$

An estimate of the near-end signal is then generated by $\Xi\eta$ and subtracted from $\rho(k)$. In the second stage, the far-end message is detected using the output of the first stage echo cancellation, and used for generating an estimate of the far-end transmit signal $\psi(k)$. Two matrices Ψ and Φ are then constructed from $\psi(k)$ and the estimation of frequency offset, following the structure of Ψ and Φ respectively. The near-end channel can then be estimated by

$$\begin{bmatrix} \hat{\eta} \\ e^{j\omega_0}\hat{\varsigma} \end{bmatrix} = (\mathbf{A}^H\mathbf{A})^{-1}\mathbf{A}^H\rho \tag{18}$$

Where A is defined as $A = [\Xi\ \Phi\ \Psi]$. By taking the far-end signal into account, the estimation precision of the near-end channel can be improved and therefore better echo cancellation can be achieved.
After the cancellation, the far-end signal is detected again and used for a new round of near-end channel identification and echo cancellation. This procedure is

repeated until a satisfactory suppression ratio and output SINR is obtained.

3.1.1. SINR of the 1st stage of echo cancellation

In the first stage of echo cancellation, the near-end channel is estimated as follows

$$\widetilde{h_{s1}} = (\widetilde{X^H}\tilde{X})^{-1}\widetilde{X^H}\mathrm{r} = \mathrm{h} + \Delta \mathrm{h}_{s1} \quad (19)$$

Where

$\Delta \mathrm{h}_{s1} = (\widetilde{X^H}\tilde{X})^{-1}\widetilde{X^H}(-\ \overline{X_q}\mathrm{h} - \bar{X}_{PN}\mathrm{h} + \bar{Y}\mathrm{g} + \mathrm{z} + \mathrm{z}_{q,Rx} + \mathrm{z}_{PN})$

represents the channel estimation error. Therefore the resulting signal after echo cancellation is

$$S_{S1} = \mathrm{r} - \tilde{X}\widetilde{h_{s1}} = \bar{Y}\mathrm{g} - \tilde{X}\Delta \mathrm{h}_{s1}\overline{X_q}\mathrm{h} - \bar{X}_{PN}\mathrm{h} + \mathrm{z} + \mathrm{z}_{q,Rx} + \mathrm{z}_{PN} \quad (20)$$

Where

$$\tilde{X}\Delta \mathrm{h}_{s1} = X\Delta h_{s1} + \overline{X_q}\Delta h_{s1} + \bar{X}_{PN}\mathrm{h}_{s1} + \bar{X}_{PN}\Delta \mathrm{h}_{s1} + X_{NL}\Delta \mathrm{h}_{s1} + X_{IQ}\Delta \mathrm{h}_{s1} \quad (21)$$

By examining eq. (20) and eq. (21), we can see that the per-subcarrier SINR of the far-end signal is limited by the last 5 terms in eq. (20) and the last 4 terms in eq. (21).The effects of $X_q\ \Delta h_{s1}$ and $X_{PN}\Delta h_{s1}$ can be ignored since their power is much smaller than X_qh and X_{PN}h, respectively. In the following we will calculate the interference power caused by the other 7 terms on the receiving subcarriers. The term $X_{NL}\Delta h_{s1}$ in eq. (21) represents the residue of the PA nonlinearity -induced nonlinear components of near-end signal and the term $X_{IQ}\Delta h_{s1}$ in eq. (21) represents the residue of the I/Q imbalance induced components of the near-end signal. The term z_{PN} in eq. (20) denotes the nonlinear components of the near-end signal resulting from LO phase noise. Therefore the per-subcarrier SINR after the 1st stage echo cancellation is

$$\mathrm{SINR}_{SC,S1} = \frac{P_{far,sc}}{\sigma^2_{Z,sc} + P_{PN,SC} + P_{QN,SC} + P_{\Delta,S1,NL,SC} + P_{\Delta,S1,IQ,SC}} \quad (22)$$

3.1.2. SINR of the 2nd stage of echo cancellation

Since the far-end signal is taken into consideration in the 2nd stage of echo cancellation, the estimation precision of the near-end channel is improved. By assuming that the far-end message is detected correctly after sufficient iterations and neglecting the estimation error of the far-end channel, the estimation error of the near-end channel is

$$\Delta \mathrm{h}_{s2} = (\widetilde{X^H}\tilde{X})^{-1}\ \widetilde{X^H}\ (-\ \overline{X_q}\mathrm{h} - \bar{X}_{PN}\mathrm{h} + \overline{Y_{NL}}\mathrm{g} + \overline{Y_{IQ}}\mathrm{g} + \mathrm{z} + \mathrm{z}_{q,Rx} + \mathrm{z}_{PN}) \quad (23)$$

In the following, we will show that the terms X_qh, X_{PN}h, z, $z_{q,Rx}$ and z_{PN} in (5.16) can be neglected. Thus the frequency-domain SINR in the 2nd stage echo cancellation is written as

$$\mathrm{SINR}_{SC,S2} = \frac{P_{far,sc}}{\sigma^2_{Z,sc} + P_{PN,SC} + P_{QN,SC} + P_{\Delta,S2,NL,SC} + P_{\Delta,S2,IQ,SC}} \quad (24)$$

By examining the denominators of (22) and (24), one may notice that the per-subcarrier SINR is mainly limited by 5 factors, i.e., the thermal noise, quantization error, phase noise, PA nonlinearity and transmitter I/Q imbalance. Compared with quantization error and LO phase noise, the output SINR is less sensitive to PA nonlinearity and I/Q imbalance since the signal components resulting from these effects are partially cancelled at the receiver side. Moreover, there are also some conventional tradeoffs and an important one is the choice of *W* and *L*.

4. SIMULATION RESULTS

In this section, we incorporate the subcarrier-based duplexing scheme investigated in the previous section into the cooperation strategy of [10] to investigate its performance. In particular we consider an OFDMA system with 20MHz bandwidth and 512 subcarriers, i.e., $N = 512$. The CP length is chosen to be 64.

Table.1.Comparison values of result analysis

QAM modulation	Non-cooperative	DF(real)	DF(ideal)
16-QAM	0.0196	0.0119	0.0117
64-QAM	0.0206	0.0129	0.0126
128-QAM	0.0228	0.0821	0.0215
256-QAM	0.0288	0.0923	0.0316

The pulse shaping filter is a Hanning windowed sinc function. A raised cosine filter with roll off factor of 0.2 is used for pulse shaping. For simplicity no convolution coding or data scrambling is applied. Both the near-end and far-end channels are modeled as multipath Rayleigh fading, with the path loss model and power delay profile (PDP). The isolation between the transmitter and receiver is assumed to be 40dB, which complements the baseband cancellation and can be achieved by the RF techniques as has been discussed. The carrier frequency offset between the near-end and far-end transmitter is set to 10 KHz.
The I/Q imbalance introduced in the up-conversion is assumed to exhibit a fixed amplitude mismatch of 1dB and fixed phase mismatch of 10 degrees between I branch and Q branch. A widely accepted solid-state power amplifier (SSPA) model [27] is used to model the effect of PA nonlinearity, with an OBO value of 10dB. For each channel realization, we firstly assume perfect isolation between different subcarriers and obtain the optimal power and subcarrier allocation strategy. The power and subcarrier assignment is then incorporated into the simulation of the previous

section to obtain the SINR values of the data streams from the source nodes to the relay nodes in the realistic model.

Since these SINR values are lower than the ideal SNR values, we scale up the transmit power of the source nodes on the corresponding subcarriers by a factor of SNR/SINR to compensate for this loss. In this way, we can obtain the overall power consumption of cooperative OFDMA systems under realistic conditions. We firstly consider a two user case where the base station and two users are placed along a straight line. The above graph shows the performance of cooperative communication in the case of using 16-QAM modulation. Fig. 5 shows the results when the data rate of user 1 and user 2 varies while *d*10 is fixed to be 50m.

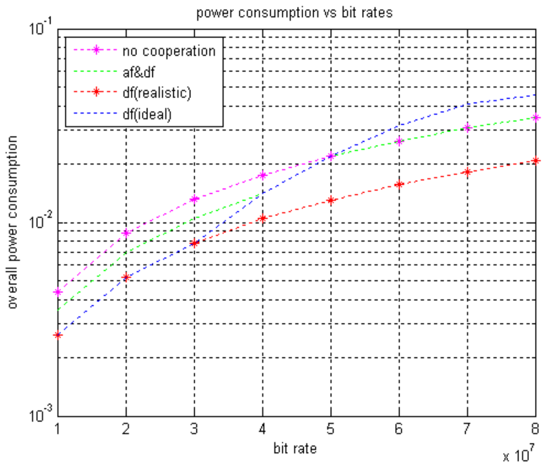

Fig.5. power consumption analysis in terms of 16-QAM

It can be seen that the difference between the power consumption of DF cooperation in the ideal case and that in the realistic case increases with the data rate. The reason is that the transmit power of user 1 increases with the data rate, therefore more interference is generated to the data streams from user 2 to user 1, and user 2 needs to scale up its transmit power in order to compensate for the SINR loss. As the data rate increases, extra transmit power required by user 2 also increases, thus the total power consumption of optimal DF cooperation will finally exceed that of AF&DF cooperation and that of no cooperation. However, it can be seen from Fig.5 that optimal DF cooperation is advantageous in most of the data rate region.

So, to overcome this problem (i.e. the increasing of extra transmit power required by the user 2) we use the next levels of QAM modulations (64,128 and 256). The results which are obtained by using different flavours of QAM modulations is as shown in below. From the below results we achieve a high data rates and gradual reduction in power consumption which is caused by the extra usage of the secondary users (user 2).

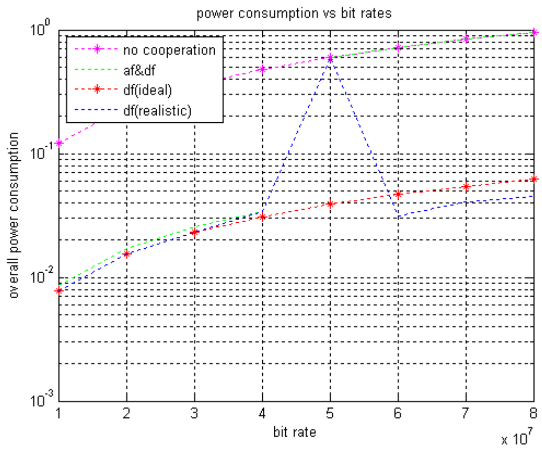

Fig.6. power consumption analysis in terms of 64-QAM

By using fig.6, fig.7 and fig.8 we have observed that the transmission bit rate is increased and power consumption reduced by increasing the levels of QAM modulation techniques as 64-QAM, 128-QAM and 256- QAM modulations. In each level of QAM modulation we have achieved the decrement in power consumption for the DF cooperation under realistic conditions compared with the non cooperative condition.

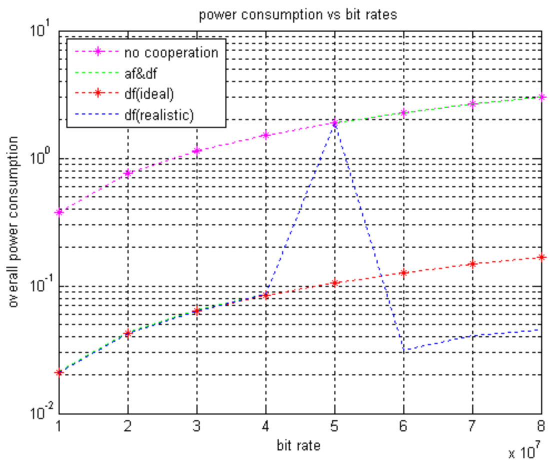

Fig.7. power consumption analysis in terms of 128-QAM

By using above comparison we investigated the sub carrier based duplexing in terms of different QAM modulation techniques performance of cooperative communication OFDMA systems with. The BER can play an important role to decrease the power consumption. The relation between the BER and power consumption is that the interference among multi users can increase the BER and decrease the data rates then automatically the usage of power can increase to transmit the messages. So, we need to reduce the bit error rate in the system. By increasing the bit rates, we can reduce the bit error rate then we can overcome this problem.

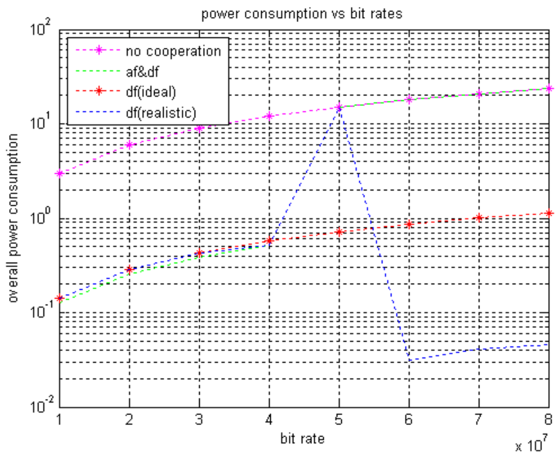

Fig.8. power consumption analysis in terms of 256-QAM

To increase the bit rate and decrease the bit error rate, we need to follow better modulation techniques other than conventional modulation techniques.

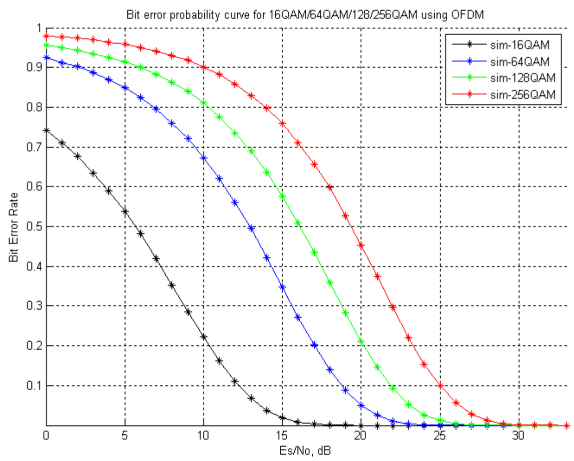

Fig.9.Bit error probability for 16/64/128/256-QAM

In our work, we use the different flavours of QAM modulations (16-QAM, 64-QAM, 128-QAM and 256-QAM). By increasing these levels, we can get high bit rates and fast transmission of bits. Therefore the bit error rate can automatically decrease. From the above graph, we have proven that the reducing possibility of bit error probability for 16/64/128/256-QAM. The approximate percentages of bit error rate are, 16-QAM is nearly 72%, 64-QAM is nearly 96%, 128-QAM is nearly 98%, 256-QAM is nearly 99.8%. So by using these QAM techniques we can reduce bit error probability efficiently compared with previous modulation techniques.

5. CONCLUSION AND FUTURE SCOPE

Cooperative communication in OFDMA systems has been shown to significantly improve wireless system performance. In this project a particular subcarrier resource allocation approach investigated [8]– [12].To perform the investigation we proposed a transceiver structure so that the system tradeoffs and limitations of this approach could be understood. The performance of the transceiver was evaluated by both analysis and computer simulation and it was shown that the non-ideal characteristics of subsystems will limit the achievable SINR. From this observation, we obtained a good improvement in reduction of power consumption particularly in relay network. To achieve this, we use different levels of QAM modulations (16-QAM, 64-QAM, 128-QAM and 256). By using these levels of QAM, we improved the data rates and decrease the bit error rates. Finally, we compared all these results by computer simulations.

Furthermore, to model the nonlinear echo cancellation, we can use low-order volterra filter in our proposed approach which involves a large number of coefficients that need to be identified. We can also use Adaptive echo cancellation instead of frame based echo cancellation which can track the changing channel response within each frame.

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