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MOBILITY AFFECTED ON CHANNEL ESTIMATION USING DIFFERENT MODULATION IN LTE

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Abstract *- In a relay network, mobility makes the entire set of channels time selective. With appropriate detection strategies that utilize partial CSI, performance improvement can be obtained over a complete non-coherent detection. Due to the multipath channel there is some inter-symbol interference (ISI) in the received signal. Therefore a signal detector (like MLSE or MAP) needs to know channel impulse response (CIR) characteristics to ensure successful equalization (removal of ISI). The channel estimation depicts that type of modulation not improves the spectral efficiency when speed became rise. It means that ICI will increase. In this work, channel estimation is implemented as an accelerator for LTE modem platforms to handle real-time channel estimation with different antenna configurations and mobility. And steady offset parameters, which produce ICI, lead to effected on channel performance. In this paper we analyze different approaches of channel estimation, like Least-squares (LS) or Linear Minimum Mean Squared Error (LMMSE) method. We presented many configuration antennas and used MMSE to increase data rate, which led to increase channel estimation and reduce ISI.*

Анотація *– У статті проведено дослідження впливу мобільності абонентів та різних видів модуляції на оцінку стану каналу в системі LTE. Оцінка враховує рівень міжсимвольної та міжканальної інтерференції. Для забезпечення максимальної пропускної здатності необхідно підтримувати високий рівень спектральної ефективності системи, що досягається завдяки використанню адаптивного вибору модуляційної схеми. Підвищення пропускної здатності для абонентів, що знаходяться на межі комірки, досягається шляхом використання механізму адаптивного формування діаграми спрямованості. Отримані результати показують досяжні межі спектральної ефективності системи з використанням різних видів модуляції.*

Аннотация *– В статье проведено исследование влияния мобильности абонентов и различных видов модуляции на оценку состояния канала в системе LTE. Оценка учитывает уровень межсимвольной и межканальной интерференции. Для обеспечения максимальной пропускной способности необходимо поддерживать высокий уровень спектральной эффективности системы, что достигается благодаря использованию адаптивного выбора модуляционной схемы. Повышение пропускной способности для абонентов, находящихся на грани ячейки, достигается путем использования механизма адаптивного формирования диаграммы направленности. Полученные результаты показывают достижимые пределы спектральной эффективности системы с использованием различных видов модуляции.*

Introduction

The 3rd Generation Partnership Project Long-Term Evolution (3GPP LTE) is an emerging mobile broadband communication system that offers high-speed mobile data service. LTE adopts Orthogonal Frequency Division Multiple Access (OFDMA) and Multiple-Input Multiple-Output (MIMO) to achieve a peak bit-rate target of 326 Mbit/s in the downlink (assuming a 4 × 4 MIMO system with 20 MHz bandwidth, 64 QAM, coding rate 1 and 19% pilot symbol overhead). The performance gain of MIMO heavily depends on the accurate estimation of Channel State Information (CSI), which is crucial for every communications system. As addressed in [1], the complexity of MIMO channel estimation is infeasible for most low-complexity receivers in practice. Hence, the pilots (reference signals) introduced in LTE have been chosen orthogonal to allow low-complexity channel estimation for multiantenna transmissions. In this work, a channel estimation it clearly that type of modulation not improve the spectral efficiency when speed became rise that mean OFDM that it make will lost or can be talk that ICI will increase.

I. Background for Channel Estimation

Fig.1 shows a generic simulation layout for a TDMA based mobile system, which exploits channel estimation and signal detection operations in equalization. The digital source is usually protected by channel coding and interleaved against fading phenomenon, after which the binary signal is modulated and transmitted over multipath fading channel. Additive noise is added and the sum signal is received.

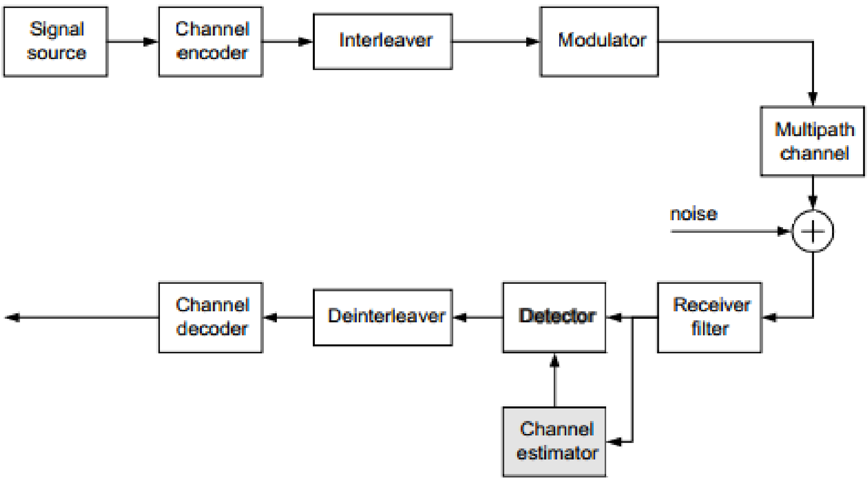


Fig. 1. Block diagram for a system utilizing channel estimator and detection

Due to the multipath channel there is some inter-symbol interference (ISI) in the received signal. Therefore a signal detector (like MLSE or MAP) needs to know channel impulse response (CIR) characteristics to ensure successful equalization (removal of ISI). Note that equalization without separate channel estimation (e.g., with linear, decision-feedback, blind equalizers [2]) is also possible, but not discussed in our research. After detection the signal is de-inter leaved and channel decoded to extract the original message.

In this paper we are mainly interested in the channel estimation part. Usually CIR is estimated based on the known training sequence, which is transmitted in every transmission burst. The receiver can utilize the known training bits and the corresponding received samples for estimating CIR typically for each burst separately. There are a few different approaches of channel estimation, like Least-squares (LS) or Linear Minimum Mean Squared Error (LMMSE) methods [3, 4].

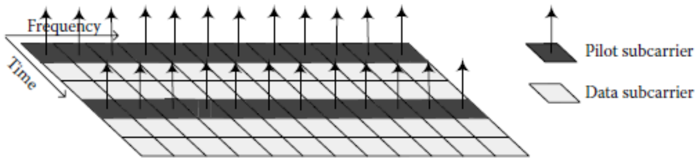
II. Pilot Structure in 3GPP

When pilot subcarriers are allocated in the time-frequency domain of an OFDMA system, different possibilities exist, as depicted in following strategy:

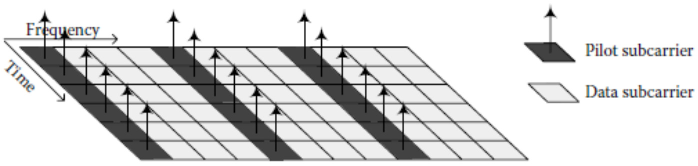
– the first strategy (Figure 2, a) — namely, block type pilot arrangement — consists of allocating a complete OFDM symbol for pilot transmission. Thus, an estimation of the channel frequency response can be obtained by least squares (LS), minimum mean square error (MMSE), or similar methods. This approach is a suitable strategy for slow time-varying channels;

– an alternative is comb type pilot arrangement, shown in Figure 2, b. In this technique, pilot subcarriers are transmitted only in some frequencies, but continuously in time. For the rest of the frequencies, the channel frequency response is obtained by linear (Lopez-Martınez et al. 2007, 987), second-order (Hsieh and Wei 1998, 217), or other interpolation methods;

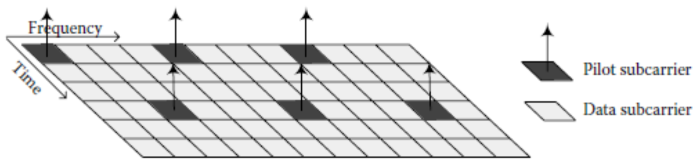
– the last strategy (Figure 2, c) is a combination of the former methods, where pilot subcarriers are spaced in both time and frequency In general, the time spacing between symbols with pilots is given by channel time variability that is, the coherence time of the channel. On the other hand, frequency spacing between pilot subcarriers is determined by the frequency selectivity of the channel, which can be quantified by its coherence bandwidth.



(a) Frequency continuous–time spaced pilot allocation (Block type)



(b) Time continuous–frequency spaced pilot allocation (Comb type)



(c) Time spaced–frequency spaced pilot allocation

Fig. 2. Pilot arrangement strategies in OFDM

Channel Estimation in OFDMA DL

In LTE downlink, cell-specific downlink reference signals are transmitted on one or several antenna ports of the eNodeB in order to facilitate the demodulation of the OFDMA signal at the UE. The structure of this reference signal corresponds to a time-frequency spaced pilot arrangement scheme (see Figure 2, c). Thus, two-dimensional time and frequency interpolation is required at the receiver. For MIMO transmission, the time-frequency allocation of reference signals on each antenna is targeted to avoid interfering with each other. Therefore, MIMO channel estimation may be addressed as the estimation of MxN individual channels, where M is the number of transmit antennas and N is the number of receive antennas. This way, channel estimation in MIMO can be tackled as a combination of different single-input, single-output (SISO) estimations (Figure 3).

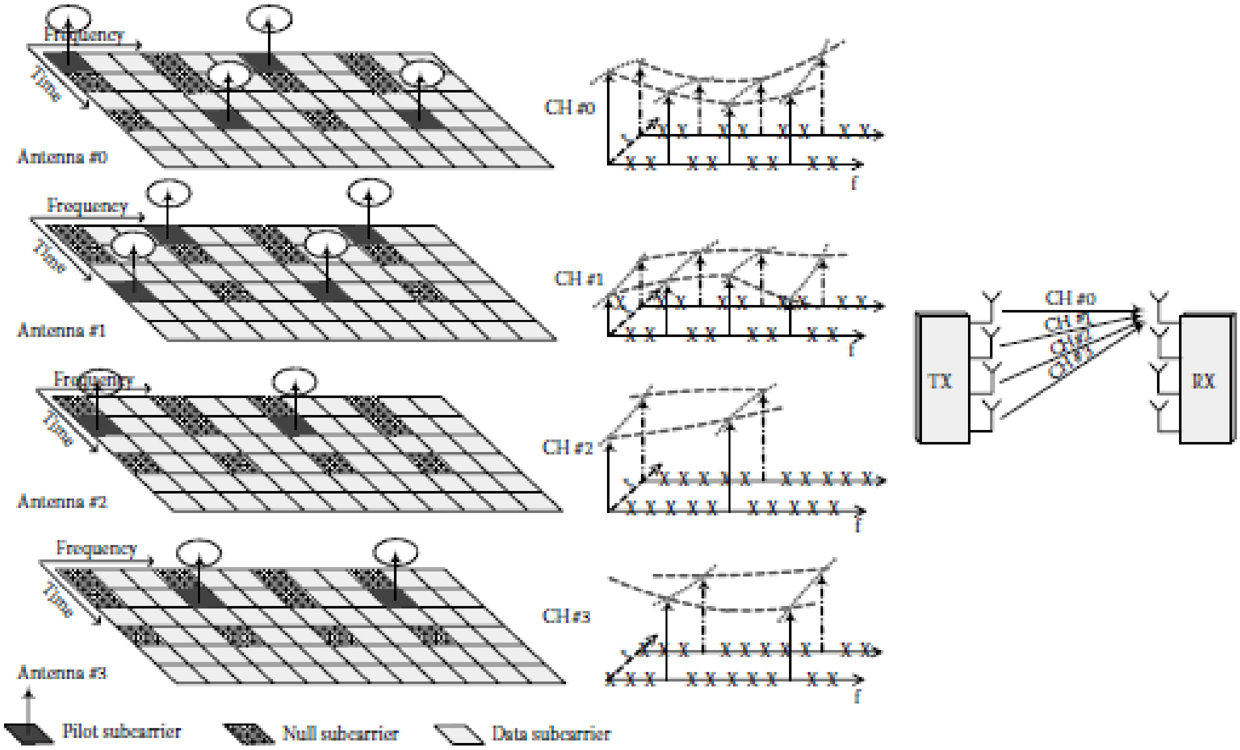


Fig. 3. Channel estimation in DL-OFDMA from MIMO reference signal (4x4 case)

Channel Estimation in SC-FDMA UL

In the case of LTE uplink, reference signals are transmitted with two different objectives. The first is to facilitate channel estimation at the eNodeB prior to demodulation (demodulation reference signal). The second, the sounding reference signal (SRS), helps to provide information of the uplink channel response in a range of frequencies greater than the allocated bandwidth for a particular UE. This allows the eNodeB to implement uplink frequency-dependent scheduling strategies. Both variants of the uplink reference signal are based on Zadoff–Chu sequences (3GPP 36.211). The demodulation reference signal is transmitted in the fourth SC-FDMA symbol of the slot and its size is equal to the uplink allocated bandwidth for that UE. Demodulation reference signal structure can be easily identified with a block type pilot arrangement scheme (see Figure 2), so classical channel estimation mechanisms for this strategy may be used. Also, time interpolation is not necessary to perform because the allocated resources for the UE are likely to be frequency hopped by the eNodeB uplink scheduler. The SRS is transmitted within a bandwidth greater than that allocated for this UE. This way, the eNodeB can have out-of-band information of the channel frequency response for this individual user, which facilitates the implementation of channel-aware scheduling mechanisms in the uplink. The transmission of an SRS may interfere with the PUSCH of other users*; hence, a broadcast parameter is defined to indicate the possible location of SRS sub-frames so that the UEs can know where to puncture their PUSCH transmission. On the other hand, an SRS is not allowed to interfere with PUCCH. The parameters that define the SRS are bandwidth, duration, periodicity, location in the sub-frame, cyclic shift for the Zadoff–Chu sequence, and transmission comb.

III. Pilot-Assisted Channel Estimation

The pilot assisted channel estimation process consists of two steps; first statistical estimation of the channel at OFDM tones consisting of reference symbols is determined using statistical methods including Squares (LS) and Minimum Mean Squares (MMSE) estimates. Different pilots assisted channel estimation schemes can be employed for the estimation of the channel effects on the transmitted signal. The response of the channel at the data sub-carriers is subsequently determined by interpolation. The interpolators used for the purpose of estimation are linear, second order, cubic or time domain interpolators derived from both the statistical and deterministic point of view [5-7]. Various publications can be found that deal with one or all these estimation criteria for pilot assisted channel estimation of OFDM applications from CIR or CFR prospective [8-13].

IV. Least Square Estimation

Least Square based parameter estimation approach aims at determining the channel impulse response from the known transmitted reference symbols in the following way:

$$G_{LS} = \left[\frac{Y_{r(1)}}{X_{r(1)}} \quad \frac{Y_{r(2)}}{X_{r(2)}} \quad \frac{Y_{r(3)}}{X_{r(3)}} \quad \frac{Y_{r(N)}}{X_{r(N)}} \right], \tag{1}$$

where $G_{LS} \in C^{N_r}$ is the estimated channel frequency response on the subcarriers which contains reference symbols? This response can be interpolated over full frequency range in order to obtain the channel frequency response for the subcarriers carrying data symbols. The interpolation can be performed in time domain or frequency domain.

The signal received in time domain can be expressed as follows:

$$Y_r = F^H A_r F_L h + \mu \,. \tag{2}$$

The channel can be estimated using Least Squares, in time domain in the following way [14]:

$$\hat{h} = \left(S^H S\right)^{-1} S^H Y_r \,. \tag{3}$$

Solving the above two equations, we get the expression for LS estimates.

$$\hat{h} \approx \left(F_L^H A_r^H A_r F_L\right)^{-1} F_L^H A_r^H F^H Y_r \,. \tag{4}$$

The term $\left(F_L^H A_r^H A_r F_L\right)^{-1} F_L^H$ in this equation is constant and can be computed offline regardless of the time varying nature of the channel. This makes LS estimator computationally simple. However an ill-conditioned problem occurs in straightway application of LS estimator due to matrix inversion. The problem can be solved in the following two ways [5].

Regularized LS Estimation

In this method, small constant term is added to the diagonal entries for regularizing the Eigen values of the matrix to be inverted. In this case, the channel impulse response becomes of the form,

$$\hat{h}_{reg} \approx \left(\alpha I + F_L^H A_r^H A_r F_L\right)^{-1} F_L^H A_r^H F^H Y_r \,. \tag{5}$$

The value of α has to be selected such that the inverse matrix is least perturbed.

Minimum Mean Square Estimation

In the above section, LS channel estimation method has been described which is computationally simple but its performance is not good. Another method to estimate the CIR is minimum mean square estimator (MMSE) which has better performance than LS but it is computationally complex.

This method intends at the minimization of the mean square error between the exact and estimated CIRs. In this section we will discuss linear minimum mean square estimator (LMMSE). The CIR can be calculated using LMMSE in the following way [14]

$$\hat{h} = R_{hYr} R_{Y_r Y_r}^{-1} Y_r \,. \tag{6}$$

Here R_{YrYr} is the auto covariance of vector Y_r and R_{hYr} is the cross covariance of vectors h and Y_r. These covariance matrices for the above equation can be calculated as

$$R_{YrYr} = E\left[Y_r Y_r^H\right], \tag{7}$$

and

$$R_{hYr} = E\left[h Y_r^H\right]. \tag{8}$$

After simplified gate

$$\hat{h} = X_r^H T_r^H \left(X_r T_r R_{hh} X_r^H T_r^H\right)^{-1} Y_r \,. \tag{9}$$

V. Frequency-Domain Channel Estimation

For frequency-domain channel estimation, first an estimate of the CTF at the pilot subcarriers is obtained. Then the full channel transfer function is calculated using an interpolation method:

– estimate at pilot position: Let $p_i, i = 0,\dots,N_p - 1$ be a set of indexes containing the subcarrier indexes that carry pilot symbols, where N_p – is the indexes of pilot symbols in the short block. A least-squares (LS) estimate of the CTF at these pilot positions can be calculated as

$$\tilde{h}_\rho[i] = \frac{z[p_i]}{d[p_i]}, \tag{10}$$

where $z[p_i]$ and $d[p_i]$ are, respectively, the received short block after the FFT and the transmitted short block in the i^{th} pilot subcarrier. h_p is an N_p long column vector containing the LS estimates of the channel at pilot subcarriers;

– wiener filtering interpolation: Wiener filtering is the optimum interpolation method in terms of MSE. Using the statistics of the channel and noise, it performs an MMSE interpolation of the estimates at pilot subcarriers, optimally reducing the effects of noise and channel distortion. The full CTF is obtained by

$$\tilde{h} = R_{hh_p} \cdot \left(R_{h_p h_p} + \sigma_w^2 I_{N_p}\right)^{-1} h_p, \tag{11}$$

where R_{hh_p} is the cross-correlation matrix of the true CTF and the true CTF at pilot subcarriers, $R_{h_p h_p}$ is the autocorrelation matrix of the true CTF at pilot subcarriers, σ_w^2 is the noise power, and I_{N_p} is the $N_p \times N_p$ identity matrix. Note that the CTF coefficients are assumed to be uncorrelated to the noise process. In another way the channel frequency response estimate is then formed by calculating the DFT of the impulse response estimate

$$\hat{H} = DFT\{\hat{h}\}. \tag{12}$$

VI. Time-Domain Channel Estimation

The channel estimation can also be performed using the time domain approach, before DFT processing of the training symbols. In this case, the channel impulse response, instead of the channel frequency response, is estimated. The received time domain signal during the two long training symbols is

$$y_{1,j} = h \cdot X_j + n_{1,j}. \tag{13}$$

Time domain convolution can be expressed as a matrix vector multiplication. The circular convolution matrix is formed from the training data as

$$X = \begin{bmatrix} x_1 & x_{64} & \cdots & x_{64-M+2} \\ x_2 & x_1 & \cdots & x_{64-M+3} \\ \vdots & \vdots & \ddots & \vdots \\ x_{64} & x_{63} & \cdots & x_{64-M+1} \end{bmatrix}, \tag{14}$$

where parameter M defines the maximum length of the impulse response that can be estimated, and X is in general a rectangular matrix. The channel impulse response vector is

$$h = [h_1 \ldots\ldots\ldots h_{M-1} \quad h_M]^T . \tag{15}$$

Then the convolution is expressed in equation (13) became

$$y_{1,j} = h \cdot X_j + n_{1,j} = Xh + n_i . \tag{16}$$

Here we can see the channel impulse response estimate can then be formed by

$$\hat{h} = h + 1/2X^{+}(n_1 + n_2), \tag{17}$$

where X^{+} denotes Moore-Penrose generalized inverse of X.

VII. Module And Parameters For Simulation

With the transport format and resource combination (TFRC) information from the link adaptation (LA) module to the transport block generator which clearly in Fig. 4. The transport block generator generates the packet based on the requested size. Then the cyclic redundancy check (CRC)-based error-detecting code is added to the corresponding packet. According to the packet size and the maximum code block size, the packet is segmented into several blocks. Each block is coded; the coding used for this study is the turbo coding defined in UTRA. The rate matching block adjusts the output bit rate according to the requirements by puncturing or repeating the coded bits. The rate matching implementation follows the description in UTRA. Afterward, the output coded bits are interleaved and are passed to the modulator. The modulator is implemented based on the supported modulation schemes defined in technical specification TR 25.814 V7.1.0 [15]: *p*/2 BPSK, QPSK, 8 PSK, and 16 QAM. For QAM modulation, a QAM remapping is performed to set the systematic bits at more reliable constellation points because this improves the decoder performance. The parameters that used in our simulation clearly in table (1) below. Single-carrier transmission is formed by the DFT spreading. The DFT spreading block precodes the PSK/QAM data symbols. Thus, all constellation symbols are mixed together to form an SC-FDMA symbol. Afterward, several SC-FDMA symbols are combined with pilot symbols in a time division multiplex (TDM) and mapped to the proper subcarriers. The 3GPP study item report [15] specifies both localized and distributed allocation mapping. The data symbols are placed on orthogonal subcarriers by the IFFT and the cyclic prefix is added. At the receiver side, the base station basically performs the reverse operations with respect to the process in the transmitter. The amplitude and phase variation due to the transmission in a frequency-selective fading are compensated by an equalizer. The equalizer is basically a one-tap frequency domain equalizer located at the output of the FFT in the receiver.

Table 1. Simulation parameters

Parameter	Value
Carrier frequency	2 GHz
Transmission BW	10 MHz
Subframe duration	0.5 ms
Subcarrier spacing	15 kHz
SC-FDM symbols/TTI	6 long blocks, 2 short blocks
CP duration	4.1 μs
FFT size/useful subcarriers	1024/600
MCS setting	π/2 BPSK:1/3, 1/6 QPSK: 1/2 , 2/3 ,3/4 8 PSK: 1/2 , 1/3, 2/3, ¾ 16 QAM: 1/2, 2/3, 3/4, 4/5
Channel code	3Gpp rel. 6 compliant turbo code with basic rate 1/3
Rate matching, inter-leaver	3Gpp rel. 6 compliant
Channel estimation	Ideal
Antenna schemes	SISO
Channel models	Typical urban 6 paths
Speed	120 km/h and 200 km/h

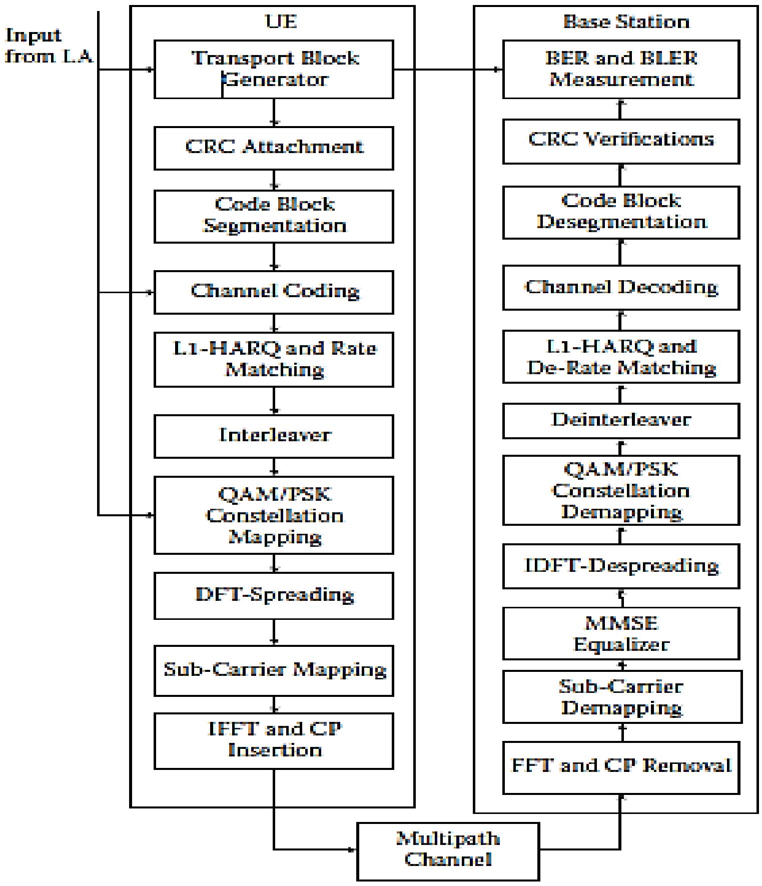


Fig. 4. Link –level simulator SC-FDMA and OFDMA

VIII. Simulation System

A simple time-domain channel estimation based on linear interpolation is assumed. In a subframe, two short blocks are transmitted. The channel estimation for each subframe is a linear interpolation between these two short blocks. Fig. 4 shows the link-level simulation results with the preceding channel estimation techniques using UE speed at 120 km/h 200 km/h. Fig 5 It is shown that the OFDMA performance is more affected by the UE speed than SC-FDMA, especially at higher modulation and coding schemes (16 QAM rate 4/5). The UE at high speed has a higher Doppler frequency and creates intercarrier interference (ICI). In OFDM, each subcarrier carries different symbols and therefore OFDM is more sensitive to the ICI and it may happen from frequency offset resulting in Inter Carrier Interference (ICI) while receiving an OFDM modulated symbol. Let us analyze the effect of frequency offset and later we will define the loss of orthogonality and resulting signal to noise ratio (SNR) loss due to the presence of frequency offset or from high speed. The analysis is accompanied by Matlab/Octave simulation scripts which clearly in Fig 6 below here we used BPSK modulation to become clearly.

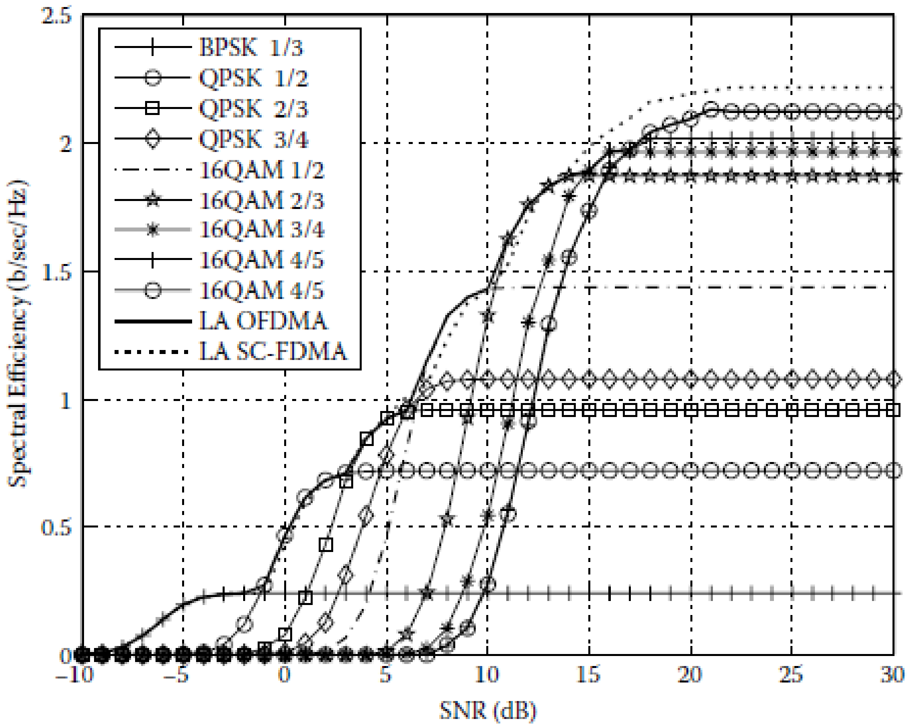


Fig. 5. SC-FDMA and OFDMA with real channel estimation at a speed of 120 km/h

Fig. 7 illustrates the results for a speed of 200 km/h. It is shown that 16 QAM rates of 3/4 and above do not contribute to the link adaptation curve and hence the system operates at low spectral efficiency. At high speed, the subcarrier orthogonality is destroyed by the ICI.

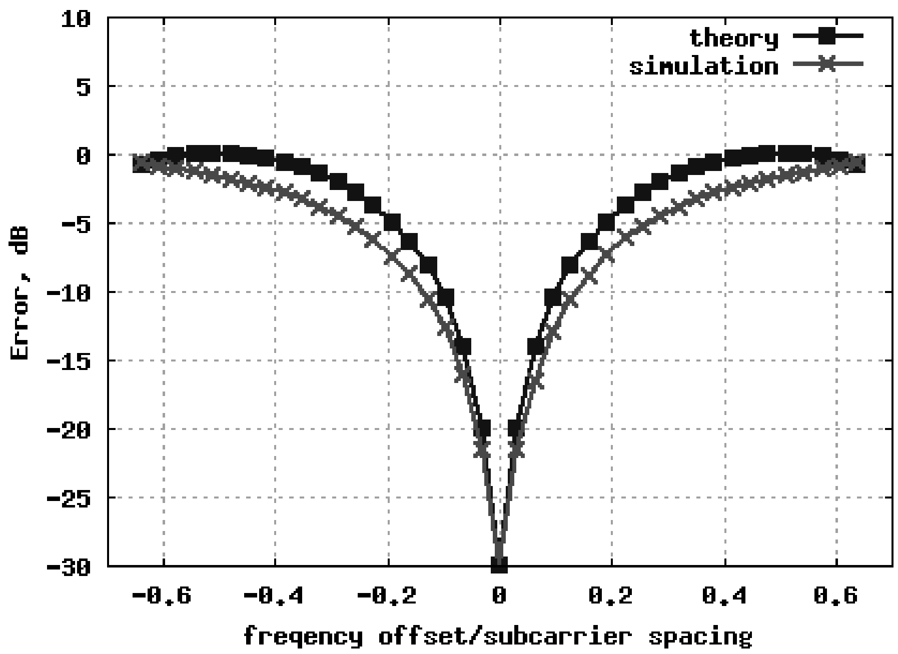


Fig. 6. Error Magnitude vs frequency offset for OFDMA

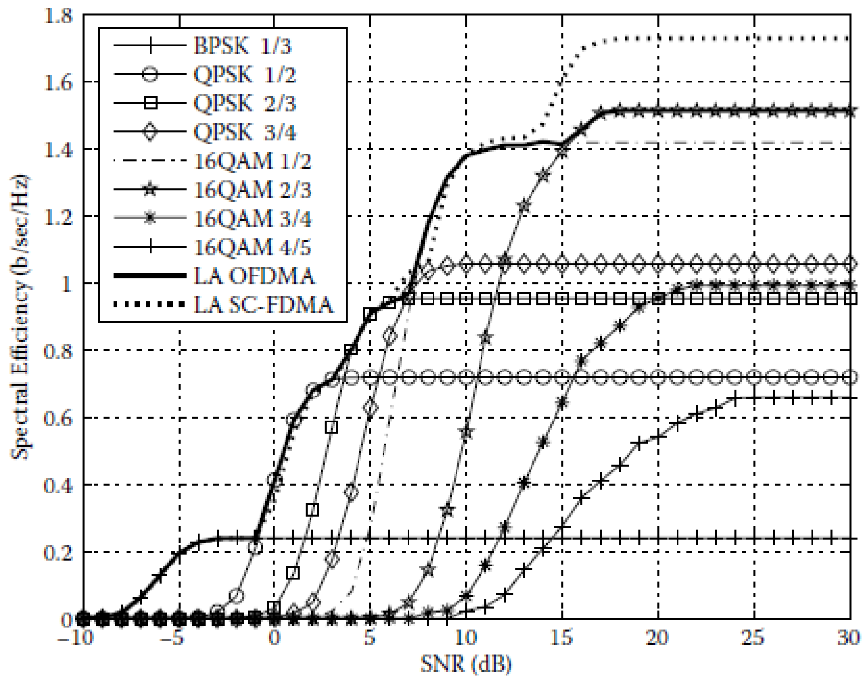


Fig. 7. SC-FDMA and OFDMA with real channel estimation at a speed of 200 km/h

Conclusion

This paper presents a channel estimation, which depicts that type of modulation not improves the spectral efficiency when speed became rise. It means that ICI will increase according to our simulation in fig. 5 and fig. 7 and throw these figures we have studied effected OFDMA on User speed more them SC-FDMA specially when we have higher order of modulation. We presented many configuration antennas and here presented MMSE to increase data rate, which led to increase channel estimation and reduce ISI it's not clear here because we concentrate on mobility of user and this will clear in next paper. And we can see many reasons effect on channel estimation one of this reason offset frequency with appear in fig. 5 lead to ICI.

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