Performance Analysis of 3GPP-LTE Downlink Control Channel (PHICH Channel) with Perfect and Imperfect Channel State Information

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ABSTRACT

LTE is an all-IP wireless network based on orthogonal frequency division multiplexing (OFDM), and over a reframed GSM band with bandwidth from 1.4M Hz to 20M Hz. The physical channels defined in the downlinks include Physical Downlink Shared Channel (PDSCH), Physical Multicast Channel (PMCH), Physical Downlink Control Channel (PDCCH), Physical Broadcast Channel (PBCCH), Physical Control Format Indicator Channel (PCFICH) and Physical Hybrid ARQ Indicator Channel (PHICH). In this paper the performance of PHICH Channel is analyzed for various fading channel models and maximum likelihood-based (ML) receiver structures are derived for the decoding of this downlink control channel in the new long term evolution (LTE) standard under the assumption that the channel state information is perfect. The performance of the same is being analyzed over Ped-B channel with imperfect channel state information. The channel is being estimated through various channel estimation techniques such as Maximum Likelihood (ML) criterion, Minimum Mean Square Error and Modified Minimum Mean Square Error and the comparison between them is made.

Key words: LTE, OFDM, PHICH, UE, BS, ML, MMSE

INTRODUCTION

With the emergence of packet-based mobile broadband systems, it is evident that a comprehensive long term evolution (LTE) is required to remain competitive in the long term. Long term goals for the system include support for high peak data rates (100 Mbps downlink and 50 Mbps uplink), low latency, improved system capacity and coverage, reduced operating costs, multi-antenna support, efficient support for packet data transmission, flexible bandwidth operations (up to 20 MHz) and seamless integration with existing systems [3]. To reach these goals, a new design for the air interface is adopted. This includes an efficient control channel design to reduce the overhead required to support data transmission. In the downlink, this control signalling encompasses scheduling grants, a control format indicator, and H-ARQ acknowledgments. This paper provides an overview of downlink control channel (PHICH Channel) design and the performance of the same is analyzed with perfect and imperfect channel state information. The paper is organized as follows. In Section II, an overview of downlink structure is provided. Section III discusses on Transmit Processing of the PHICH channel. The decision metric for the H-ARQ Channel is derived and its performance is analyzed under the assumption of perfect CSI in Section IV. The Section V provides an insight of various estimation techniques with the simulation results. This paper focuses on the performance analysis of the PHICH between the user equipment (UE) and the base station (BS) in three types of channels: (1) static (additive white Gaussian noise (AWGN), (2) frequency flat fading, (3) Ped-B channel models.

DOWNLINK FRAME STRUCTURE

In the downlink OFDM is selected as the air interface and it is straight forward to exploit frequency selectivity of the multi-path channel with low complexity receivers. Furthermore, due to its frequency domain nature, OFDM enables flexible bandwidth operation with low complexity. Table 1 provides the LTE downlink system numerology [1].

Table 1 LTE Downlink System Numerology

Channel Bandwidth (MHz)	1.4	3.0	5.0	10.0	15.0	20.0
Frame duration	10 ms					
Sub-carrier spacing	15 kHz					
Number of Physical Resource Blocks N_{RB}^{DL}	6	15	25	50	75	100
FFT Size N	128	256	512	1024	1024	2048
Sampling Frequency (MHz) f_s	1.92	3.84	7.68	15.36	23.04	30.72

The downlink subframe structure is shown in [1]. Each frame is comprised of two slots of length 0.5ms (7 OFDM symbols). Within each slot reference symbols are located in the 1st and 5th OFDM symbols. This structure allows low complexity high performance channel estimation techniques such as ML, MMSE and IFFT based estimators.

PHYSICAL HYBRID ARQ INDICATOR CHANNEL

The PHICH carries physical hybrid - ARQ ACK/NACK. Data arrives to the coding unit in form of indicators for HARQ acknowledgement [3]. Figure 1 shows the PHICH transport channel and physical channel processing on hybrid-ARQ data, w_n is the spreading code for n-th user in a PHICH group, obtained from an orthogonal set of codes. In LTE, $2M$ spreading sequences are used in a PHICH group, where $M = 4$ for normal CP and 2 for extended CP. The first set of M spreading sequences are formed by $M \times M$ Hadamard matrix, and the second set of M spreading sequences are in quadrature to the first set.

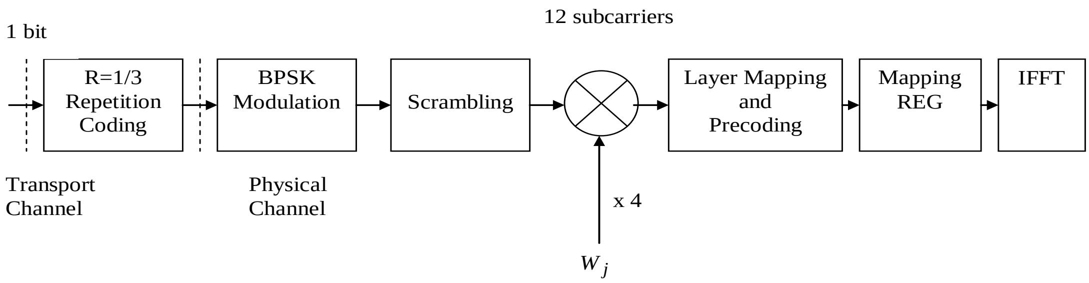

Fig 1: PHICH Transmit Processing

The repetition code is performed on the data transmitted from the BS to the UE and then the coded data is modulated using $\pi/4$ BPSK modulation scheme. In order to obtain orthogonality between the coded bits, they are multiplied with the help of the Hadamard sequence as shown in [1] and then transmitted.

PHICH WITH SINGLE TRANSMIT ANTENNA

The received signal is processed and the decision metric is derived as follows. The resource elements are demapped and the output representing the i-th resource element group and the k-th receive antenna is given by,

$$\mathrm{y}_{i,k} = \mathrm{h}_{i,k} \circ \left(x_1 \sqrt{\frac{P_1}{2}} \mathrm{w}_1 + \sum_{n=2}^{M} \sqrt{\frac{P_n}{2}} \mathrm{w}_n x_n + j \sum_{n=1}^{M} \sqrt{\frac{\tilde{P}_n}{2}} \mathrm{w}_n \tilde{x}_n \right) + \mathrm{u}_{i,k}, i = 1,2,3 \quad \ldots (1)$$

where $\mathbf{y}_{i,k}$ is a $M \times 1$ vector, P_n and $\tilde{P}_n$ $n = 1, \ldots M$ are the power levels of the M orthogonal codes (for the normal CP case), $x_1 \in (1,-1)$ is the data bit value of the desired user HI, x_n and $\mathbf{h}_{i,k}$ is a $M \times 1$ complex channel frequency response vector. Without loss of generality it is assumed that the desired HI channel to be decoded uses the first orthogonal code denoted as $\mathbf{w}_1$. The second and third terms in equation (1) denote the remaining $2M - 1$ spreading codes used for the other HI channels within a PHICH group. The term $\mathbf{u}_{i,k}$ denotes the thermal noise, which is modelled as circularly symmetric zero-mean complex Gaussian with covariance $E\left[\mathbf{u}_{i,k}\mathbf{u}_{i,k}^H\right]=\sigma_u^2\mathbf{I}$. The ML decoding is given by $z=\sum_{k=1}^{K} z_k$ where K is the number of antennas at the UE receiver and $z_k=\sum_{i=1}^{3} z_{i,k}$ where $z_{i,k}=\mathrm{Re}\left\{\left\langle \mathbf{y}_{i,k}\circ\hat{\mathbf{h}}_{i,k}^*,\mathbf{w_1}\right\rangle\right\}$ … (2)

where the estimated channel frequency response $\hat{\mathbf{h}}_{i,k}$ is given by $\hat{\mathbf{h}}_{i,k}=\mathbf{h}_{i,k}+\mathbf{e}_{i,k}$, $\mathbf{e}_{i,k}$ is the estimation error which is uncorrelated with $\mathbf{h}_{i,k}$ and zero mean complex Gaussian with covariance $\sigma_e^2\mathbf{I}$. By expanding (2), we get that

$$z_{i,k}=\mathrm{Re}\left(\begin{array}{l}\left\langle \mathbf{h}_{i,k}\circ\hat{\mathbf{h}}_{i,k}^*\circ\mathbf{w_1},\mathbf{w_1}\right\rangle x_1\sqrt{\frac{P_1}{2}}+\sum_{n=2}^{M}\left\langle \mathbf{h}_{i,k}\circ\hat{\mathbf{h}}_{i,k}^*\circ\mathbf{w}_n,\mathbf{w_1}\right\rangle\sqrt{\frac{P_n}{2}}x_n\\ +j\sum_{n=1}^{M}\left\langle \mathbf{h}_{i,k}\circ\hat{\mathbf{h}}_{i,k}^*\circ\mathbf{w}_n,\mathbf{w_1}\right\rangle\sqrt{\frac{\tilde{P}_n}{2}}\tilde{x}_n+\left\langle \mathbf{u}_{i,k}\circ\hat{\mathbf{h}}_{i,k}^*,\mathbf{w_1}\right\rangle\end{array}\right) \quad \text{… (3)}$$

Note that $\left\langle \mathbf{w}_i,\mathbf{w}_j\right\rangle=\begin{cases}M, i=j\\ 0, i\neq j\end{cases}$. Thus (3) becomes

$$\begin{aligned} z_k=&\sum_{i=1}^{3}\sum_{m=1}^{M}\left|h_{i,k}^{(m)}\right|^2\sqrt{\frac{P_1}{2}}x_1+\mathrm{Re}\left(\sum_{i=1}^{3}\sum_{m=1}^{M}h_{i,k}^{(m)}e_{i,k}^{(m)*}\right)\sqrt{\frac{P_1}{2}}x_1-\mathrm{Im}\left(\sum_{i=1}^{3}\sum_{m=1}^{M}h_{i,k}^{(m)}e_{i,k}^{(m)*}\right)\sqrt{\frac{\tilde{P}_1}{2}}\tilde{x}_1\\ &+\mathrm{Re}\left(\sum_{i=1}^{3}\sum_{m=1}^{M}h_{i,k}^{(m)*}u_{i,k}^{(m)}\right)+\mathrm{Re}\left(\sum_{i=1}^{3}\sum_{m=1}^{M}e_{i,k}^{(m)*}u_{i,k}^{(m)}\right)\end{aligned} \quad \text{… (4)}$$

For ideal channel estimation, error variance is zero, and then due to the orthogonality property of the spreading codes, no interference is introduced to $\mathbf{w}_1$ from the other HI channels within a PHICH group.

However, in the presence of channel estimation error, self-interference and co-channel interference are introduced as seen in the second and third terms, respectively in (4). Since $\left|\tilde{x}_1\right|^2=1$ and $\left|x_1\right|^2=1$, the SINR of the decision statistic z is thus given by

$$\gamma_z^{non-idealCE}=\sum_{k=1}^{K}\frac{P_1\left(\left(\sum_{i=1}^{3}\sum_{m=1}^{M}\left|h_{i,k}^{(m)}\right|^2\right)^2\right)}{\frac{\sigma_e^2\left(P_1+\tilde{P}_1\right)}{2}\left(\sum_{i=1}^{3}\sum_{m=1}^{M}\left|h_{i,k}^{(m)}\right|^2\right)+\sigma_u^2\sum_{i=1}^{3}\sum_{m=1}^{M}\left|h_{i,k}^{(m)}\right|^2+3M\sigma_u^2\sigma_e^2} \quad \text{… (5)}$$

In the case of a static AWGN channel with a single antenna at the UE receiver, i.e., $h_{i,k}^{(m)}=h,\forall i,m,k$, the SINR is simply given by

$$\gamma_z^{non-idealCE}=\frac{P_G P_1\left|h\right|^4}{0.5\sigma_e^2\left(P_1+\tilde{P}_1\right)\left|h\right|^2+\sigma_u^2\left|h\right|^2+\sigma_u^2\sigma_e^2} \quad \text{… (6)}$$

where P_G =12 in equation (6) is the processing gain obtained from the spreading code of length 4, and (3,1) repetition code in the case of normal CP. For ideal channel estimation, $\sigma_e^2 = 0$ and the SNR of the decision statistic z is thus given by

$$\gamma_z^{idealCE} = \frac{P_G P_1 |h|^2}{\sigma_u^2} \quad \ldots (7)$$

The probability of error in the AWGN case with a single receive antenna is simply

$$P_b^{(HI)} = \frac{1}{2} P\left(z < 0 \middle| HI = 0\right) + \frac{1}{2} P\left(z > 0 \middle| HI = 1\right) = \frac{1}{2} erfc\left(\sqrt{P_G \gamma}\right) \quad \ldots (8)$$

where γ is the SNR. The PHICH performance for SISO and SIMO systems in static AWGN is shown in Figure 3

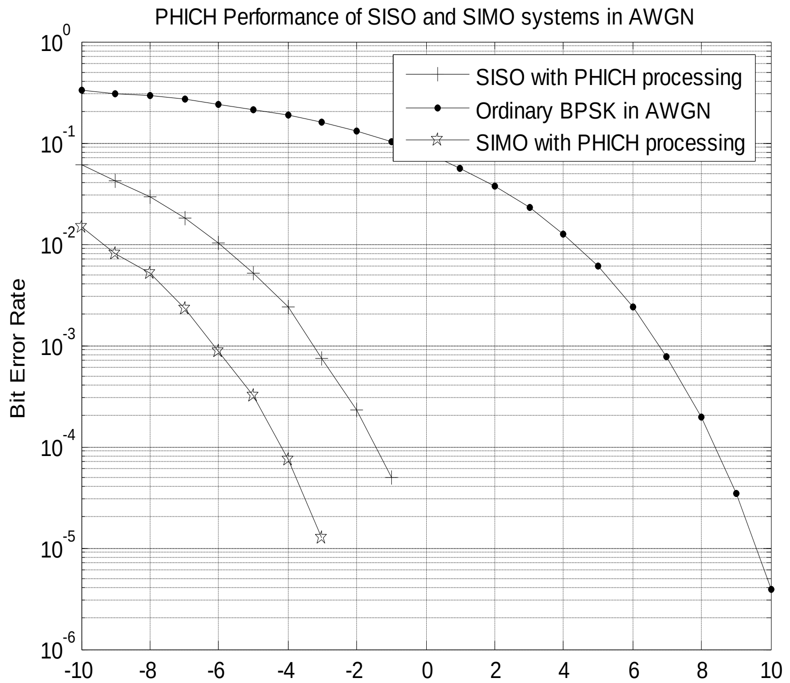

Figure 2 PHICH Performance of SISO and SIMO systems in AWGN

In Figure 2 the comparison is made with the ordinary BPSK modulated curve. It is found that there is a 3dB gain between the SISO and SIMO. The probability of error formula for an ordinary BPSK modulation in AWGN [4] is given by,

$$P_e^{BPSK} = \left(\frac{1}{2}\right) erfc\left(\sqrt{\gamma}\right) \quad \ldots (9)$$

where γ is the signal to noise ratio in dB. Now these systems' performance for Rayleigh flat fading channel is studied through Figures 3 and 4.

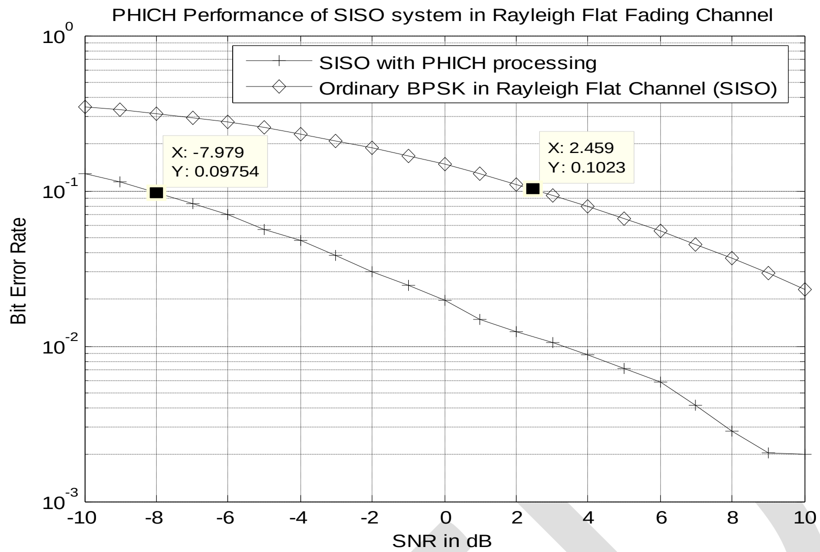

Figure 3 PHICH Performance of SISO systems in Rayleigh Flat Fading Channel

The probability of error formula for the BPSK modulated curve in Rayleigh flat fading channel [4] for a SISO system is given by,

$$P_e = E\left[Q\left(\sqrt{2|h|^2 SNR}\right)\right] = \frac{1}{2}\left(1-\sqrt{\frac{SNR}{1+SNR}}\right) \quad \dots (10)$$

The SISO PHICH system has got approximately 10.438dB gain against the BPSK modulation in Flat fading channel because the processing gain of the PHICH system is 12.

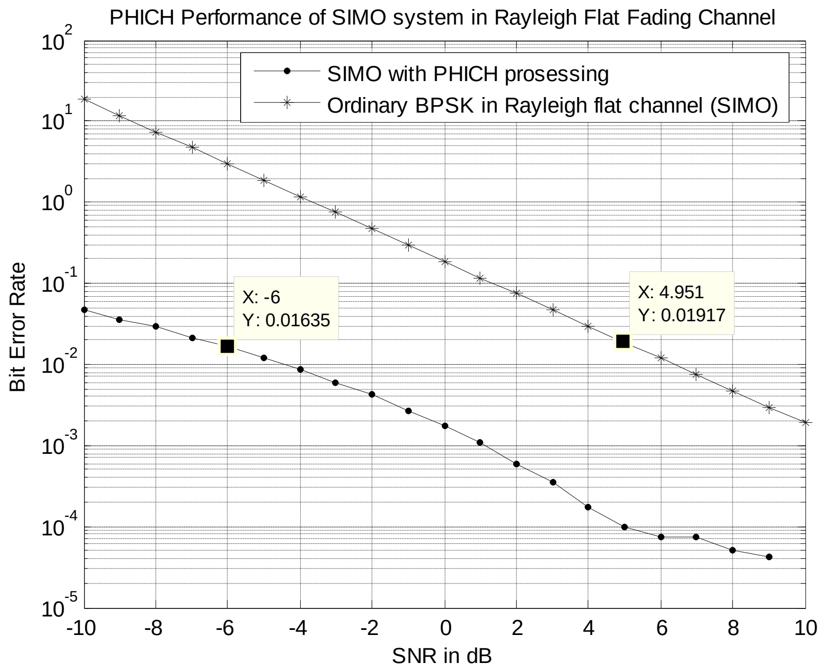

Figure 4 PHICH Performance of SIMO system in Rayleigh Flat Fading Channel

The probability of error formula for the BPSK modulated curve in Rayleigh flat fading channel for a SIMO system [4] is given by,

$$P_e = \left(\frac{1}{4\bar{\gamma}_c}\right)^L \binom{2L-1}{L} \quad \ldots (11)$$

where $\bar{\gamma}_c = \dfrac{2E}{N_0} E\left(\alpha_k{}^2\right)$

The SIMO PHICH system has got approximately 10.951dB gain against the BPSK modulation in Flat fading channel.

PHICH WITH MULTIPLE TRANSMIT ANTENNAS

The received signal is processed and the decision metric is derived as follows. The resource elements are demapped The output at the l-th layer on the k-th receive antenna, and i-th resource element group (REG) is given by

$$\mathbf{y}_{l,k}^{(i)} = \mathbf{H}_{l,k}^{(i)} \mathbf{d}_l^{(i)} + \mathbf{u}_{l,k}^{(i)} \quad 0 \le l \le M_{symb}^{layer} - 1,\ 1 \le k \le K,\ i = 1,2,3 \quad \text{s}\ldots (12)$$

where $M_{symb}^{layer} = \dfrac{M_{symb}}{3\times 2} = 2$, $\mathbf{y}_{l,k}^{(i)}$ is a 2×1 received signal vector, $\mathbf{d}_l^{(i)}$ is 2×1 transmit signal vector and $\mathbf{u}_{l,k}^{(i)}$ denotes 2×1 thermal noise vector, and its each element is modelled as circularly symmetric zero-mean complex Gaussian with covariance $E\left[\mathbf{u}_{l,k}^{(i)} \mathbf{u}_{l,k}^{(i)H}\right] = \sigma_u^2 \mathbf{I}$. The channel matrix $\mathbf{H}_{l,k}^{(i)}$ is given by

$$\mathbf{H}_{l,k}^{(i)} = \frac{1}{\sqrt{2}} \begin{bmatrix} h_{k,1}^{(l)(i)} & -h_{k,2}^{(l)(i)} \\ h_{k,2}^{(l)(i)*} & h_{k,1}^{(l)(i)*} \end{bmatrix} \quad \ldots (13)$$

where $h_{k,m}^{(l)(i)}$ is complex channel frequency response between m-th transmit antenna and k-th receive antenna, at l-th symbol layer in i-th REG. The transmit signal vector $\mathbf{d}^{(i)}$ is generated by layer mapping and precoding the HI data vector $\mathbf{x}$ in i-th REG. The 4×1 vector $\mathbf{x}$ is given by

$$\mathbf{x} = x_1 \sqrt{\frac{P_1}{2}} \mathbf{w}_1 + \sum_{n=2}^{M} \sqrt{\frac{P_n}{2}} \mathbf{w}_n x_n + j \sum_{n=1}^{M} \sqrt{\frac{\tilde{P}_n}{2}} \mathbf{w}_n \tilde{x}_n \quad \ldots (14)$$

P_n and $\tilde{P}_n$ $n = 1,2,3,4$ are the power levels of the 8 spreading codes.

The ML decision statistic, is given by $z = \sum_{k=1}^{K} \tilde{z}_k$ where

$$\tilde{z}_k = \sum_{i=1}^{3} \tilde{z}_k^{(i)} = \sum_{i=1}^{3} \mathrm{Re}\left(\sum_{l=0}^{M_{symb}^{layer}-1} \left\langle \mathbf{z}_{l,k}^{(i)}, \mathbf{w}_1 \right\rangle \right) \quad \ldots (15)$$

where $\mathbf{z}_{l,k}^{(i)} = \mathbf{H}_{l,k}^{(i)^H}\mathbf{H}_{l,k}^{(i)}\mathbf{d}_l^{(i)} + \mathbf{H}_{l,k}^{(i)^H}\mathbf{u}_{l,k}^{(i)} \quad 0 \le l \le M_{symb}^{layer} - 1,\ 1 \le k \le K, i = 1,2,3$ ... (16)

In flat fading channel, $\mathbf{H}_{l,k}^{(i)} = \mathbf{H}_k\ \forall l,i$. Then the decision statistic z is given by,

$$z = \sum_{k=1}^{K}\sum_{i=1}^{3}\left(\mathbf{H}_k^H\mathbf{H}_k\sum_{l=0}^{M_{symb}^{layer}-1}\mathrm{Re}\left(\left\langle \mathbf{d}_l^{(i)}, \mathbf{w}_1\right\rangle\right) + \sum_{l=0}^{M_{symb}^{layer}-1}\mathrm{Re}\left(\left\langle \mathbf{H}_k^H\mathbf{u}_{l,k}^{(i)}, \mathbf{w}_1\right\rangle\right)\right) \quad \text{… (17)}$$

The instantaneous SNR of z is evaluated to be

$$SNR_{\tilde{z}_k} = \sum_{k=1}^{K}\frac{6P_1\left(\left|h_{k,1}\right|^2 + \left|h_{k,2}\right|^2\right)}{\sigma_u^2} \quad \text{… (18)}$$

In the case of a static AWGN channel with a single antenna at the UE receiver, i.e., $h_{i,k} = h, \forall i,k$, the SNR is given by $SNR_{\tilde{z}_k} = |h|^2\frac{12P_1}{\sigma_u^2}$. The probability of error is given by (8).

The PHICH performance for MISO and MIMO systems in static AWGN is shown in Figure 5.

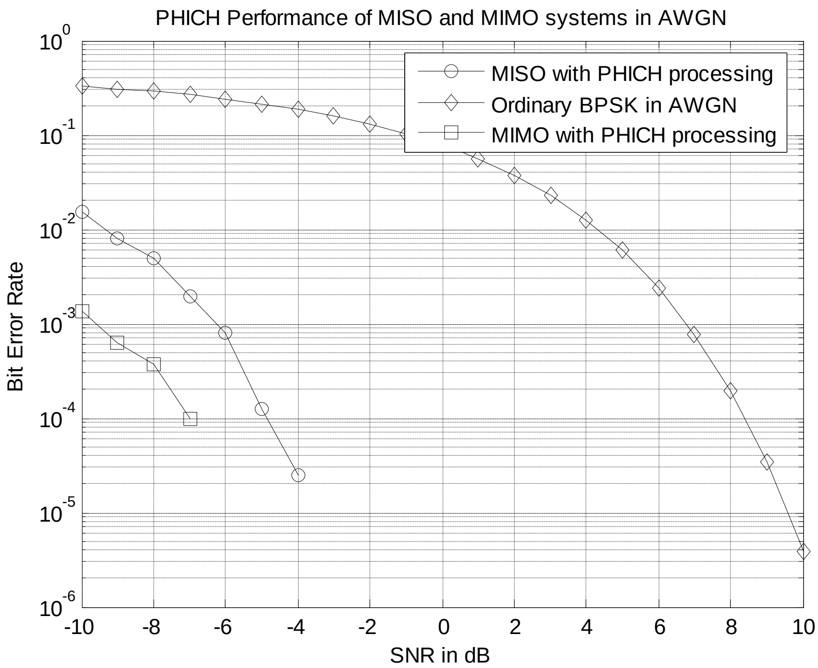

Figure 5 PHICH Performance of MISO and MIMO systems in AWGN

In this figure the performance of the PHICH channel in a MISO and MIMO systems are compared with the ordinary BPSK modulated curve. It is found that the PHICH channel has a processing gain of 24 and so there is a 3dB gain for a MIMO PHICH system to that of MISO PHICH system. The probability of error formula for an ordinary BPSK modulation in AWGN is given by equation (9).

Now these systems' performance for Rayleigh flat fading channel is studied through Figures 6 and 7.

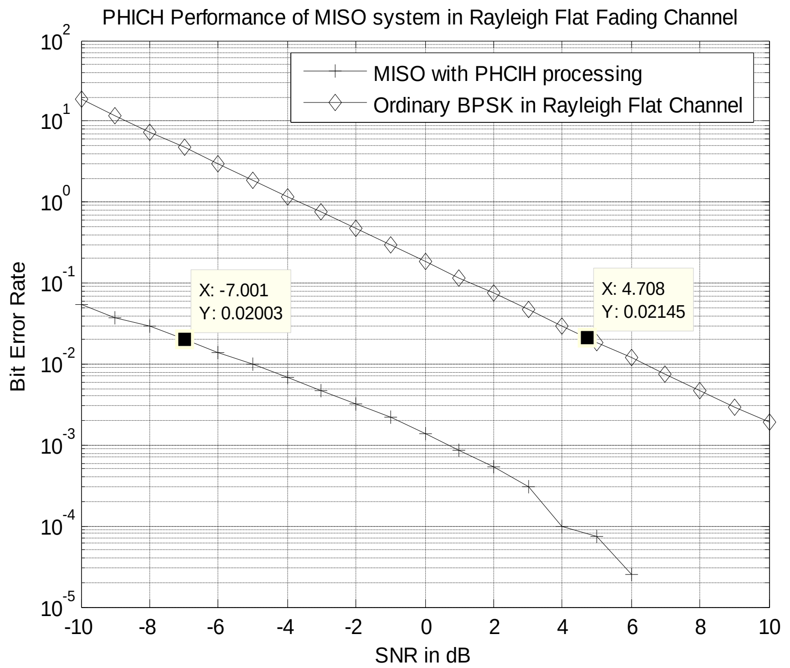

Figure 6 PHICH Performance of MISO system in Rayleigh Flat Fading Channel

The probability of error formula for the BPSK modulated curve in Rayleigh flat fading channel for a MISO and MIMO systems is given by equation (11).

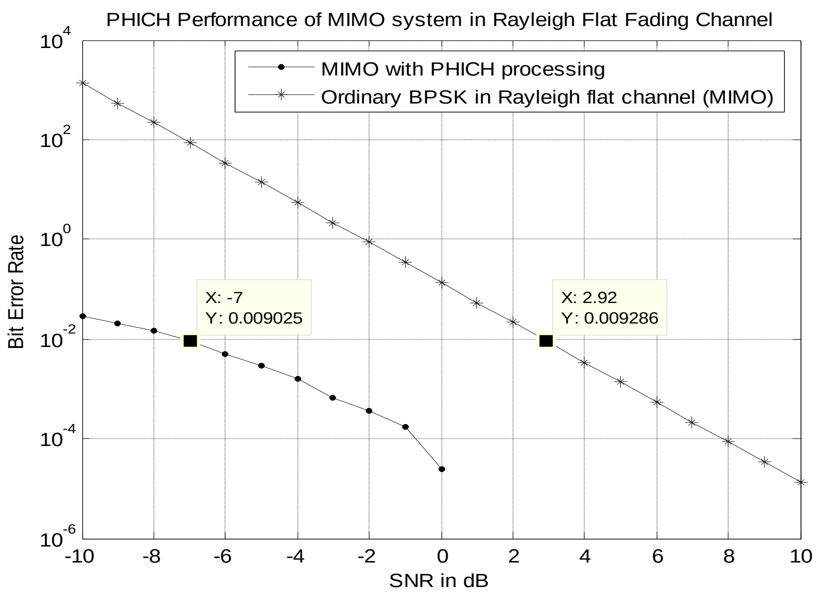

Figure 7 PHICH Performance of MIMO system in Rayleigh Flat Fading Channel

The performance of PHICH channel is analysed for various systems in different channels under the assumption that the channel state information is perfect. And the receiver structure described above is termed as Genie Receiver [8]. Now considering that the channel state information is imperfect, the channel is being estimated through various estimation techniques and they are briefly explained in the following section and the PHICH channel performance including this is clearly shown through simulations.

ESTIMATION TECHNIQUES

A. ML Estimation (Frequency Domain)

Consider the received signal model as $Y = HX + N$. Here Y is the received signal vector, H is complex channel frequency response vector, X is the combined transmitted vector of all the 8 users and N is the circularly symmetric zero-mean complex Gaussian noise. Then the ML estimate-on is given by,

$$\hat{\mathbf{H}}_{FD} = \left(\mathbf{X}^T \mathbf{X}\right)^{-1}\left(\mathbf{X}^T \mathbf{Y}\right) \quad \ldots (19)$$

The channel is estimated at the pilot symbol locations and the channels in all other locations are obtained through linear interpolation.

B. ML Estimation (Time Domain)

The ML estimate uses the knowledge of number of channel taps. From the Least Squares (LS) estimation, the ML estimation is derived.

$$\mathbf{H}_{LS} = \mathbf{X}^{-1}\mathbf{Y} \quad \ldots (20)$$

$$\hat{h}_{ML} = \left(\mathbf{B}^H \mathbf{B}\right)^{-1}\left(\mathbf{B}^H \mathbf{H}_{LS}\right) \quad \ldots (21)$$

where B is the FFT matrix which considers the maxim-um length of the channel taps.

$$\hat{\mathbf{H}}_{TD} = FFT\left(\hat{h}_{ML}\right) \quad \ldots (22)$$

C. MMSE Estimation

The MMSE estimation for the frequency response at the pilot symbol locations is given by [7],

$$\hat{H}_{MMSE} = R_{HH}\left[R_{HH} + \sigma_N^2\left(XX^H\right)^{-1}\right]^{-1}\hat{\mathbf{H}}_{LS} \quad \ldots (23)$$

where $R_{HH} = E\left(HH^H\right)$.

The MMSE estimator yields much better performance, especially under the low SNR scenarios. A major drawback of the MMSE estimator is its high computational complexity.

D. Modified MMSE Estimation

Modified MMSE Estimators reduces the complexity. Among them an optimal low-rank MMSE (OLR-MMSE) estimator is considered here and it combines the following three simplification algorithms [7].

The first simplification of MMSE estimator is to define the average SNR as $SNR = E\left\{|X_k|^2\right\}/\sigma_N^2$ and the term $\beta = E\left\{|X_k|^2\right\} \Big/ E\left\{1/|X_k|^2\right\}$ is a constant. Here the transmitted bits are the H-ARQ acknowledgments and they are $\pi/4$ BPSK modulated and hence $|X_k|^2 = 1$, therefore $\beta = 1$.

The second simplification is based on the low rank approximation. Here the taps with significant energy is alone considered and L is given by, $L = \left\lceil {T_G}/{T_S} \right\rceil N$, where N is the number of reference signal locations and $\left\lceil {T_G}/{T_S} \right\rceil$ is chosen among {1/32, 1/16,1/8,1/4}and so the effective size of the matrix is reduced dramatically.

The third simplification uses the singular value decomposition of the autocorrelation matrix of the generated channel. $R_{HH} = U \Lambda V^H$, where U is the unitary matrix containing the singular vectors, Λ is a diagonal matrix containing the singular values $\lambda_0 \geq \lambda_1 \geq \ldots \geq \lambda_{N-1}$ on its diagonal. The SVD also dramatically reduces the calculation complexity on the matrices.

Combining all simplification techniques, the OLR-MMSE estimator is explained as follows. The system first determines the number of ranks required by the estimator, denoted by p, which should be no smaller than $L+1$. Then given the signal constellation, the noise variance and the channel autocovariance matrix R_{HH} , the receiver pre-calculates β, SNR, the unitary matrix U and the singular values λ_k s. It thus obtains the $(N \times N)$ diagonal matrix Δ_p with entries,

$$\delta_k = \begin{pmatrix} \left[{\lambda_k}/{\left(\lambda_k + \frac{\beta}{SNR} \right)} \right], k = 0,1,\ldots, p-1 \\ 0, k = p,\ldots, N-1 \end{pmatrix} \quad \ldots (24)$$

The OLR estimator with rank p is thus given by,

$$H_{OLR-MMSE} = U \Delta_p U^H \hat{\mathbf{H}}_{LS} \quad \ldots (25)$$

SIMULATION RESULTS

In order to analyze the performance of PHICH Channel with imperfect channel state information the reference signal (Section 6.10 [1]) is generated in its locations and the PHICH data are transmitted in its locations as shown in (Section 6.9.3 [1]). The following simulation results undergo frequency selective fading (Ped-B channel). The power delay profile of this channel is given in Table 2.

Table 2 Power Delay Profiles for Pedestrian B Channel Model

Ped-B Channel Model	**Delay (n sec)**	0	200	800	1200	2300	3700
	Average Power (dB)	0	-0.9	-4.9	-8.0	-7.8	-23.9

The results are taken for various bandwidths shown in Table 1. As the bandwidth increases the number of resource elements and the number of reference signals also increases and hence the estimation would yield better results. This receiver structure is termed as Mismatched Receiver [8]. The simulation is carried out for both SISO and SIMO systems.

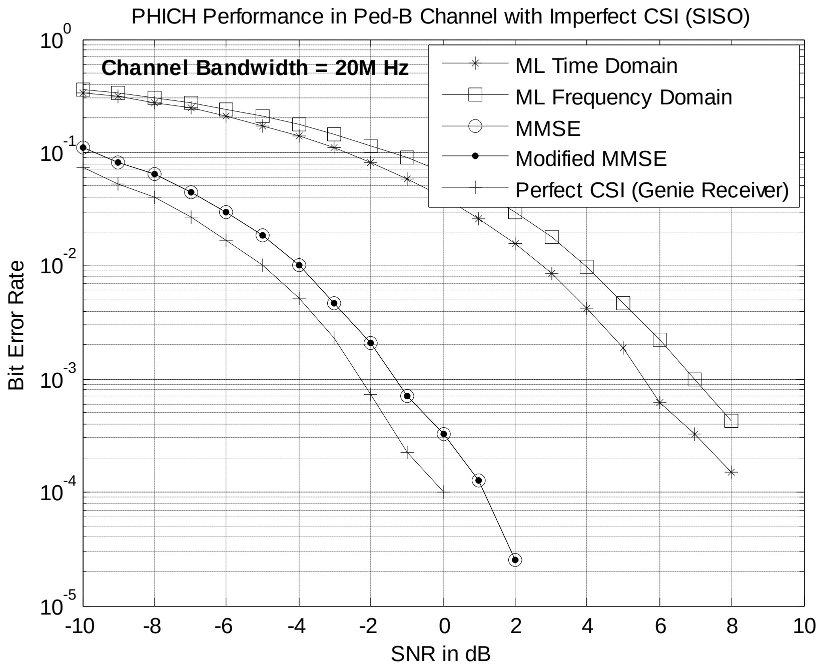

Figure 8 PHICH Performance in Ped-B Channel with Imperfect CSI (SISO) – BW = 20M Hz

The Figure 8 shows the simulation results of the performance of PHICH channel of a SISO system with bandwidth 20MHz experiencing Ped-B frequency selective fading environment. Here the CSI is imperfect and hence the channel is estimated through various techniques. In ML (Frequency domain) the increase in bandwidth does not make any significant change in the decibels (dB) difference with the Genie Receiver [8]. The difference is approximately 8dB.

In the ML (Time domain), the length of the channel taps are considered in generating the FFT matrix. The length increases with the bandwidth and hence the dB difference between Genie Receiver and ML (Time domain) does not have much change and it is approximately 9dB.

Whereas in MMSE estimation, the statistics of the channel is studied and is used in the estimation of that channel and hence it performs well. So the difference in dB between the Genie Receiver and MMSE varies from 5 dB to 1 dB as the bandwidth increases.

In the case of Modified MMSE the complexity is reduced through the simplification algorithms and hence it performs worse at lesser bandwidth but as the bandwidth increases it matches with the MMSE estimation.

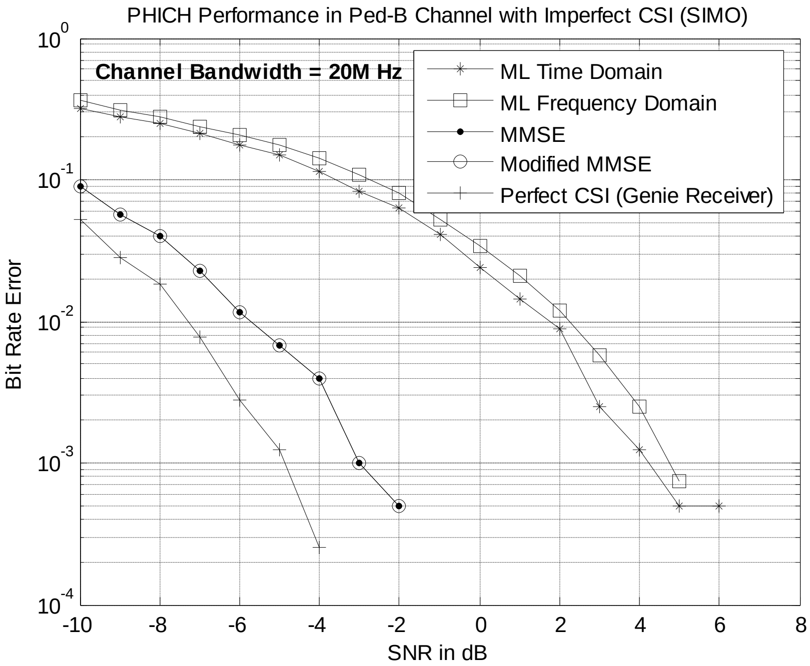

Figure 9 PHICH Performance in Ped-B Channel with Imperfect CSI(SIMO) – BW = 20M Hz

As it is a SIMO system Figure 9 shows that from the ML (Frequency) the ML (Time) has a 1 dB excess difference compared to the SISO system. In the case of MMSE the difference varies from 6 dB to 1.5 dB as there is an increase in bandwidth. The analysis of Modified MMSE holds good for both the systems. It is evidently shown from the above simulated results, as the bandwidth increases from 1.4MHz to 20 MHz the Modified MMSE estimation moves towards the MMSE channel estimation. MMSE channel estimation performs better than all the estimation techniques at very low SNR.

CONCLUSION

The LTE downlink (PHICH) channel's performance is analysed in the presence of AWGN and Flat fading environment for SISO, SIMO, MISO and MIMO systems with Genie Receiver structure (Perfect CSI). The ML (Frequency Domain), ML (Time Domain), MMSE and Modified MMSE Channel estimation techniques for LTE Downlink (PHICH) Channel has been thoroughly investigated. MMSE estimator using the channel statistics for estimation performs well. ML channel estimation method requires the knowledge of channel length for estimation. Because of this knowledge, it outperforms MMSE at high SNR when the channel length is decreased. Better channel estimation is achieved for all the bandwidth. The performance analysis in MSO and MIMO system of PHICH channel in Ped-B frequency selective fading environment is the focus of next-stage of work.

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