A Practical Performance Analysis of CRS-Aided Channel Estimation Algorithms for LTE Downlink System

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Abstract. Channel estimation algorithms are employed in 3GPP Long Term Evolution (LTE) downlink system to help with coherent demodulation. Several CRS-aided channel estimation algorithms over multipath Rayleigh fading channel have been investigated in this paper. Based on theoretical analysis and simulation, Wiener interpolation channel estimator is proposed for LTE downlink system. In consideration of practical implementation and universality for different channels, we propose that CRSs in Wiener interpolation should be contained in the time window of $10e^{-3}$ second as well as in the frequency window of two adjacent resource blocks, which symmetrically distribute around the current estimated resource block in frequency domain.

Keywords: 3GPP LTE, channel estimation, CRS-aided channel estimation, Wiener interpolation.

1 Introduction

The third generation partnership project (3GPP) long term evolution (LTE), which is favored by most telecommunication operators all over the world, has been generally recognized as the internationally powerful mobile communication system [1].The LTE targets at significantly increased instantaneous peak data rates, higher average spectrum efficiency and the cell-edge user throughput efficiency [2]. Some advanced technologies which are new to cellular applications are employed by 3GPP LTE physical layer, e.g. multiple Input Multiple Output (MIMO) and Orthogonal Frequency Division Multiplexing (OFDM) are adopted.

MIMO technology as well as adaptive technology is used in LTE to enhance the data rate and system performance, which makes the peak data rate reach over 100Mbit/s in the downlink (DL). Multiple antennas provide with an additional degree of freedom to the channel scheduler. Depending on user's selection over individual resource blocks (RBs) in the spatial domain, different MIMO schemes could be used in the 3GPP standards.

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In the physical layer of 3GPP LTE DL system, OFDM has been used due to its high bandwidth efficiency and the capability to mitigate the inner symbol interference (ISI) in a severe multi-path fading channel. 3GPP LTE uses orthogonal frequency division multiplexing access (OFDMA) scheme for transmission in DL and single carrier frequency division multiple access (SCFDMA) for uplink. A dynamic channel estimation algorithm tracking of fading channel at the receiver is necessary before demodulation of OFDM symbols, because the radio channel is mostly frequency-selective and time-varying for wideband mobile communication systems [3]. In OFDM symbols, cell-specific reference signals (CRS) are used to do help with channel estimation. The receiver could estimate the whole channel response of each OFDM symbol by processing the received signals at predefined positions of CRS. Such kind of CRS-aided channel estimation has been proven as a feasible method for OFDM systems. CRS-aided channel estimation algorithms can be based on Least Square (LS) or Minimum Mean Square Error (MMSE) [4]. LS is relatively simple, which dose not need relevant channel information and is easily affected by noise. As for MMSE, channel response is estimated with the help of channel statistic information and relevance between subcarriers.

Although many papers about channel estimation have been published, to the best of our knowledge, there are few methods based on numerous practical elements. Moreover, there is no detailed performance evaluation on how to select CRS. In this paper, Wiener filter interpolation with different configurations and linear interpolation have been investigated for 3GPP LTE DL. Analysis of channel estimation performance has been done on all the antenna ports. Based on statistic property of different channels and performance v.s. complexity ratio, an appropriate channel estimation algorithm has been selected.

The rest of the paper is organized as follows: 3GPP DL system model is introduced in Section 2, and some channel estimation algorithms are presented in Section 3. Scenario, parameters description and the analysis of channel estimation performance are shown in Section 4, followed by Section 5 which concludes this paper.

2 System Model in LTE DL

2.1 CRS Structure

In LTE DL, CRS sequence $r_{l,n_s}(m)$ shall be mapped to complex-valued modulation symbols $a_{k,l}^{(p)}$ used as reference symbols for antenna port p in slot n_s, where $r_{l,n_s}(m)$, $a_{k,l}^{(p)}$ and the mapping function between them are defined in Section 6.10.1 in [5]. As shown in figure 1, the white grid represents resource elements (REs) for data transmission, and the colored grid means REs for CRSs. The shaded grid denotes unused RE, because REs used for reference signal transmission on any of the antenna ports in a slot shall not be used for any transmission on any other antenna port in the same slot and set to zero. Based on the definition in [5], the m of CRS is decided by the length of cyclic prefix, cell ID and antenna port. And the density of CRS on antenna port 2 (or 3) is half as much as it on antenna port 0 (or 1).

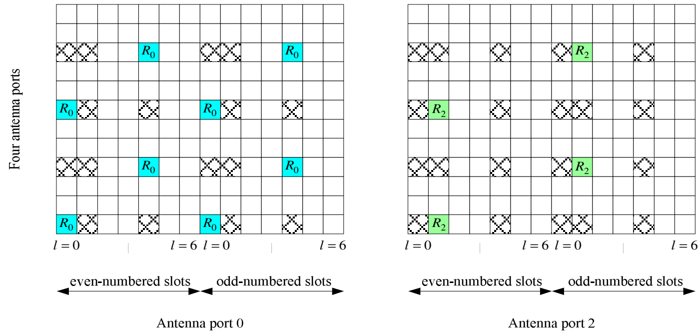

Fig. 1. Mapping of DL reference signals (normal cyclic prefix) [5]

2.2 OFDM in LTE

As shown in figure 2, $x(k)$ is the input data at the transmitter and $y(k)$ is the output data at the receiver. $w(i)$ is complex additive white Gaussian noise(AWGN). At the transmitter, a modulated sequence $x(k)$ is transformed into a N-point sequence $x(n)$, where N is the length of IFFT. A cyclic prefix for avoiding Inter-symbol interference is added before transmitting. It is an inverse process at the receiver, and channel estimation is done after FFT in receiver.

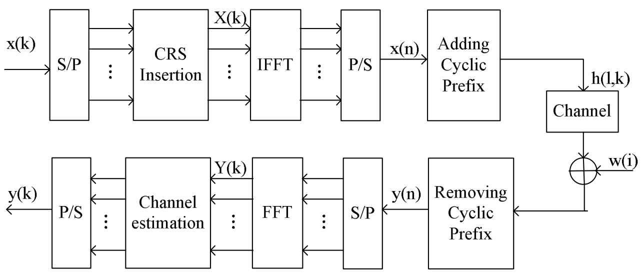

Fig. 2. The block diagram of OFDM transmitter and receiver

2.3 Frame Structure

There are two radio frame structures supported in 3GPP LTE. One is applicable to both full duplex and half duplex FDD, which is shown in figure 3. The other is applicable to TDD. In [5], seven kinds of uplink-downlink configurations for TDD frames are also presented.

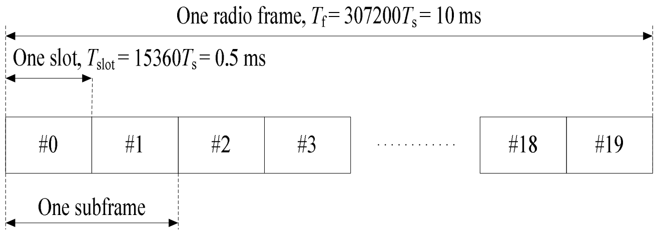

Fig. 3. FDD frame structure [5]

3 Channel Estimation Algorithms

A number of channel estimation techniques have been proposed to estimate channel response in LTE DL, such as LS, MMSE and maximum likelihood (ML). In all the CRS-aided techniques, channel response on CRS is estimated first based on the above principles. And two-dimension Wiener interpolation [6] or linear interpolation is performed to estimated channel response on data REs based on the known channel response on CRSs.

3.1 Two-Dimension Wiener Interpolation

The time-variable channel impulse response could be written as:

$$h(t,\tau)=\sum_{l=0}^{L-1}\alpha_l(t)\delta(\tau-\tau_l) \ . \tag{1}$$

Then, the frequency response of the channel at time t is written as:

$$H(t,f)=\int_{-\infty}^{\infty}h(t,\tau)e^{-j2\pi f\tau}d\tau=\sum_{l=0}^{L-1}\alpha_l(t)e^{-j2\pi f\tau_l} \ . \tag{2}$$

Assume that $\alpha_l(t)$ has the following correlation function in time domain:

$$r_{\alpha_l}(t+\triangle t,t)=E\left\{\alpha_l(t+\triangle t)\cdot\alpha_l^*(t)\right\}=\sigma_l^2 r_t(\triangle t) \ . \tag{3}$$

Assume that different paths are independent, and then the correlation function in frequency domain written as:

$$\begin{aligned} r_H(\triangle t,\triangle f) &= E\left\{H(t+\triangle t,f+\triangle f)\cdot H^*(t,f)\right\} \\ &= \sum_{l=0}^{L-1} r_{\alpha_l}(\triangle t)e^{-j2\pi\triangle f\tau_l} = r_t(\triangle t)\sum_{l=0}^{L-1}\sigma_l^2 e^{-j2\pi\triangle f\tau_l} \ . \end{aligned} \tag{4}$$

If the channel is normalized, that is

$$\sum_{l=0}^{L-1}\sigma_l^2 = 1 \ . \tag{5}$$

And $r_t(\triangle t)$ could be computed by the zero-order Bessel function of 1st kind [7], f_d is the Doppler frequency spread.

$$r_t(\triangle t) = J_0(2\pi f_d \triangle t). \tag{6}$$

Assume that the propagation channel do not vary during the duration of one OFDM symbol. After sampling, channel response in frequency domain could be written as:

$$H(n,k) = H\left(nT, k\frac{1}{NT_s}\right) = \sum_{l=0}^{L-1}\alpha_l(nT)e^{-j2\pi\frac{d_l \cdot k}{N}} \ . \tag{7}$$

$$r_H(n,k) = r_H\left(nT, k\frac{1}{NT_s}\right) == r_t(nT)\sum_{l=0}^{L-1}\sigma_l^2 e^{-j2\pi\frac{d_l \cdot k}{N}} \ . \tag{8}$$

where $d_l = \tau_l / T_s$, T_s is the sampling time, and T is the duration of an OFDM symbol (including cyclic prefix). What's more, n means the interval of two REs in time domain; k means the interval of two REs in frequency domain.

$$T = (N_{CP} + N) \cdot T_s \ . \tag{9}$$

where N_{CP} is the length of cyclic prefix, N is the length of IFFT/FFT.

Assume that the duration of cyclic prefix is larger than impulse response of channel and there is no inter-symbol interference. After the CRSs are faded by the channel, the corresponding received signal could be written as:

$$\vec{Y} = \vec{H} + \vec{W} = \begin{pmatrix} H(n_1,k_1) \\ H(n_2,k_2) \\ \vdots \\ H(n_P,k_P) \end{pmatrix} + \vec{W} \ . \tag{10}$$

All the vectors in (10) are $P \times 1$ vectors, $\vec{Y}$ could correspond to any combination of CRSs. n_p and k_p are the indices of CRSs in time and frequency domain, respectively, where $p = 0,1,\cdots,P$. $\vec{W}$ is the vector of additive white Gaussian noise, the variance of each of which is σ_w^2.

Based on the time-frequency two-dimensional Wiener, channel response in frequency of any RE could be written as:

$$\hat{H}(n,k) = \vec{C}(n,k)\vec{Y} \quad . \tag{11}$$

where,

$$\vec{C}(n,k) = \vec{r}_{H(n,k)\vec{Y}} \cdot \vec{r}_{\vec{Y}\vec{Y}}^{-1} \quad . \tag{12}$$

$$\begin{aligned} \vec{r}_{H(k,n)\vec{Y}} &= E\left\{H(k,n)\cdot\vec{Y}^H\right\} \\ &= E\left\{H(k,n)\cdot\left(H^*(n_1,k_1),\quad H^*(n_2,k_2),\quad \cdots \quad H^*(n_P,k_P)\right)\right\} \\ &= \left(r_H(n-n_1,k-k_1),\quad r_H(n-n_2,k-k_2),\quad \cdots \quad r_H(n-n_P,k-k_P)\right) \end{aligned} \quad . \tag{13}$$

$$\begin{aligned} \vec{r}_{\vec{Y}\vec{Y}} &= E\left\{\vec{Y}\vec{Y}^H\right\} = E\left\{\vec{H}\vec{H}^H\right\} + \sigma_n^2\vec{I} \\ &= \begin{bmatrix} r_H(0,0) & r_H(n_1-n_2,k_1-k_2) & \cdots & r_H(n_1-n_P,k_1-k_P) \\ r_H^*(n_1-n_2,k_1-k_2) & r_H(0,0) & \cdots & \vdots \\ \vdots & \vdots & \ddots & \vdots \\ r_H^*(n_1-n_P,k_1-k_P) & \cdots & \cdots & r_H(0,0) \end{bmatrix} + \sigma_n^2\vec{I} \end{aligned} \quad . \tag{14}$$

3.2 Linear Interpolation

The basic assumption of linear interpolation is that the relationship of channel response between estimated REs and known REs is linear. Then, channel response of RE could be calculated base on the slope. Assume that there are two known REs, which can be denoted as (x_1, y_1), (x_2, y_2), and the unknown RE can be denoted as (x, y) accordingly. Then the estimated channel response of RE can be written as:

$$y = \frac{x - x_1}{x_2 - x_1} y_2 + \frac{x_2 - x}{x_2 - x_1} y_1 . \tag{15}$$

Two-dimension linear interpolation can be divided into two one-dimension linear interpolation. Generally, one-dimension linear interpolation in time (or frequency) domain is done firstly, and then the other one-dimension linear interpolation in frequency (or time) domain is done. One-dimensional interpolation in time domain can be done as follows:

$$\hat{H}(n+m,k) = (1 - \frac{m}{\Delta})H(n,k) + \frac{m}{\Delta}H(n+\Delta,k) \quad . \tag{16}$$

where $0 \le m \le \Delta$ and Δ is the duration between two known REs.

4 Performance and Complexity Analysis with Different Configuration

A cellular OFDM downlink system over multipath Rayleigh fading environment is considered to compare the performance of the above estimators. In simulation, channel types are pedestrian A 3km/hour, vehicular A 30km/hour and vehicular A120km/hour are selected, which are defined in Rec. ITU-R M.1225 [8] and denoted as PedA, VehA30 and VehA120 in the following context, respectively.

4.1 Parameters Description

The performance of channel estimators is evaluated at bandwidth of 5MHz, 10MHz and 20MHz. The performance of channel estimators at 20MHz is shown and analyzed. The performance at 5MHz and 10MHz is similar with it at 20MHz.

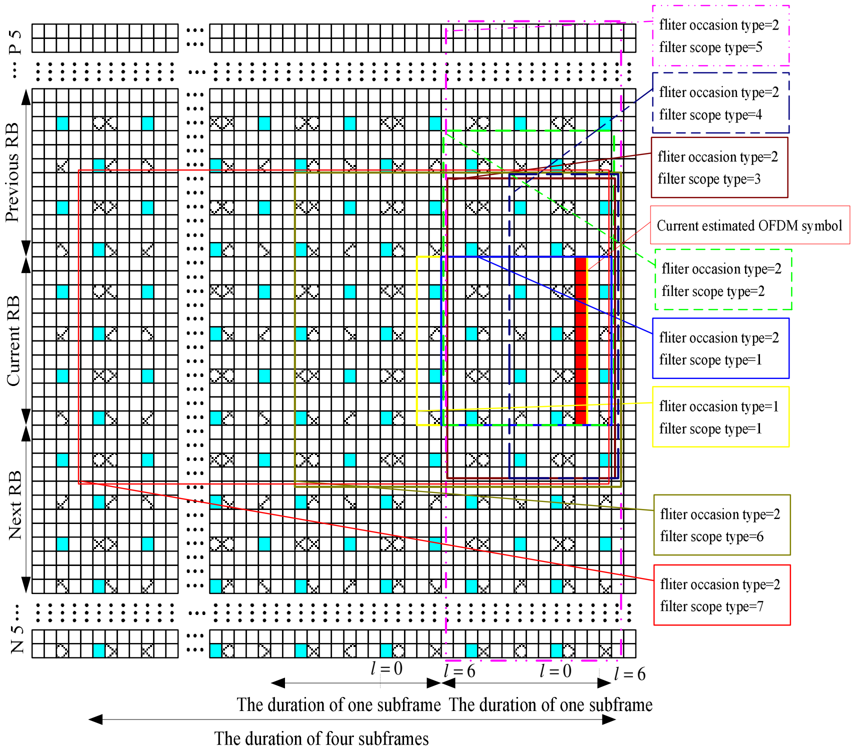

Fig. 4. Various types of configurations in Wiener interpolation on port 0

As shown in (11), the performance of 2D Wiener interpolation is determined by $\bar{C}(n,k)$, which is determined by selected CRSs. Thus the performance and complexity of 2D Wiener interpolation are fully decided by pattern of the selected CRSs. In order to choose a better interpolation method with appropriate configuration, antenna port 0 is taken as an example, and figure 4 shows different CRSs pattern in 2D Wiener interpolation.

Wherein “filter occasion type” indicates when channel estimation is done in time domain. It is used in both linear and Wiener interpolation. “filter occasion type = 1” represents that channel estimation (CE) for an OFDM symbol is done as soon as it is received. And “2” represents that CE is done for the previous 2 or 3 OFDM symbols in the time domain on Port 0 when a new OFDM symbol with CRS is received. If it is Port 2, “2” represents that CE is done for the previous 6 OFDM symbols in the time domain when a new OFDM symbol with CRS is received. “filter scope type” is only used in Wiener interpolation, which indicates CRS used in CE. Moreover, “the previous (or next) XX RBs” means the previous (or next) XX RBs in frequency domain. “filter scope type = 1” represents that CRSs in current RB in 10^{-3} second are used, which is equal to the duration of one subframe. “2” represents that CRSs in previous and current RB in 10^{-3} second are used for CE. “3” represents that CRSs in half of previous and next RBs, as well as the current RB in 10^{-3} second are used. “4” represents that CRSs which are closer to current OFDM symbol from CRSs in “3” are chosen to be used. “5” represents that CRSs in previous five and next five RBs in 10^{-3} second are used, as well as current RB. “6” represents that CRSs in half of previous and next RBs, as well as the current RB in 2×10^{-3} second are used, which is equal to the duration of two subframes. “7” represents that CRSs in half of previous and next RBs, as well as the current RB in 4×10^{-3} second are used. And for VehA120, 4×10^{-3} second is approximate to its coherence time.

4.2 Channel Statistic Information

The coherence time of the three channel types are different from each other, which is shown in table 1.

Table 1. Coherence Time over Three Types of Channel

	PedA	VehA30	VehA120
Coherence time (s)	0.1714	0.0171	0.0043

4.3 Performance of Channel Estimation on Port 0

As shown in figure 5, the performance of Wiener channel estimator becomes better as the growth of CRSs, which is used in Wiener interpolation. When MSE is 10^{-1}, the required SNR can be decreased about 10 dB from “filter scope type=1” to “filter scope type=7” over PedA. Meanwhile, figure 6 and 7 show the similar tendency. Yet, the required SNR is decreased about 5 dB when MSE is 10^{-1} over VehA120, because part of selected CRSs in Wiener interpolation with “filter scope=7” approach to the limit of channel coherence time.

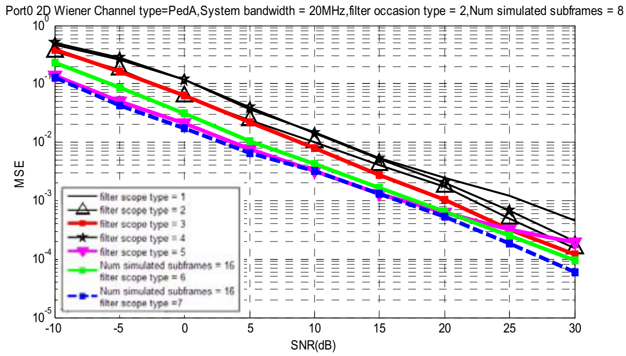

Fig. 5. The performance of Wiener channel estimator over PedA on port 0

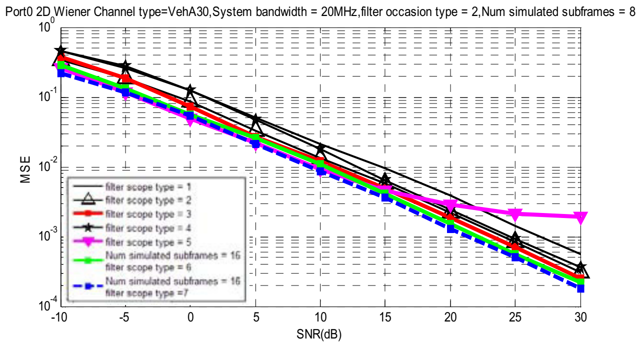

Fig. 6. The performance of Wiener channel estimator over VehA30 on port 0

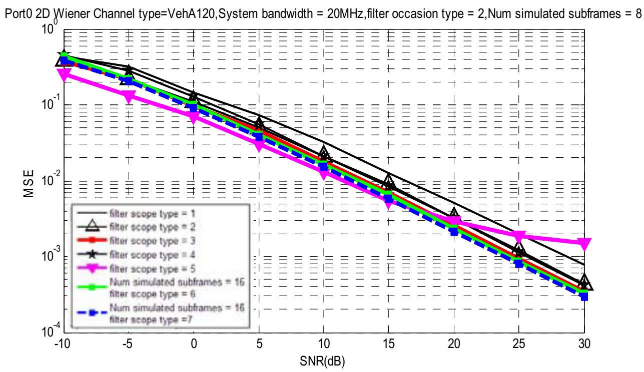

Fig. 7. The performance of Wiener channel estimator over VehA120 on port 0

As for figure 6 and figure 7, the performance of Wiener channel estimator does not become better when filter scope is 5 over VehA30 and VehA120, because the selected CRSs in “filter scope=5” are from more than ten RBs in frequency domain, which exceeds the coherence bandwidth. Based on statistic properties of channel, the coherence bandwidth is only about 398kHz over VehA30 and VehA120, which is approximate to two RBs in frequency domain. Nevertheless, Wiener estimator with "filter scope=5" uses much uncorrelated CRSs, the effect of which is equal to noise. Thus there is a MSE bound when SNR is high. When the duration of CRS is 10^{-3} second, the performance of Wiener channel estimator with "filter scope=3" is better than others over VehA30 and VehA120.

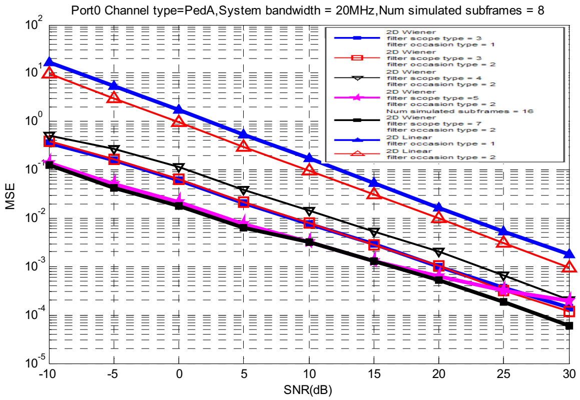

Fig. 8. The performance of channel estimation over PedA on port 0

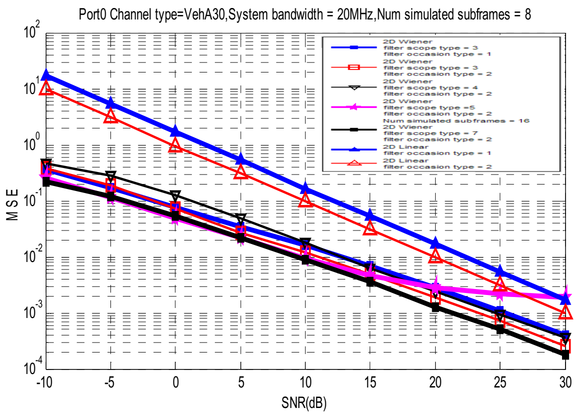

Fig. 9. The performance of channel estimation over VehA30 on port 0

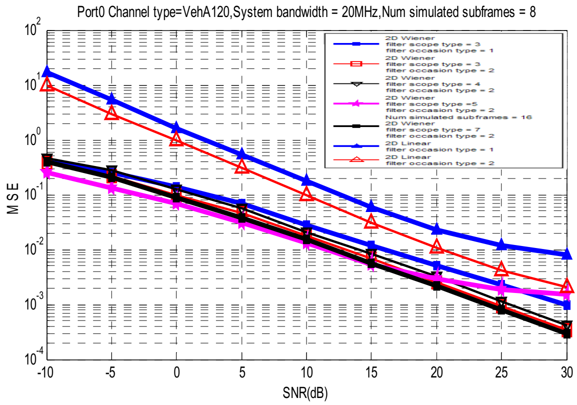

Fig. 10. The performance of channel estimation over VehA120 on port 0

Figure 8-10 show the performance of linear channel estimator and Wiener channel estimator over different channels. The performance becomes better as the growth of SNR in both linear channel estimator and Wiener channel estimator. Yet the performance of Wiener channel estimator is always better than linear channel estimator, especially when SNR is low. When MSE is 10^{-1} over PedA, the required SNR in linear channel estimator (filter occasion =1) is about 12.30 dB, while the required SNR in Wiener channel estimator (filter occasion=1, filter scope =3) is -2.65 dB. With the growth of velocity, Doppler frequency shift becomes larger. Accordingly, coherence time becomes smaller. Coinciding with this, Figure 8-10 above show that the performance of channel estimation becomes worse as the velocity grows. Moreover, the performance of Wiener estimator with “filter occasion=2” is much better than that of Wiener estimator with “filter occasion=1”, especially when users’ velocity is high. Table 2 shows the value of SNR over different channels when MSE is 10^{-1}, and table *3* shows the value of SNR over different channels when MSE is 10^{-2}. Wherein, the parameters of Wiener estimator consist of filter scope type and filter occasion type, respectively. And the parameter of linear estimator is filter occasion type.

Table 2. The Value of SNR (dB) When MSE is 10^{-1}

Channel Type	2D-Wiener (s=3,o=1)	2D-Wiener (s=3,o=2)	2D-Wiener (s=7,o=2)	2D-Linear (o=1)	2D-Linear (o=2)
PedA	-2.65	-2.45	-8.90	12.30	9.75
VehA30	-1.65	-1.75	-4.00	12.20	9.78
VehA120	2.35	-0.15	-0.62	12.63	9.83

Table 3. The Value of SNR (dB) When MSE is 10^{-2}

Channel Type	2D-Wiener (s=3,o=1)	2D-Wiener (s=3,o=2)	2D-Wiener (s=7,o=2)	2D-Linear (o=1)	2D-Linear (o=2)
PedA	8.55	8.90	2.80	22.20	19.90
VehA30	12.82	11.30	9.20	22.20	19.87
VehA120	16.1	13.10	12.17	27.20	20.30

4.4 Performance of Channel Estimation on Port 2

As it is defined in [5], the density of CRS on port 0 is twice of CRS on port2. Thus the performance of CRS-aided channel estimation on port 0 is better than it on port 2, although the same channel estimator is used.

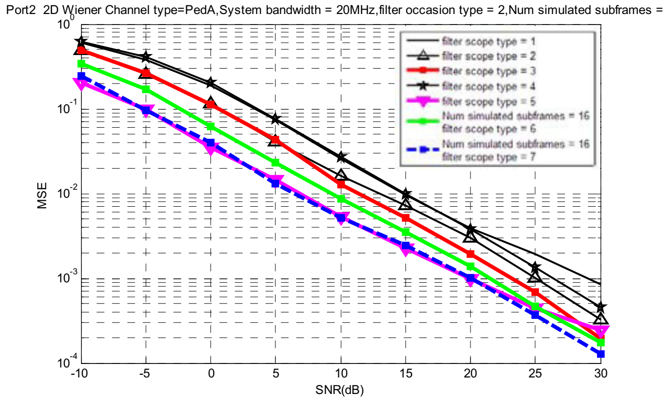

Fig. 11. The performance of Wiener channel estimator over PedA on port 2

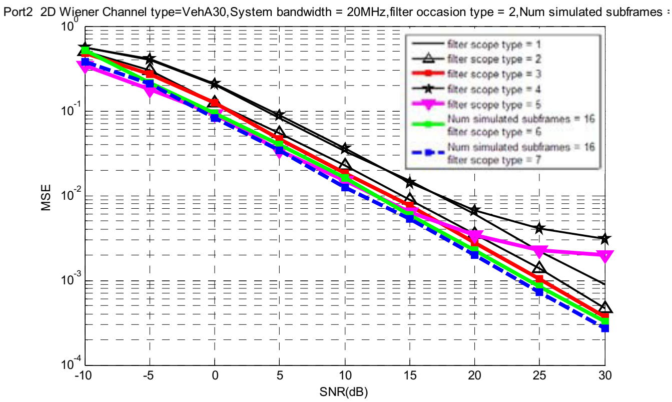

Fig. 12. The performance of Wiener channel estimator over VehA30 on port 2

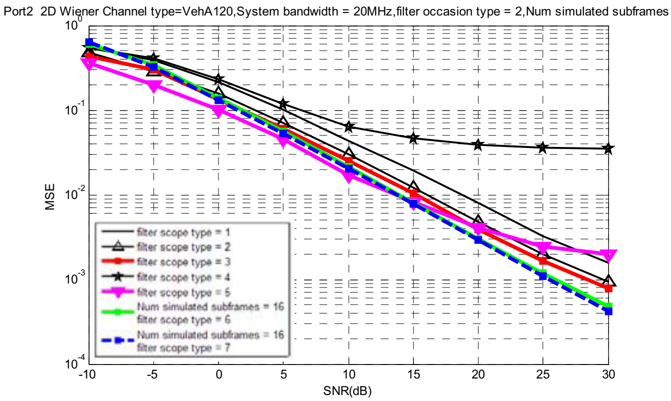

Fig. 13. The performance of Wiener channel estimator over VehA120 on port 2

Figure 11-13 show that the performance of Wiener estimator becomes better with the growth of CRSs, no matter in frequency domain or in time domain. There is also a MSE bound over VehA30 and VehA120 when filter scope is 5, because part of selected CRSs used in Wiener estimator are not in coherence bandwidth. And when the duration of CRS is 10^{-3} second, the performance of Wiener channel estimator with "filter scope=3" is better than others over VehA30 and VehA120.

Table 4. The Value of SNR (dB) When MSE is 10^{-1}

Channel Type	2D-Wiener (s=3,o=1)	2D-Wiener (s=3,o=2)	2D-Wiener (s=7,o=2)	2D-Linear (o=1)	2D-Linear (o=2)
PedA	0.15	0.60	-5.17	12.53	9.85
VehA30	1.58	1.10	-1.10	12.55	9.78
VehA120	5.77	1.85	1.50	>30.00	10.40

Table 5. The Value of SNR (dB) When MSE is 10^{-2}

Channel Type	2D-Wiener (s=3,o=1)	2D-Wiener (s=3,o=2)	2D-Wiener (s=7,o=2)	2D-Linear (o=1)	2D-Linear (o=2)
PedA	11.65	11.40	6.45	22.55	19.70
VehA30	16.45	13.40	11.25	22.65	19.70
VehA120	21.70	15.17	13.70	>30.00	>30.00

The contrast between linear estimator and Wiener estimator is shown in table 4, when MSE is 10^{-1}. The required SNR in linear channel estimator (filter occasion =1) is about 12.53 dB, while the require SNR in Wiener channel estimator (filter occasion=1, filter scope =3) is 0.15 dB, when MSE is 10^{-1} over PedA. Yet the performance of channel estimator is worse than it on port 0, no matter linear estimator or Wiener estimator. And when MSE is 10^{-2}, the contrast is shown in table 5. Wherein, the parameters of Wiener estimator consist of filter scope type and filter occasion type, respectively. And the parameter of linear estimator is filter occasion type.

5 Conclusion

In this paper, two-dimension linear interpolation and two-dimension Wiener interpolation channel estimation methods for LTE DL have been investigated over multipath Rayleigh fading environment. Based on theoretical analysis and simulation, we propose that channel estimation should be done for the previous 2 or 3 OFDM symbols in the time domain on port 0 when a new OFDM symbol with CRS is received. If it is on port 2, channel estimation should be done for the previous 6 OFDM symbols in the time domain when a new OFDM symbol with CRS is received. In view of performance, computational complexity and universality for different channels, we propose that Wiener interpolation channel estimator should be used, which uses CRSs in half of previous and next RB, as well as the current RB in 10^{-3} second (one subframe). Wherein, "the previous and next RB" means the previous and next RB in frequency domain. In time domain, the CRSs in the time window of 10^{-3} second should be used.

References

1. Hou, J., Liu, J.: A novel channel estimation algorithm for 3GPP LTE downlink system using joint time-frequency two-dimensional iterative Wiener filter. In: IEEE International Conference on Communication Technology (November 2010)
2. 3GPP TR 36.814 V0.4.1, Further Advancements for E-UTRA Physical Layer Aspects (Release 9) (February 2009)
3. Bahai, A.R.S., Saltzberg, B.R.: Multi-Carrier Digital Communication: Theory and Applications of OFDM. Kluwer Academic/Plenum (1999)
4. van de Beek, J.-J., Edfors, O., Sandell, M., Wilson, S.K., Borjesson, P.O.: On channel estimation in OFDM systems. In: IEEE 45th Vehicular Technology Conference, pp. 815–819 (1995)
5. 3GPP TS 36.211 V9.1.0, Technical Specification Group Radio Access Network for E-UTRA Physical channels and Modulation (Release 9) (March 2010)
6. Qin, Y., Hui, B., Chang, K.H.: Performance evaluation of pilot-based channel estimation and equalization techniques in 3GPP LTE downlink. In: Proc. Of KICS Summer Conference (November 2009)
7. Zhou, W., Lam, W.H., A novel method of Doppler shift estimation for OFDM systems. In: IEEE Military Communication Conference (November 2008)
8. Recommendation ITU-R M.1225, Guidelines for Evaluation of Radio Transmission Technologies for IMT-2000(Question ITU-R 39/8) (1997)