

PHY Abstraction for System Level Performance Evaluation in Inter-Carrier Interference Environment

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Abstract— There are two computer simulation methods to evaluate the performance of communication systems, i.e. the link level simulation and system level simulation. The PHY abstraction is the link-to-system mapping methodology for the system level simulation. Generally, for the system level simulation, synchronization errors such as carrier frequency offset or symbol timing offset are not taken into consideration because the accuracy of the error estimation by using the preamble or ranging is very high. However, there may be the large synchronization errors in the distributed synchronization environment such as the direct communication in IEEE 802.16n. Therefore, the synchronization errors need to be taken into consideration even in the system level simulation.

Keywords— PHY Abstraction, SLS, LLS, ICI

I. INTRODUCTION

There are two computer simulation methods, i.e. the link level simulation (LLS) and system level simulation (SLS), to evaluate the performance of communication systems. The LLS is used for the performance evaluation of the physical link according to physical layer modelling and channel behaviours between one transmitter and one receiver. On the other hand, the SLS deals with higher level issues such as interference, resource allocation and power control, regarding the entire communication system, i.e. among multiple transmitters and multiple receivers. So there exist a large number of links in the system level simulation. The receivers should decide whether the decoding of the data burst received through a physical link is success or not during the system level simulation. The PHY abstraction is the methodology which is used to predict the quality of physical link, specifically block error rate (BLER) of the data burst, according to the channel, interference, transmit power, transmit scheme and receiver structure with simple computational way[1-2].

There are some PHY abstraction methodologies for the system level performance evaluation of the mobile communication systems such as IEEE 802.16[3-4]. IEEE 802.16 is one of IMT-Advanced technologies proposed by IEEE standard board. IEEE 802.16 technology supports orthogonal frequency division multiplexing (OFDM) and various multi-input and multi-output (MIMO) transmission schemes in order to achieve higher data rate. But the PHY abstraction methods don't take synchronization error into consideration because of robustness of synchronization error estimation performance by using preamble and ranging.

Now a days, the standardization activities related to direct communications are proceeding briskly in many organizations for various application such as public protection and disaster relief (PPDR) or social network service. In IEEE, especially, a project is underway to support direct communication among two or more devices. In such a communication environment, the devices may be synchronized by distributed way without centralized arbitrator such as base station (BS)[5]. In this case, it is inescapable that interference caused by the synchronization error increases among the devices. Even though synchronization error is small, the power of interference signal may be large according to the position of the devices because there is no ranging or power control mechanism as cellular network. These effects are too critical to be ignored.

In this work, we suggest PHY abstraction method which take inter carrier interference (ICI) into consideration, caused by carrier frequency offset (CFO) between transmitters and receivers.

II. SYSTEM MODEL

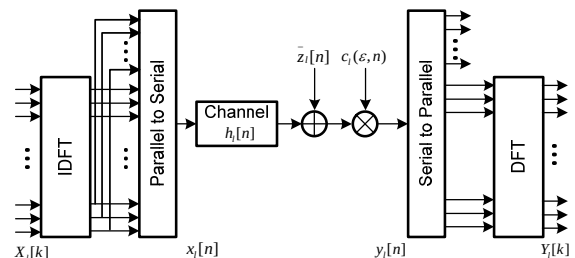


Figure 1. Discrete time baseband model of the OFDM system.

Figure 1 illustrates the baseband model of the OFDM system in the ICI environment caused by the CFO. To generate the time domain samples $x_l[n]$ of the l -th OFDM symbol, the inverse discrete Fourier transform (IDFT) is performed on the transmit symbol $X_l[k]$ with zero mean and variance σ_x^2 on the frequency domain as follows:

$$x_l[n] = \begin{cases} \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1+N_{cp}} X_l[k] e^{j2\pi k(n-N_{cp})/N}, & \text{for } 0 \leq n \leq N + N_{cp} \\ 0, & \text{otherwise} \end{cases} \quad (1)$$

where N and N_{cp} are the number of data symbols per a OFDM symbol and the number of cyclic prefix (CP) samples, respectively.

The time domain sample $x_i[n]$ passes through the channel $h_i[n]$ and the noise $\bar{z}_i[n]$ is added. If the channel $h_i[n]$ is time-invariant within each OFDM symbol and the noise $\bar{z}_i[n]$ is complex Gaussian noise with zero mean and variance σ_z^2 , the $\bar{y}_i[n]$ can be written as follows:

$$\bar{y}_i[n] = h_i[n] * x_i[n] + \bar{z}_i[n] \quad (2)$$

where $*$ means the convolution operation defined as follows:

$$h_i[n] * x_i[n] \triangleq \sum_{t=-\infty}^{+\infty} h_i[t] x_i[n-t] \quad (3)$$

We can define the signal-to-noise power ratio (SNR) as follows:

$$SNR \triangleq \frac{\sigma_x^2}{\sigma_z^2} \quad (4)$$

Also, the $\bar{y}_i[n]$ is multiplied by $c_i(\varepsilon, n)$ as shown in the figure 1 because of the CFO Δf between transmitter and receiver. The $c_i(\varepsilon, n)$ which generates phase rotation of time domain sample can be expressed as follows:

$$c_i(\varepsilon, n) = e^{j2\pi\varepsilon(1(N+N_{cp})+n-N_{cp})/N} \quad (5)$$

where ε is the carrier frequency offset normalized by subcarrier spacing f_{sub} defined as follows:

$$\varepsilon \triangleq \frac{\Delta f}{f_{sub}} \quad (6)$$

Therefore, the received time domain sample $y_i[n]$ is represented as follows:

$$y_i[n] = c_i(\varepsilon, n) \bar{y}_i[n] \\ = c_i(\varepsilon, n) (h_i[n] * x_i[n]) + z_i[n] \quad (7)$$

where

$$z_i[n] = c_i(\varepsilon, n) \bar{z}_i[n] \quad (8)$$

The received symbol $Y_i[k]$ on the frequency domain can be obtained by the discrete Fourier transform (DFT) of the received sample $y_i[n]$ after removing the CP as follows:

$$Y_i[k] = \frac{1}{\sqrt{N}} \sum_{n=N_{cp}}^{N-1+N_{cp}} y_i[n] e^{j2\pi k(n-N_{cp})/N} \quad (9)$$

If the delay spread of the channel $h_i[n]$ is smaller than CP length, there is no inter-symbol interference (ISI). Therefore $Y_i[k]$ can be represented as follows:

$$Y_i[k] = C_i(\varepsilon, k) * (H_i[k] X_i[k]) + Z_i[k] \quad (10) \\ = C_i(\varepsilon, 0) H_i[k] X_i[k] \\ + \sum_{\substack{r=-\infty \\ r \neq 0}}^{\infty} C_i(\varepsilon, r) H_i[k-r] X_i[k-r] + Z_i[k]$$

where $C_i(\varepsilon, k)$, $H_i[k]$, $Z_i[k]$ is the DFT output of $c_i(\varepsilon, n)$, $h_i[n]$ and $z_i[n]$, respectively. If the normalized carrier frequency offset is $-0.5 \leq \varepsilon \leq 0.5$, $C_i(\varepsilon, k)$ is given as follows[6]:

$$C_i(\varepsilon, k) = \sum_{n=N_{cp}}^{N-1+N_{cp}} c_i(\varepsilon, n) e^{-j2\pi(n-N_{cp})k/N} \quad (11) \\ = \left(\frac{\sin(\pi(\varepsilon-k))}{N \sin(\pi(\varepsilon-k)/N)} e^{j\pi(\varepsilon-k)(1-1/N)} \right) e^{j(2\pi\varepsilon l(1+N_{cp}/N))}$$

III. PHY ABSTRACTION METHODOLOGY

A. PHY Abstraction Methodology

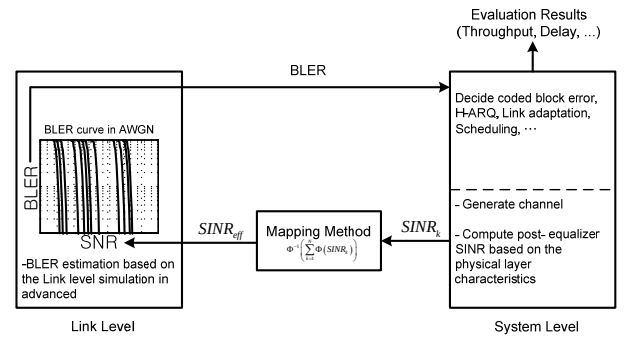


Figure 2. Link-to-System mapping procedure.

The purpose of the PHY abstraction is to estimate BLER of a data burst based on the signal-to-interference and noise power ratio (SINR), data burst size and code rate of the forward error correction (FEC) coding. Figure 2 shows PHY abstraction procedure. The post-equalizer SINR per subcarrier (or per MIMO stream) $SINR_k$ should be calculated depending on the physical layer characteristics such as channel, interference, MIMO transmit scheme and receiver structure. And then, the $SINR_k$ is mapped to the effective SINR $SINR_{eff}$ by a mapping function. The mapping function is the method to obtain the mean value of the $SINR_k$. Several effective SNR mapping (ESM) approaches to predict the BLER are proposed such as mean instantaneous capacity[7-9], exponential-effective SNR mapping (EESM)[10-13] and mutual information effective SINR mapping (MIESM) based on the received bit mutual information rate (RBIR) or based on the mean mutual information per bit (MMIB)[14-15]. In the ESM approaches, $SINR_{eff}$ is given as follows:

$$SINR_{eff} = \Phi^{-1} \left(\frac{1}{M} \sum_{i=1}^M \Phi(SINR_i) \right) \quad (12)$$

where M and $\Phi(\bullet)$ are the number of modulated symbol in a coded block and mapping function which is invertible, respectively. The $SINR_{eff}$ is used to decide the error of data burst based on the BLER curve generated in AWGN channel by the link level simulation in advance.

B. SINR Calculation in ICI environment

To calculate post-equalizer SINR in which ICI caused by CFO is reflected, we define subcarrier index $k_{l,D}$ of desired signal and subcarrier index $k_{l,I}$ of interference signal at any symbol time l . If the set of $k_{l,D}$ is an $\mathbf{K}_{l,D}$ and the set of $k_{l,I}$ is the $\mathbf{K}_{l,I}$, the equation (10) can be modified as follows:

$$Y_l[k_{l,D}] = C_{l,D}(\varepsilon_D, 0)H_{l,D}[k_{l,D}]X_{l,D}[k_{l,D}] + ICI_{self} + ICI_{adj} + Z_l[k_{l,D}] \quad (13)$$

where the ICI_{self} and ICI_{adj} are represented as follows:

$$ICI_{self} = \sum_{\substack{s \in \mathbf{K}_{l,D} \\ s \neq k_{l,D}}} C_{l,D}(\varepsilon_D, s - k_{l,D})H_{l,D}[s]X_{l,D}[s] \quad (14)$$

$$ICI_{adj} = \sum_{s \in \mathbf{K}_{l,I}} (C_{l,I}(\varepsilon_I, s - k_{l,D})H_{l,I}[s] \cdot \sqrt{P_{rx,I}/P_{rx,D}} \cdot X_{l,I}[s]) \quad (15)$$

where $C_{l,D}(\varepsilon, k)$ and $C_{l,I}(\varepsilon, k)$ are the $C_l(\varepsilon, k)$, $H_{l,D}[k]$ and $H_{l,I}[k]$ are the $H_l[k]$, $X_{l,D}[k]$ and $X_{l,I}[k]$ are the $X_l[k]$, ε_D and ε_I are the ε which are relevant to the desired and interference signal, respectively. Additionally, we can define signal to interference power ratio (SIR) as follows:

$$SIR \triangleq \frac{P_{rx,D}}{P_{rx,I}} \quad (16)$$

where $P_{rx,D}$ and $P_{rx,I}$ are the received power of the desired signal and interference signal, respectively.

For a given channel $H_{l,D}[k_{l,D}]$, $H_{l,I}[k_{l,I}]$, normalized frequency offset ε_D , ε_I , SIR , the signal-to-noise and interference power ratio $SINR_{k_{l,D}}$ at any symbol time can be obtained as follows:

$$SINR_{k_{l,D}} = \frac{E \left[\left| C_{l,D}(\varepsilon_D, 0)H_{l,D}[k_{l,D}]X_{l,D}[k_{l,D}] \right|^2 \right]}{E \left[\left| ICI_{self} + ICI_{adj} + Z_l[k_{l,D}] \right|^2 \right]} \quad (17)$$

The power of desired signal is equal to

$$E \left[\left| C_{l,D}(\varepsilon_D, 0)H_{l,D}[k_{l,D}]X_{l,D}[k_{l,D}] \right|^2 \right] = E \left[\left| C_{l,D}(\varepsilon_D, 0)H_{l,D}[k_{l,D}] \right|^2 \right] \sigma_x^2 \quad (18)$$

If $X_{l,D}[k_{l,D}]$ and $X_{l,I}[k_{l,I}]$ are uncorrelated with those in other subcarriers, ICI_{self} and ICI_{adj} are also uncorrelated with $X_{l,D}[k_{l,D}]$ and $X_{l,I}[k_{l,I}]$ because ICI_{self} and ICI_{adj} are the linear combination of $X_{l,D}[k_{l,D}]$ and $X_{l,I}[k_{l,I}]$, respectively. So the power of noise and interference is equal to

$$E \left[\left| ICI_{self} + ICI_{adj} + Z_l[k] \right|^2 \right] = E \left[\left| ICI_{self} \right|^2 \right] + E \left[\left| ICI_{adj} \right|^2 \right] + E \left[\left| Z_l[k] \right|^2 \right] \quad (19)$$

If we assume the variance of $X_{l,D}[k_{l,D}]$ and $X_{l,I}[k_{l,I}]$ are equal to the σ_x^2 , the power of ICI_{self} and ICI_{adj} are obtained as follows:

$$E \left[\left| ICI_{self} \right|^2 \right] = E \left[\left| \sum_{\substack{s \in \mathbf{K}_{l,D} \\ s \neq k_{l,D}}} C_{l,D}(\varepsilon_D, s - k_{l,D})H_{l,D}[s]X_{l,D}[s] \right|^2 \right] = \sum_{\substack{s \in \mathbf{K}_{l,D} \\ s \neq k_{l,D}}} \left| C_{l,D}(\varepsilon_D, s - k_{l,D})H_{l,D}[s] \right|^2 \sigma_x^2$$

$$E \left[\left| ICI_{adj} \right|^2 \right] = E \left[\left| \sum_{s \in \mathbf{K}_{l,I}} (C_{l,I}(\varepsilon_I, s - k_{l,D})H_{l,I}[s] \cdot \sqrt{P_{rx,I}/P_{rx,D}} \cdot X_{l,I}[s]) \right|^2 \right] = \sum_{s \in \mathbf{K}_{l,I}} \left| C_{l,I}(\varepsilon_I, s - k_{l,D})H_{l,I}[s] \right|^2 \frac{P_{rx,I}}{P_{rx,D}} \sigma_x^2 \quad (21)$$

At last, the power of $Z_l[k_{l,D}]$ is obtained as follows:

$$E \left[\left| Z_l[k_{l,D}] \right|^2 \right] = \sigma_z^2 \quad (22)$$

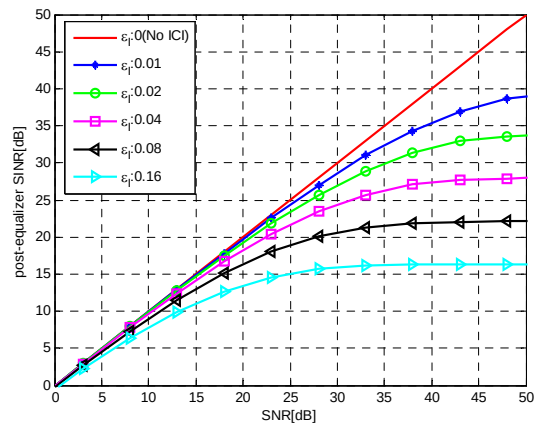


Figure 3. SNR vs. post-equalizer SINR based on the IEEE 802.16n direct communication frame structure according to the normalized carrier frequency

offset in case of the SIR is -20dB and frequency offset of the desired signal is zero.

If the subcarrier allocation is fixed, the $SINR_{i,k}$ is the function of ε_D , ε_i and SIR . Figure 7 is the numerically calculated SNR vs. post-equalizer SINR according to the normalized carrier frequency offset and SIR on the IEEE 802.16n direct communication frame structure.

IV. SIMULATION

In this section, we conduct the link level simulation and PHY abstraction on the same condition to verify the validation of the PHY abstraction method by applying physical layer parameters of IEEE 802.16n direct communication. The simulation parameters and assumption are listed in Table 1. The link curves used to estimate the BLER is shown as figure 4. The frame structure, resource allocation and interference scenario for simulation is explained in the section IV-A and IV-B. We performed the simulation for various modulation orders, code rates and data burst sizes.

TABLE 1. SIMULATION PARAMETERS AND ASSUMPTIONS

Parameters	Value	Note
Frame structure	IEEE 802.16n direct communication	Section IV-A
Used sub-channel	Disired signal: Sub-Channel 0 Interference signal: Sub-Channel 1	Section IV-A
Interference Scenario	Adjacent channel interference by CFO of the interference signal	Section IV-B
Channel Model	AWGN, Urban Micro (NLOS)	
Link-to-system Mapping method	MIESM-RBIR	Section III-A
Nominal bandwidth	10MHz	
FFT size	1024	
CP length	1/8 (128 samples)	
Carrier frequency	2.3GHz	
Sampling frequency	11.2 MHz	
MIMO Schemes	Space frequency block coding (SFBC)	Receiver: MRC
Antenna Configuration	Transmitter: 2 Receiver: 2	
MCS level (M × Rc)	2(4QAM) × 0.370 = 0.740 4(16QAM) × 0.463 = 1.852 6(64QAM) × 0.914 = 5.484	M: Modulation order Rc: Code rate
Data burst size [bit] (S _b)	320 for MCS level 0.740 800 for MCS level 1.852 2368 for MCS level 5.484	
Forward error correction	Turbo code	Mother code rate : 1/3
Normalized CFO	[-0.02, -0.04, -0.08]	
SIR [dB]	[0, -20 -40]	

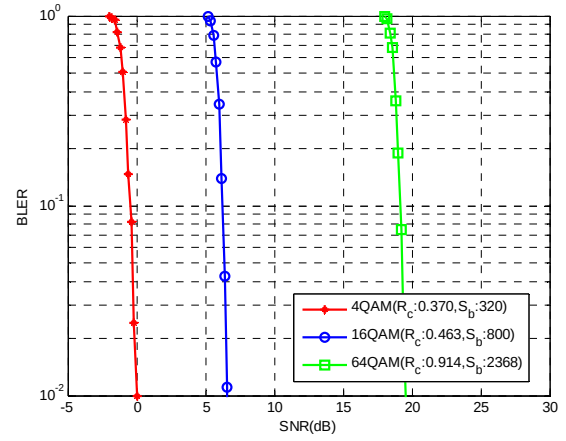


Figure 4. Link curves for BLER estimation

A. Frame Structure for Direct Communication in IEEE 802.16n

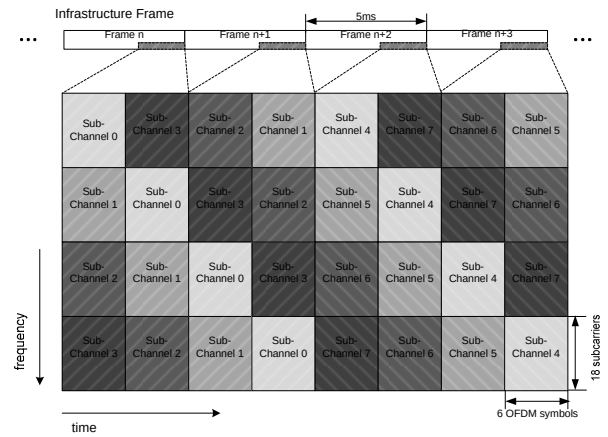


Figure 5. Example of frame structure for the direct communication based on the IEEE 802.16n

Figure 2 shows the example of the frame structure for direct communication based on in IEEE 802.16n. Some part of the infrastructure resource is assigned for the mobile station (MS) which want to directly communicate each other without controlling of the BS. There can be multiple sub-channels in the direct communication frame and the MS tries to reserve the resource by contention-based way and interference avoidance mechanism is operated. A sub-channel consists of 4 sub-block with 18 subcarriers by 6 OFDM symbols.

B. Interference Scenario

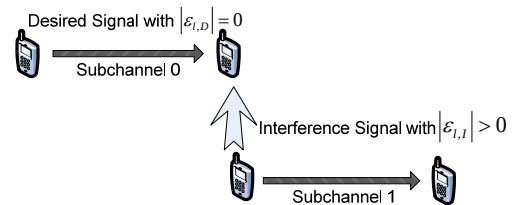


Figure 6. Adjacent channel interference scenario

In direct communication environment, there may be large ICI according to the position of devices. For PHY abstraction in ICI environment caused by CFO, we should take adjacent channel interference into consideration as shown in the Figure 6. Since the devices which directly communicate each other can estimate accurately the CFO among them, the effect of CFO the desired signal is not critical in practical communication environment. So we assume that there is the ICI from interference signal only.

C. Simulation Results

Figure 7-9 and figure 10-12 show BLER performance according to the CFO and SIR in AWGN and urban micro (UMi) non-line of sight (NLOS) channel, respectively. BLER estimation is conducted by using the proposed PHY abstraction method at the same time. The dotted lines are the simulation result and the markers are the estimated result. When the power of the interference signal is small, BLER performance is almost same as the no ICI case. But the performance degradation occurs even if the CFO is small when the power of the interference signal is large. Anyway, the predicted BLER is very little different from the simulation results regardless of the CFO and SIR.

A little difference between simulation and estimated result are observed after error floor occurred as shown in the figure 8, especially 64QAM mode with large CFO. However, MS will not try to communicate with high modulation order in those cases because the power of interference is too large to satisfy the requirement such as error vector magnitude (EVM). So it will not have significant impact in system level simulation.

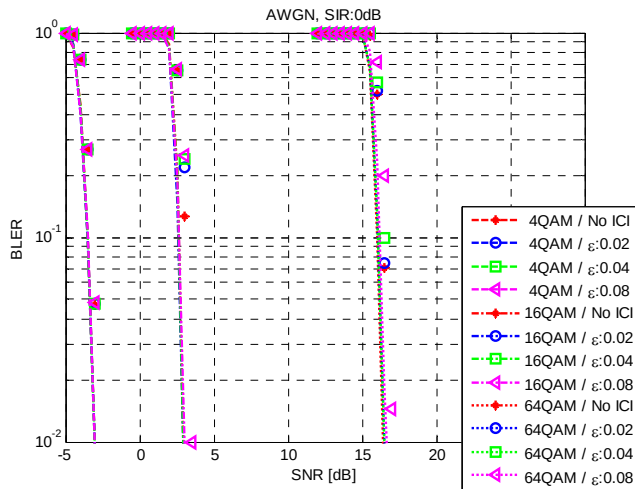


Figure 7. BLER in AWGN, SIR: 0dB – Simulation vs. Estimated (dotted lines: simulation results, markers: estimated results)

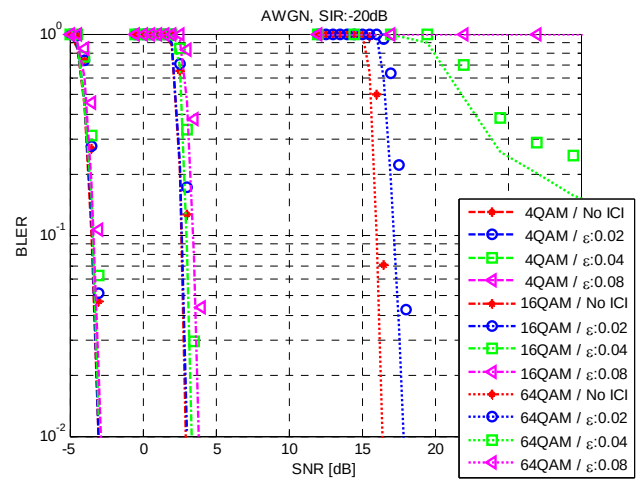


Figure 8. BLER in AWGN, SIR: -20dB – Simulation vs. Estimated (dotted lines: simulation results, markers: estimated results)

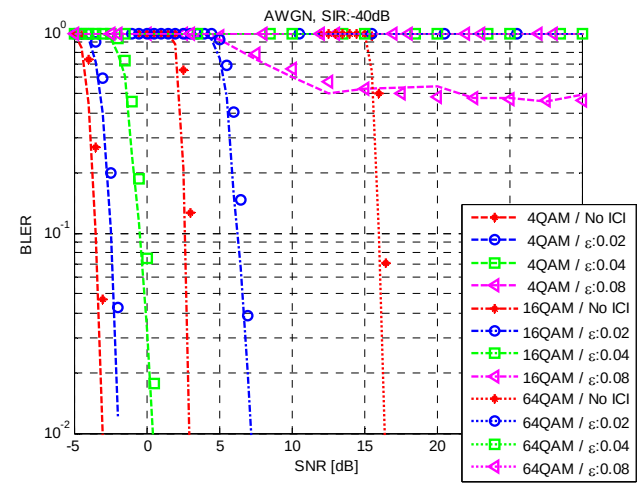


Figure 9. BLER in AWGN, SIR: -40dB – Simulation vs. Estimated (dotted lines: simulation results, markers: estimated results)

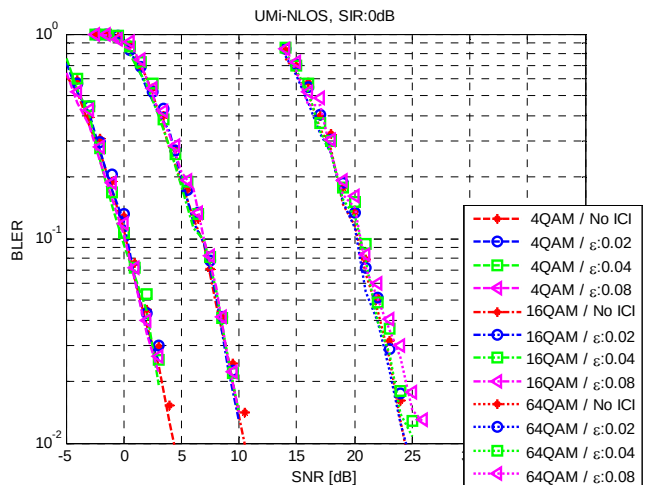


Figure 10. BLER in UMi, SIR: 0dB – Simulation vs. Estimated (dotted lines: simulation results, markers: estimated results)

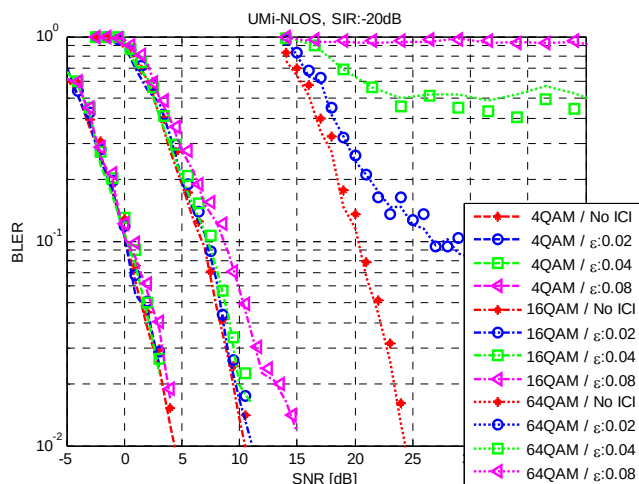


Figure 11. BLER in UMi, SIR: -20dB – Simulation vs. Estimated (dotted lines: simulation results, markers: estimated results)

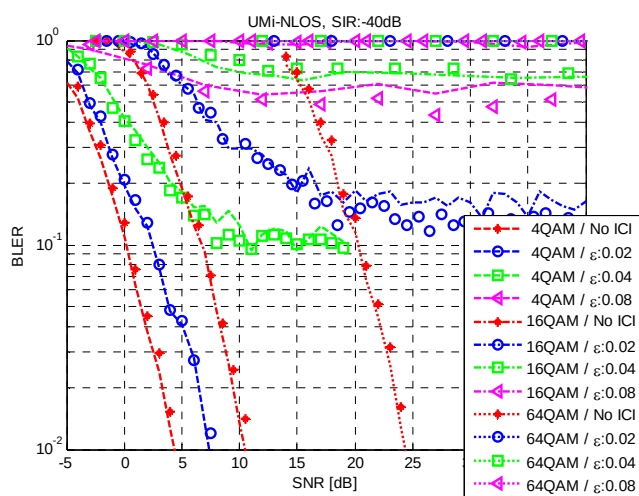


Figure 12. BLER in UMi, SIR: -40dB – Simulation vs. Estimated (dotted lines: simulation results, markers: estimated results)

V. CONCLUSIONS

In this paper, we proposed PHY abstraction method to predict BLER in ICI environment caused by CFO for system level performance evaluation. We can dynamically calculate the effective SINR based on the normalized CFO, channel of desired signal and interference signal and SIR. We conducted the link level simulation and PHY abstraction on the same

condition to verify the validation of the PHY abstraction method. We confirmed that there is little difference between the simulation results and the PHY abstraction results. The proposed method can be simply inserted in system level simulation. Although, the simulation is performed on the IEEE 802.16n frame structure, the PHY abstraction method is applicable to any other system based on OFDM and MIMO with linear receiver.

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