Doppler Spread Estimation for Broadband Wireless OFDM Systems

Jun Tao, Jingxian Wu, and Chengshan Xiao

Abstract— **In this paper, we present a new Doppler spread estimation algorithm for broadband wireless orthogonal frequency division multiplexing (OFDM) systems with time-varying and frequency-selective Rayleigh fading. The algorithm is developed by analyzing the statistical properties of the power of received signals in the time domain, thus it excludes the influence of intercarrier interference introduced by channel variation within one OFDM symbol. The operation of the algorithm doesn't require the knowledge of fading coefficients, transmitted data symbols, or signal-to-noise ratio (SNR). It works well under time-selective and frequency-selective Rayleigh fading channel with SNR as low as 0 dB. Moreover, unlike existing algorithms, the proposed algorithm takes into considerations of the discrete-time channel inter-tap correlation, as the case in practical systems. Simulation results demonstrate that this new algorithm can accurately estimate a wide range of Doppler spread with low estimation latency and high computational efficiency.**

I. Introduction

Doppler spread estimation or mobile speed estimation has a wide range of applications for single carrier systems [1] – [3], and has received extensive attentions [4] – [6]. However, few algorithms have been discussed on Doppler spread estimation for orthogonal frequency division multiplexing (OFDM) systems [7], [8]. Doppler spread estimation is important for broadband wireless OFDM system [9], where time selectivity of channel can not be ignored even within one OFDM symbol [10]. In [7], auto-covariance of post-DFT (Discrete Fourier Transform) frequency-domain signal is utilized to estimate Doppler spread or mobile speed. The algorithm can only be applied to signals on pilot tones. Consequently, to extract the necessary statistics for Doppler spread estimation, large number of OFDM blocks are required, and this results in long estimation delay. In [8], autocorrelation of time-domain channel estimations is utilized to estimate Doppler spread. However, the wireless receiver still has to know the fading channel coefficients before estimating the Doppler spread. Also, this method provides satisfactory estimation accuracy only when the signal-to-noise ratio (SNR) is high.

As mentioned above, most existing algorithms require the knowledge of channel fading coefficients before Doppler estimation can be performed. This results in high computational complexities at the receiver. In addition, previous estimation algorithms designed for frequency selective fading assume a symbol-spaced discrete-time channel model with uncorrelated channel taps, which, as shown in [11] and [12], is a very special case of practical systems. The uncorrelated channel tap assumption may significantly limit the practical value of those algorithms.

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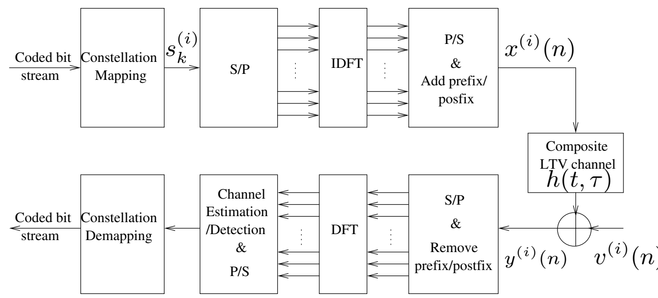

Fig. 1. OFDM system model (only $y^{(i)}(n)$ is used during Doppler spread estimation.)

In this paper, a novel algorithm, which does not suffer from any aforementioned limitation, is proposed to estimate Doppler spread for broadband OFDM system under doubly selective (time-selective and frequency-selective) Rayleigh fading. Compared to other algorithms in the literature, the new algorithm has three main advantages. First, the estimation statistics are collected from all the signals at the receiver, and it doesn't require the knowledge of fading coefficients or transmitted data symbols. So the limitations in [7], [8] are avoided. Second, the new algorithm is not affected by intercarrier interference (ICI) and is robust over a wide range of SNR. Third, the algorithm is developed based on practical discrete-time channel model with correlated channel taps. Numerical simulations demonstrate that it can provide very good estimation accuracy even for SNR as low as 0 dB.

II. System Model and Assumptions

Consider an OFDM system shown in Fig. 1. For the ith ($i \in \mathbb{Z}$) OFDM symbol, a set of N modulated symbols, $s^{(i)} = [s_0^{(i)}, ..., s_{N-1}^{(i)}] \in \mathcal{C}^{1\times N}$, are multiplexed onto N subcarriers. The modulated complex symbols, $s_k^{(i)}$, are assumed to be zero-mean random variables with correlation given by

$$E\left\{ s_{k_1}^{(i)} \left[s_{k_2}^{(j)} \right]^* \right\} = \delta(k_1 - k_2)\delta(i-j). \tag{1}$$

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This assumption is valid for modulation schemes such as M-ary phase shift keying (MPSK), M-ary quadrature amplitude modulation (MQAM), etc.

Performing N-point normalized inverse discrete Fourier transform (IDFT) [10] leads to the time-domain samples, $x^{(i)}(n)$, as

$$x^{(i)}(n) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} s_k^{(i)} e^{j\frac{2\pi}{N}kn}, -N_p \leq n < N + N_q, \quad (2)$$

where N_p and N_q are the lengths of cyclic prefix and cyclic postfix, respectively. Cyclic postfix is introduced to avoid intersymbol interference (ISI) caused by the noncausal parts of the equivalent channel.

The time domain signals are passed through transmit filter, and then serially transmitted over wireless channel. Let $h(t,\tau) = p_T(\tau) \otimes g_c(t,\tau) \otimes p_R(\tau)$ denote the composite impulse response (CIR) of the channel, where $\otimes$ represents convolution, $p_T(\tau)$ and $p_R(\tau)$ are the impulse responses of the transmit filter and receive filter, respectively, and $g_c(t,\tau)$ is the time-varying impulse response of the physical channel. The physical channel $g_c(t,\tau)$ is assumed to be wide-sense stationary uncorrelated scattering (WSSUS) Rayleigh distributed [13].

At the receiver, the output of the receive filter is sampled at rate $1/T_s$. After the prefix and postfix are discarded, the time domain samples at the output of the receive filter can be represented by [12]

$$y^{(i)}(n) = \sum_{l=-L_1}^{L_2} h^{(i)}(n,l)x^{(i)}(n-l) + v^{(i)}(n), \quad \text{for } n = 0, 1, \cdots, N-1, \quad (3)$$

where $y^{(i)}(n) \triangleq y((iN_{sym}+n)T_s)$ is the nth sample of the ith received OFDM symbol, with $N_{sym} = N_p + N + N_q$ being the number of samples contained in one OFDM symbol including prefix and postfix, $v^{(i)}(n)$ is the additive white Gaussian noise (AWGN) sample with variance σ^2, and $h^{(i)}(n,l) \triangleq h((iN_{sym}+n)T_s, lT_s)$ is the discrete-time version of the continuous-time CIR $h(t,\tau)$. Due to the insertion of cyclic prefix and postfix, the signal sample, $y^{(i)}(n)$, contains information contributed only from the ith transmitted OFDM symbol. The non-negative integers, L_1 and L_2, define the tap index range of the discrete-time CIR, $h^{(i)}(n,l)$, for $l = -L_1, \cdots, L_2$. The values of L_1 and L_2 depend on the transmit filter, $p_T(\tau)$, receive filter, $p_R(\tau)$, and power delay profile of the physical channel [12].

The CIR, $h(t,\tau)$ or $h^{(i)}(n,l)$, is generally non-causal if $p_T(\tau)$ and $p_R(\tau)$ have zero delay. It should be noted that the underlying physical channel is always causal. In addition, the presence of transmit filter and receive filter introduces inter-tap correlation in the delay domain l or τ, even though the physical fading is WSSUS. For Rayleigh fading, the discrete-time CIR, $h^{(i)}(n,l)$, is a wide-sense stationary zero mean Gaussian distributed random process with auto-correlation given by [12]

$$E\{h^{(i_1)}(n_1,l_1)h^{(i_2)*}(n_2,l_2)\} = C_{l_1,l_2} J_0\{2\pi f_d[(i_1-i_2)N_{sym} + (n_1-n_2)]T_s\}, \quad (4)$$

where C_{l_1,l_2} is correlation coefficient between the l_1th channel tap and the l_2th channel tap, $J_0(\cdot)$ is the zero-order Bessel function of the first kind, f_d is maximum Doppler spread, and $E(\cdot)$ and $(\cdot)^*$ stand for mathematical expectation and complex conjugate, respectively. The value of C_{l_1,l_2} is determined by transmit filter, channel power delay profile, and receive filter. For energy normalized Rayleigh fading channel, $\sum_{l=-L_1}^{L_2} C_{l,l} = 1$.

III. Development of Doppler Spread Estimation Algorithm

In this section, the theory for Doppler spread estimation is developed by analyzing the second-order and fourth-order statistics of the received signal $y^{(i)}(n)$.

A. *Statistics of Received Signals*

Define the auto-correlation of the received signal, $y^{(i)}(n)$, and the auto-correlation of received signal power, $|y^{(i)}(n)|^2$, as

$$R_{yy}(s,u,n,m) \triangleq E\left\{y^{(s)}(n+m)\left[y^{(u)}(n)\right]^*\right\}, \quad (5a)$$

$$R_{|y|^2|y|^2}(s,u,n,m) \triangleq E\left\{|y^{(s)}(n+m)|^2|y^{(u)}(n)|^2\right\}, \quad (5b)$$

where $s, u \in \mathbb{Z}$ are OFDM symbol indices, and n, m are time index and time lag in samples, respectively.

Substituting (2), (3) into (5), we have the auto-correlations expressed as

$$R_{yy}(s,u,n,m) = \left\{\frac{1}{N}\sum_{l_1=-L_1}^{L_2}\sum_{l_2=-L_1}^{L_2}\sum_{k=0}^{N-1} C_{l_1,l_2}J_0(2\pi f_d m T_s) \exp[j2\pi(l_2-l_1+m)k/N]\right\}\delta(s-u) + \sigma^2\delta(m)\delta(s-u), \quad (6)$$

$$\begin{aligned} R_{|y|^2|y|^2}(s,u,n,m) &= 1 + 2\sigma^2 + \sigma^4 + 2\sigma^2\delta(m)\delta(s-u) + \sigma^4\delta(m)\delta(s-u) + \\ & J_0^2\{2\pi f_d[(s-u)N_{sym}+m]T_s\} \sum_{l_1=-L_1}^{L_2}\sum_{l_2=-L_1}^{L_2} |C_{l_1,l_2}|^2 + \\ & \frac{1}{N^2}\sum_{l_1=-L_1}^{L_2}\sum_{l_2=-L_1}^{L_2}\sum_{l_3=-L_1}^{L_2}\sum_{l_4=-L_1}^{L_2}\sum_{\substack{k_1=0 \\ k_1\neq k_2}}^{N-1}\sum_{k_2=0}^{N-1} \left[C_{l_1,l_2}C_{l_3,l_4} + C_{l_1,l_4}C^*_{l_2,l_3}J_0^2(2\pi f_d m T_s)\right] \times \\ & \exp\left\{j\frac{2\pi}{N}[k_1(m-l_1+l_4) - k_2(m+l_3-l_2)]\right\}\delta(s-u). \quad (7) \end{aligned}$$

It should be noted that the identity, $J_0\{2\pi f_d[(s-u)N_{sym} + m]T_s\}\delta(s-u) = J_0(2\pi f_d m T_s)\delta(s-u)$, is used in the derivation of (6) and (7). The detailed derivations of (6) and (7) are omitted here for brevity. Observing (6) and (7) reveals that both $y(n)$ and $|y(n)|^2$ are wide-sense stationary – they are functions of symbol index difference $s-u$ and time difference m, while independent of the starting symbol index s and starting time n. We denote $R_{yy}(s,u,n,m) \triangleq R_{yy}(s-u,m)$ and $R_{|y|^2|y|^2}(s,u,n,m) \triangleq R_{|y|^2|y|^2}(s-u,m)$ in the sequel unless otherwise specified.

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Setting $s-u=0, m=0$ in (6), we have $R_{yy}(0) \triangleq R_{yy}(s=u,n,m=0)$

$$R_{yy}(0) = \frac{1}{N}\sum_{l_1=-L_1}^{L_2}\sum_{l_2=-L_1}^{L_2} C_{l_1,l_2}\sum_{k=0}^{N-1} e^{j2\pi(l_2-l_1)k/N} + \sigma^2, \\ = 1+\sigma^2. \quad (8)$$

Therefore, the auto-covariance of the received signal power, $V_{|y|^2|y|^2}(s-u,m) = E[|y^{(s)}(n+m)|^2|y^{(u)}(n)|^2] - E[|y^{(s)}(n+m)|^2]E[|y^{(u)}(n)|^2]$, can be written as

$$V_{|y|^2|y|^2}(s-u,m) = R_{|y|^2|y|^2}(s-u,m) - (1+\sigma^2)^2. \quad (9)$$

The statistics described in (6), (7) and (9) are expressed as functions of the Doppler spread f_d. However, the mathematical expressions of the statistics are extremely complicated, thus it would not be easy to directly extract f_d from them. To facilitate the development of the Doppler spread estimation algorithm, we offer three remarks of the statistics.

Remark 1: From (6), received signals coming from two different OFDM symbol intervals ($s \neq u$) are always mutually uncorrelated regardless of m. Within the same OFDM symbol interval, when the time lag m meets $L < m < N-L$, where $L = L_1 + L_2$, the autocorrelation of received signal is also zero. Therefore, the autocorrelation, $R_{yy}(s-u,m)$, is not a good candidate for Doppler spread estimation.

Remark 2: When $s \neq u$, (7) and (9) can be simplified as follows

$$R_{|y|^2|y|^2}(s-u,m) = 1+2\sigma^2+\sigma^4+V_{|y|^2|y|^2}(s-u,m) \quad (10)$$

$$V_{|y|^2|y|^2}(s-u,m) = \\ J_0^2\{2\pi f_d[(s-u)N_{sym}+m]T_s\}\sum_{l_1=-L_1}^{L_2}\sum_{l_2=-L_1}^{L_2}|C_{l_1,l_2}|^2. \quad (11)$$

In (11), the effects of additive white Gaussian noise (AWGN) is removed.

Remark 3: When $s = u$, the expression of the auto-correlation or auto-covariance has a complicated extra term compared to the $s \neq u$ case, as can be seen from (7) or (9). This extra term is undesirable for Doppler spread estimation.

From the analysis above, we conclude that the auto-covariance of the signal power evaluated at $s \neq u$ case is the most suitable candidate for Doppler spread estimation. Compared to other statistics, $V_{|y|^2|y|^2}(s-u,m)$ at $s \neq u$ has the following two advantages. First, it is directly related to maximum Doppler spread in an expression easy to analyze. Second, the effect of AWGN is completely removed, and this is highly desirable for the design of a robust algorithm.

B. Doppler Spread Estimation

We are now in a position to develop the Doppler spread estimation algorithm based on the auto-covariance given in (11). Denote

$$\alpha = \sum_{l_1=-L_1}^{L_2}\sum_{l_2=-L_1}^{L_2}|C_{l_1,l_2}|^2 \quad (12)$$

and let $z = 2\pi f_d[(s-u)N_{sym}+m]T_s$, then (11) can be written in a compact form as

$$V_{|y|^2|y|^2}(z) = \alpha J_0^2(z). \quad (13)$$

It's difficult to estimate Doppler spread f_d from (13) directly due to unknown variable α. Instead, we resort to the normalized auto-covariance defined as

$$V_N(k,z) \triangleq \frac{V_{|y|^2|y|^2}(kz)}{V_{|y|^2|y|^2}(z)} = \frac{J_0^2(kz)}{J_0^2(z)} \ \ (k>1 \ \text{ integer}). \quad (14)$$

When the integer k is properly chosen such that kz is small, second-order approximation $J_0(x) \cong 1 - x^2/4$ can be applied to (14), which results in

$$\frac{V_{|y|^2|y|^2}(kz)}{V_{|y|^2|y|^2}(z)} \cong \frac{[1-(kz)^2/4]^2}{(1-z^2/4)^2}. \quad (15)$$

From (15), z is solved as

$$z = 2\sqrt{\frac{1-R_N(k,z)}{k^2-R_N(k,z)}}, \quad (16)$$

where $R_N(k,z) = \sqrt{V_N(k,z)}$. Let $s-u=1$ and $m=0$, then $z = 2\pi f_d N_{sym} T_s$. The estimation for Doppler spread is finally given by

$$\tilde{f}_d = \frac{\sqrt{\frac{1-R_N(k,z)}{k^2-R_N(k,z)}}}{\pi N_{sym}T_s} = \frac{F_1(k,z)}{\pi N_{sym}T_s}, \quad (17)$$

where

$$F_1(k,z) \triangleq \sqrt{\frac{1-R_N(k,z)}{k^2-R_N(k,z)}}. \quad (18)$$

It is worth pointing out that other choices of s, u, and m may also be used during estimation. In this paper, however, we fix $z = 2\pi f_d N_{sym} T_s$ without loss of generality.

Similarly, if we utilize fourth-order approximation: $J_0(x) \cong 1 - x^2/4 + x^4/64$, then

$$\frac{V_{|y|^2|y|^2}(kz)}{V_{|y|^2|y|^2}(z)} \cong \frac{[1-(kz)^2/4+(kz)^4/64]^2}{(1-z^2/4+z^4/64)^2}, \quad (19)$$

and z is solved as

$$z = 2\sqrt{2}\sqrt{\frac{k^2-R_N(k,z)-(k^2-1)\sqrt{R_N(k,z)}}{k^4-R_N(k,z)}}. \quad (20)$$

From (20) and the fact that $z = 2\pi f_d N_{sym} T_s$, Doppler spread estimation using fourth-order approximation is

$$\hat{f}_d = \frac{\sqrt{2}\sqrt{\frac{k^2-R_N(k,z)-(k^2-1)\sqrt{R_N(k,z)}}{k^4-R_N(k,z)}}}{\pi N_{sym}T_s} = \frac{F_2(k,z)}{\pi N_{sym}T_s}, \quad (21)$$

where

$$F_2(k,z) \triangleq \sqrt{2}\sqrt{\frac{k^2-R_N(k,z)-(k^2-1)\sqrt{R_N(k,z)}}{k^4-R_N(k,z)}}. \quad (22)$$

The mobile speed can be calculated from the Doppler spread as $v = f_d c/f_c$, where c is the speed of light and f_c is the carrier frequency.

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The newly proposed method has several advantages over existing algorithms. First, it doesn't require the estimation of fading channel coefficients. Thus, all transmitted signals, including unknown data and known pilot, can be used in the estimation. Second, effect of additive noise is removed in the estimation thanks to the auto-covariance of received signal power. Third, the new method is developed based on a practical discrete-time channel model with correlated taps. Finally, the estimation is performed over time domain signals, thus it's not affected by ICI.

IV. A Practical Doppler Spread Estimation Algorithm

Based on the theoretical analysis presented in Section III, a practical Doppler estimation algorithm with low estimation latency is presented in this section.

From (17) and (21), the proper operation of the Doppler estimation algorithm requires the knowledge of the auto-covariance function, $V_{|y|^2|y|^2}(k,z)$, which can in turn be approximated using time average. To obtain an accurate time-averaged approximation, long data sequences are needed. This will result in large estimation delay with high computational complexity. To address this problem, we pass the received signal power, $p^{(i)}(n) = |y^{(i)}(n)|^2$, through a low pass filter to suppress the effects of AWGN and power fluctuations. The output of the low pass filter can be represented by

$$\hat{p}^{(i)}(n) = \sum_{l=0}^{L_f} p^{(i)}(n-l)f(l), \tag{23}$$

where L_f is the order of low pass filter with coefficients $f(l)$.

With the output of the low pass filter, the auto-covariance approximation can be collected from M consecutive OFDM symbols as

$$V_{|y|^2|y|^2}(s-u=k,m) = \sum_{i=0}^{M-k-1}\sum_{n=0}^{N-1}[\hat{p}^{(i+k)}(n+m)-\bar{p}][\hat{p}^{(i)}(n)-\bar{p}], \tag{24}$$

where $\bar{p}$ is the mean value of the signal power calculated as

$$\bar{p} = \frac{1}{MN}\sum_{i=0}^{M-1}\sum_{n=0}^{N-1}\hat{p}^{(i)}(n). \tag{25}$$

After obtaining estimation of auto-covariance, we still have one problem left: how to choose the integer value of k used in (17) and (21)? The selection of k depends on two factors: first, kz should be small such that approximations in (15) and (19) still hold; second, for a given Doppler spread f_d, or $z = 2\pi f_d N_{sym} T_s$, the normalized auto-covariance, $V_N(k,z)$, should not be too close to 1, such that enough information can be collected from $V_N(k,z)$ for Doppler spread estimation. However, there is no single value of k that can achieve the above two objectives over all the practical Doppler spread values. Therefore, we classify the Doppler spread into three categories, "low", "medium", and "high". Different values of k are used for each of the three categories.

Classification of Doppler spread is performed at the receiver by using the auto-covariance of received signal power. In (9), when $s = u$, we choose the value of m such that $T_m = mT_s$ is less than $30\mu s$, then $J_0(2\pi f_d m T_s) \approx 1$ even if f_d is as high as 500Hz. In this case, the auto-covariance $V_{|y|^2|y|^2}(s-u=0,m)$ tends to a constant value irrelevant to f_d. We denote this constant as $V_{|y|^2|y|^2}(T_m)$. Further, we choose $s-u=k_0$, such that $T_{blk} = k_0 T_{sym} \approx 5ms$, where $T_{sym} = N_{sym}T_s$ is the time duration of one OFDM symbol including prefix and postfix. Similarly, define $V_{|y|^2|y|^2}(T_{blk}) \triangleq V_{|y|^2|y|^2}(s-u=k_0, m=0)$. Then the classification of Doppler spread can be achieved by comparing the ratio of auto-covariances with certain thresholds, as in the following

$$\frac{V_{|y|^2|y|^2}(T_{blk})}{V_{|y|^2|y|^2}(T_m)} \begin{cases} > 0.95, & low, \\ < 0.3, & high, \\ otherwise, & medium. \end{cases} \tag{26}$$

It should be noted that the threshold values are chosen for illustration purpose only. Other values can be obtained through optimization to get better estimation accuracy.

After channel classification, the value of k is chosen as

$$k = \begin{cases} 10k_0, & low, \\ k_0, & medium, \\ \lceil 0.1k_0 \rceil, & high. \end{cases} \tag{27}$$

The values used in (27) are obtained based on empirical simulation results. The values of k can then be substituted into (17) and (21) to get the estimation of Doppler spread.

V. Simulation

The DVB-H system is used as an example to illustrate the effectiveness of the proposed algorithm in practical OFDM systems.

In the simulation, one OFDM symbol excluding prefix and postfix has time duration of $T_u = 224\mu s$. Sampling interval is $T_s = T_u/N$, where the value of N is related to system operating mode. Three modes are provided in DVB-H: 2K-mode, 4K-mode and 8K-mode, corresponding to $N = 2048$, $N = 4096$ and $N = 8192$, respectively. The 2K-mode with $N = 2048$ is used in the simulation. The length of the cyclic prefix and cyclic postfix is $N/8 = 256$ samples. One OFDM frame consists 68 OFDM symbols, and four frame forms one super-frame. The transmit filter $p_T(\tau)$ and receive filters $p_R(\tau)$ are normalized square root raised cosine filter with roll-off factor 0.3. The power delay profile of the wireless physical fading channel has 120 taps with $T_s/2$ spacing between adjacent taps. The average power of the first 40 taps ramps up linearly and the last 80 taps ramps down linearly, and the total power of the fading channel is normalized to unity. Low pass filter has $L_f = 50$ taps. $k_0 = 20$ is chosen in (27) for DVB-H system. In this simulation, the maximum Doppler spread estimation is obtained by averaging over the two estimations from (17) and (21) as $\bar{f}_d = (\hat{f}_d + \tilde{f}_d)/2$.

Fig. 2 shows the simulation results of the proposed Doppler spread estimation algorithm in a DVB-H system with 64QAM modulation. One estimation is performed over $N_f = 24$ consecutive OFDM frames, which corresponds to a time duration less than 0.5s. The mean and standard deviation of the estimated Doppler spread under various SNR are plotted in the figure. It's clear from the figure that the proposed algorithm provides accurate and reliable estimation of the Doppler spread for SNR as low as 0 dB. More accurate estimations can be achieved at higher SNR.

In Fig. 3, the standard deviation of estimated Doppler spread is plotted as a function of the number of OFDM frames used in one estimation. As expected, the estimation standard deviation decreases monotonically when more OFDM frames are used during estimation. This is intuitive because more frames lead to a better time-averaged approximation of the auto-covariance.

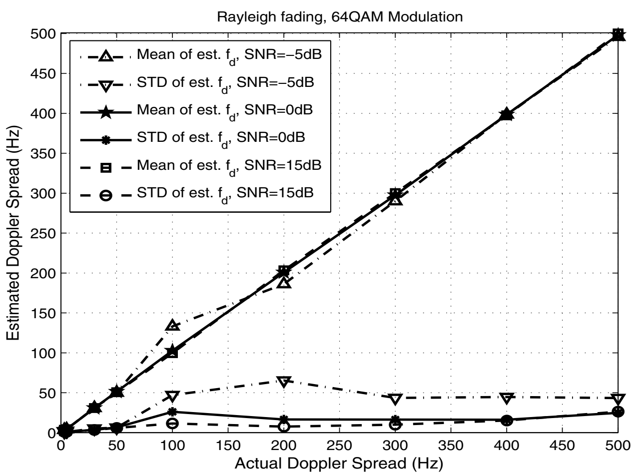

Fig. 2. Doppler spread estimation for frequency selective Rayleigh fading channel with 64QAM modulated signals.

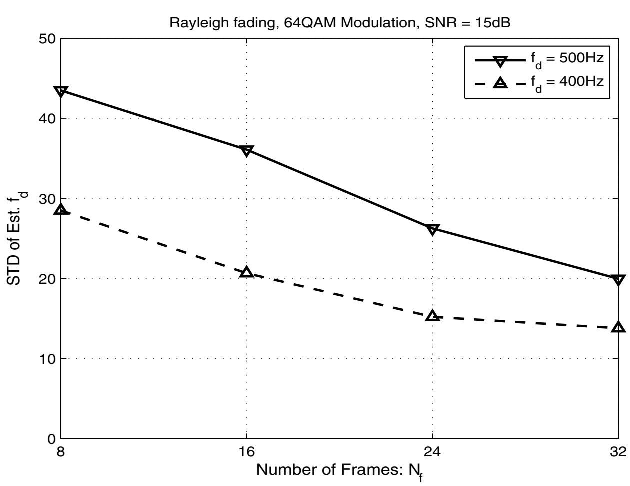

Fig. 3. Standard deviation of estimated Doppler v.s. number of OFDM frames used in estimation for Rayleigh fading channel with 64QAM modulation.

VI. Conclusion

An accurate low latency Doppler spread estimation algorithm was presented in this paper for OFDM systems with time-varying and frequency-selective Rayleigh fading. The algorithm was developed by analyzing the statistical properties of received signals containing unknown transmitted signal and unknown channel fading. The estimation is performed with the auto-covariance of the power of the Pre-DFT time domain signals at the receiver. Since the algorithm is operated in the time domain, its performance is not affected by the ICI induced by the time variation within one OFDM symbol. In addition, a practical algorithm was proposed for accurate and high efficiency Doppler estimation. Extensive simulations have shown that our new algorithm works well for doubly selective Rayleigh fading channel with SNR as low as 0 dB.

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