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Mobile Speed Estimation for Broadband Wireless Communications

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***Abstract*— In this paper, a new algorithm is proposed to estimate mobile speed for broadband wireless communications, which often encounter large number of fading channel taps causing severe intersymbol interference. Theoretical analysis is first derived and practical algorithm is proposed based on the analytical results. The algorithm employs a modified auto-covariance of received signal power to estimate the speed of mobiles. The algorithm is based on the received signals which contain unknown transmitted data, unknown frequency selective multipaths possibly including line-of-sight (LOS) component, and random receiver noise. The algorithm works well for frequency selective Rayleigh and Rician channels. The algorithm is very resistant to noise, it provides accurate speed estimation even if the signal-to-noise (SNR) is as low as 0dB. Simulation results indicate that the new algorithm is very reliable and effective to estimation mobile speed corresponding maximum Doppler up to 500Hz. The algorithm has high computational efficiency and low estimation latency, with results being available within one second after communication is established.**

I. Introduction

In wireless communication networks, the estimation of mobile user's speed plays an important role to enhance the network performance. Therefore, in the last twelve years, mobile speed estimation has received extensive attention in the literature [1]-[24]. Most existing algorithms are proposed for frequency nonselective fading channels, and developed based on channel fading coefficients. This implies that the wireless receiver has to accurately estimate the fading channel coefficients before estimating the mobile speed. This may significantly limit the value and applications of these algorithms. Moreover, many existing algorithms provide good estimation accuracy only when the signal-to-noise ratio (SNR) is high. If the SNR is moderate or low, then the speed estimation accuracy tends to be unsatisfactory. Furthermore, to our best knowledge, all the existing algorithms are unable to provide good speed estimation results for frequency selective Rician fading channels.

In this paper, a new algorithm is presented for mobile speed estimation under frequency selective Rician fading channels. The new algorithm is based only on the received signal with noise. It provides very good estimation accuracy without knowing the fading channel coefficients or transmitted data. It is not only computationally efficient but also robust to intersymbol interference and noise.

II. Channel Model and Preliminaries

A. Physical Fading Channel and Its Discrete-Time Model

Consider a SISO wireless channel shown in Fig. 1. Assume that the transmit pulse shaping filter $p_T(\tau)$, the receive matched filter $p_R(\tau)$ and the physical fading $g(t,\tau)$ are normalized with unit energy.

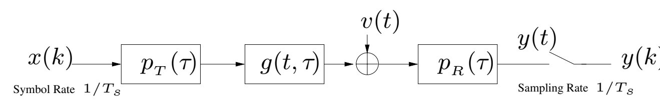

Fig. 1. The baseband block diagram of a SISO wireless fading channel.

For a broadband channel with line of sight (LOS) component, the physical fading $g_c(t,\tau)$ is frequency selective Rician fading given by

$$g_c(t,\tau) = \frac{g(t,\tau)}{\sqrt{1+K}} + \frac{K}{\sqrt{1+K}} h_{LOS}(t)\delta(\tau) \tag{1}$$

where K is the Rice factor, $g(t,\tau)$ is wide-sense stationary uncorrelated scattering (WSSUS) [25] Rayleigh fading with normalized unit energy, and the LOS component is assumed to be $h_{LOS}(t) = \exp(j2\pi f_d t \cos\theta_0 + j\phi_0)$ with f_d being the maximum Doppler frequency, θ_0 and ϕ_0 being the angle of arrival and the initial phase, respectively.

Assume that the transmit and receive filters have zero delays, the received signal $y(t)$ is sampled with timing at $\tau = 0$ and symbol interval T_s. Then the composite channel impulse response $h(t,\tau) = p_R(\tau) \otimes g_c(t,\tau) \otimes p_T(\tau)$ can be accurately converted to the following discrete-time fading channel model [26]

$$y(n) = \frac{1}{\sqrt{1+K}} \sum_{l=-L_1}^{L_2} h_l(n) x(n-l) + \frac{K}{\sqrt{1+K}} h_{LOS}(n) \sum_{l=-P_1}^{P_2} \sigma_l x(n-l) + v(n) \tag{2}$$

where $x(n)$ is the nth transmitted symbol, $v(n)$ is the additive white Gaussian noise, $h_l(n)$ is the lth tap fading channel coefficient at time instant n, L_1 and L_2 are non-negative integers, L_1+L_2+1 is the Rayleigh fading channel length which depends on the transmit filter, power delay profile and the receive filter [26], $h_{LOS}(n) = \exp(j2\pi f_d n T_s \cos\theta_0 + j\phi_0)$

is the LOS channel coefficient at time instant n, $\sigma_l = \int_{-\infty}^{\infty} p_T(s)p_R(lT_s - s)ds$ is the sampled value at $\tau = lT_s$ of the convolution between the transmit and receive filters, and $\sum_{l=-P_1}^{P_2} \sigma_l^2 = 1$, P_1 and P_2 are non-negative integers which depend on the transmit and receive filters. For practical systems, $P_1 \le L_1$ and $P_2 \le L_2$.

B. Statistics of The Discrete-time Fading Channel

It is known from (2) that even if the physical Rayleigh fading $g(t,\tau)$ is causal, the composite fading $p_R(\tau)\otimes g(t,\tau)\otimes p_T(\tau)$ is generally noncausal when $p_R(\tau)$ and $p_T(\tau)$ have zero delay. Therefore, the $h_l(n)$ is usually noncausal as indicated in (2). Moreover, even if $g(t,\tau)$ is WSSUS fading, $h_l(n)$ will generally have inter-tap correlation in addition to the temporal correlation. Adopting Clarke's two-dimensional (2-D) isotropic model [27] for the physical Rayleigh fading $g(t,\tau)$, we can obtain statistical properties of the discrete-time fading $h_l(n)$ as shown in [26]. Some key statistics are stated below.

The discrete-time fading $h_l(n)$ is wide-sense stationary zero-mean complex Gaussian, and its cross-correlation is given by

$$E\left\{h_{l_1}(n_1)h_{l_2}^*(n_2)\right\} = C_{l_1,l_2} J_0[2\pi f_d T_s(n_1-n_2)] \quad (3)$$

where $E(\cdot)$ stands for expectation, $(\cdot)^*$ denotes complex conjugate, C_{l_1,l_2} is the inter-tap correlation coefficient between l_1th tap and l_2th tap, $J_0(\cdot)$ is the zero-order Bessel function of the first kind. It is noted that for normalized Rayleigh fading channel, $\sum_{l=-L_1}^{L_2} C_{l,l} = 1$. It is also noted that C_{l_1,l_2} is usually non-zero for $l_1 \neq l_2$ even if the physical channel $g(t,\tau)$ is WSSUS fading [26], [28].

C. Assumption on the Transmitted Signal

Assuming that the binary information is equally likely and independent from bit to bit, and the wireless transmitters employ modulations such as M-ary phase shift keying (MPSK), M-ary quadrature amplitude modulation (MQAM), and M-ary amplitude shift keying (MASK), etc. The modulated data symbol $x(n)$ is zero-mean random variable with correlation given by

$$E\{x(n)x^*(m)\} = \delta(n-m) \quad (4)$$

where $\delta(\cdot)$ is the Kronecker delta function.

III. Theoretical Analysis and Doppler Approximation

In this section, we present key second-order and fourth-order statistics of the received signal $y(n)$, which are useful for designing mobile speed estimation.

Definition: The autocorrelation of the received signal, the autocorrelation of the received signal power, and auto-covariance of the received signal power are defined by

$$R_{yy}(m) = E\{y(n+m)y^*(n)\} \quad (5)$$

$$R_{|y|^2|y|^2}(m) = E\{|y(n+m)|^2|y(n)|^2\} \quad (6)$$

$$V_{|y|^2|y|^2}(m) = E\{[|y(n+m)|^2 - R_{yy}(0)][|y(n)|^2 - R_{yy}(0)]\} \quad (7)$$

Theorem: The received discrete-time signal $y(n)$ has the following statistics

$$R_{yy}(m) = \left(1+\sigma^2\right)\delta(m) + \sum_{\substack{l=-L_1 \\ r-l=m,m\neq 0}}^{L_2} \sum_{r=-L_1}^{L_2} C_{l,r} J_0[\omega_d(r-l)T_s] + \exp(j\omega_d m T_s \cos\theta_0) \sum_{\substack{p=-P_1 \\ q-p=m,m\neq 0}}^{P_2} \sum_{q=-P_1}^{P_2} \sigma_p \sigma_q^* \quad (8)$$

$$\begin{aligned} R_{|y|^2|y|^2}(m) = {} & \left(1+\sigma^2\right)^2 + 2\sigma^2\delta(m) + \sigma^4\delta(m) \\ & + \frac{J_0^2(\omega_d m T_s)}{(1+K)^2} \sum_{l=-L_1}^{L_2} \sum_{s=-L_1}^{L_2} |C_{l,s}|^2 \\ & + \frac{1}{(1+K)^2} \sum_{\substack{l=-L_1 \\ t-l=m,s-r=m,l\neq r,s\neq t}}^{L_2} \sum_{r=-L_1}^{L_2} \sum_{s=-L_1}^{L_2} \sum_{t=-L_1}^{L_2} [C_{l,r}C_{s,t} + C_{l,t}C_{r,s}J_0^2(\omega_d m T_s)] \\ & + \left(\frac{K}{1+K}\right)^2 \sum_{\substack{p=-P_1 \\ w-p=m,u-q=m,p\neq q,u\neq w}}^{P_2} \sum_{q=-P_1}^{P_2} \sum_{u=-P_1}^{P_2} \sum_{w=-P_1}^{P_2} \sigma_p\sigma_q^*\sigma_u\sigma_w^* \\ & + \frac{2K}{(1+K)^2} \sum_{\substack{l=-L_1 \\ w-l=m,u-r=m,l\neq r,u\neq w}}^{L_2} \sum_{r=-L_1}^{L_2} \sum_{u=-P_1}^{P_2} \sum_{w=-P_1}^{P_2} C_{l,r}\sigma_u\sigma_w^* \\ & + \frac{2KJ_0(\omega_d m T_s)\cos(\omega_d m T_s \cos\theta_0)}{(1+K)^2} \\ & \times \Bigg[\sum_{\substack{p=-P_1 \\ p=r,w=s}}^{P_2} \sum_{w=-P_1}^{P_2} \sum_{r=-L_1}^{L_2} \sum_{s=-L_1}^{L_2} \sigma_p\sigma_w^* C_{r,s} \\ & + \sum_{\substack{p=-P_1 \\ w-p=m,s-r=m,p\neq r,w\neq s}}^{P_2} \sum_{w=-P_1}^{P_2} \sum_{r=-L_1}^{L_2} \sum_{s=-L_1}^{L_2} \sigma_p\sigma_w^* C_{r,s} \Bigg] \end{aligned} \quad (9)$$

$$V_{|y|^2|y|^2}(m) = R_{|y|^2|y|^2}(m) - \left(1+\sigma^2\right)^2 \quad (10)$$

where σ^2 is the noise power, and $\omega_d = 2\pi f_d$.

Proof: The proof is lengthy and complex, details are omitted for brevity.

Based on the derived statistics of received signal $y(n)$, we have five remarks as follows.

Remark 1: The autocorrelation of the received signal is zero when the time lag is larger than or equal to $L = L_1 + L_2$, where $L+1$ is the channel length, this is different from the autocorrelation of the fading channel coefficients. Therefore, those algorithms which rely on the autocorrelation (or auto-covariance) of the fading channel coefficients will not work if they are applied to the received signal without knowing the fading channel coefficients. Moreover, knowing the fading channel coefficients are computationally very expensive. Therefore, the algorithms which utilize the autocorrelation and/or auto-covariance of the fading channel coefficients are not realistic for practical wireless systems.

Remark 2: For frequency flat Rayleigh fading channel, $K = 0$, $L_1 = L_2 = 0$, the autocorrelation and auto-covariance of the received signal power are given by $R_{|y|^2|y|^2}(m) = \left(1+\sigma^2\right)^2 + J_0^2(\omega_d m T_s) + 2\sigma^2\delta(m) + \sigma^4\delta(m)$ and $V_{|y|^2|y|^2}(m) = J_0^2(\omega_d m T_s) + 2\sigma^2\delta(m) + \sigma^4\delta(m)$, respectively. Thus, for wireless systems with reasonable high signal-to-noise ratio (SNR) (*i.e.*, σ^2 is small compared to unity), the normalized autocorrelation $R_{|y|^2|y|^2}(m)/R_{|y|^2|y|^2}(0)$ and/or the normalized auto-covariance $V_{|y|^2|y|^2}(m)/V_{|y|^2|y|^2}(0)$ can be employed for estimating mobile speed as "slow" or "fast" as done in [12], [15], where a low-pass filter has been employed to remove out-of-band noise to improve estimation performance.

Remark 3: For frequency flat Rician fading channel, $L_1 = L_2 = 0$ and $C_{0,0} = 1$, the normalized auto-covariance $V_{|y|^2|y|^2}(m)/V_{|y|^2|y|^2}(0)$ of the received signal power can still be employed for mobile speed estimation if the Rice factor K is known and the SNR is high (*e.g.*, SNR $\geq$ 20 dB). However, when K is unknown, and/or the SNR is not high, then the speed estimation accuracy is generally unsatisfactory. That is why many existing algorithms only considered noise-free and small Rice factor scenarios.

Remark 4: For frequency selective Rayleigh fading channel, $K = 0$, the normalized autocorrelation $R_{|y|^2|y|^2}(m)/R_{|y|^2|y|^2}(0)$ and normalized auto-covariance $V_{|y|^2|y|^2}(m)/V_{|y|^2|y|^2}(0)$ of the received signal power may still be employed for estimating mobile speed as "slow" or "fast" with degraded accuracy as done in [12], [15].

Remark 5: For frequency selective Rician fading channel, the normalized auto-covariance $V_{|y|^2|y|^2}(m)/V_{|y|^2|y|^2}(0)$ may not provide satisfactory results for estimating mobile speed as "slow" or "fast". Because the normalized value $V_{|y|^2|y|^2}(m)/V_{|y|^2|y|^2}(0)$ can be small for both fast speed (large Doppler) and slow speed (small Doppler) due to realistic SNR, Rice factor K, and C_{l_1,l_2}, where C_{l_1,l_2} is inter-tap correlation of frequency selective fading channels.

To make use of the auto-covariance (10) for mobile speed estimation in realistic frequency selective Rician fading channels, we need to remove the effect of the additive noise and to mitigate the influence of the frequency selectivity (10). To achieve this goal, we take a close look into the auto-covariance given by (10), and we find

$$V_{|y|^2|y|^2}(m) = \frac{1}{(1+K)^2}\left[aJ_0^2(\omega_d m T_s) + 2KbJ_0(\omega_d m T_s) \times \cos(\omega_d m T_s \cos\theta_0)\right], \qquad m \geq L, \quad (11)$$

where

$$a = \sum_{l=-L_1}^{L_2}\sum_{s=-L_1}^{L_2} |C_{l,s}|^2 \quad (12)$$

$$b = \sum_{p=-P_1}^{P_2}\sum_{\substack{w=-P_1 \\ p=r,w=s}}^{P_2}\sum_{r=-L_1}^{L_2}\sum_{s=-L_1}^{L_2} \sigma_p \sigma_w^* C_{r,s}. \quad (13)$$

It is known that $\omega_d L T_s$ is generally much smaller than unity for all the practical wireless systems, therefore, $J_0(\omega_d L T_s) \doteq 1$ and $\cos(\omega_d L T_s \cos\theta_0) \doteq 1$, and the modified auto-covariance $V_{|y|^2|y|^2}(m)/V_{|y|^2|y|^2}(L)$ is given by

$$\frac{V_{|y|^2|y|^2}(m)}{V_{|y|^2|y|^2}(L)} \doteq \frac{aJ_0^2(\omega_d m T_s) + 2KbJ_0(\omega_d m T_s)\cos(\omega_d m T_s \cos\theta_0)}{a+2Kb}, \qquad m \geq L. \quad (14)$$

Unfortunately, K, a, b and θ_0 are unknown parameters. Hence, (14) may not be directly employed for mobile speed estimation. However, if we utilize the approximation $J_0(x) \cong 1 - \frac{x^2}{4}$ and $\cos(x) \cong 1 - \frac{x^2}{2}$ for x being small, then we obtain the following approximation

$$\frac{V_{|y|^2|y|^2}(m)}{V_{|y|^2|y|^2}(L)} \cong \frac{a+2Kb-(\pi f_d m T_s)^2\left[2a+2Kb+4Kb\cos^2\theta_0\right]}{a+2Kb}, \qquad m \geq L. \quad (15)$$

Since θ_0 is the angle of arrival of the LOS component, for one specific phone call, θ_0 can be either a fixed value or a time-varying parameter depending on the mobility of the mobile. For another phone call, θ_0 can be another fixed value or another time-varying parameter. Therefore, in the entire sample space, θ_0 may be treated as a random variable with uniform distribution over $(0, 2\pi]$, and the mathematical expectation of $\cos^2\theta_0$ is equal to $\frac{1}{2}$. Thus, equation (16) can be further approximated by

$$\frac{V_{|y|^2|y|^2}(m)}{V_{|y|^2|y|^2}(L)} \cong \frac{a+2Kb-(\pi f_d m T_s)^2\left[2a+4Kb\right]}{a+2Kb} = 1 - 2\left(\pi f_d T_s m\right)^2, \qquad m \geq L. \quad (16)$$

From (16), we can obtain the maximum Doppler estimation as follows:

$$f_d \cong \frac{1}{\pi m T_s}\sqrt{\frac{V_{|y|^2|y|^2}(L) - V_{|y|^2|y|^2}(m)}{2V_{|y|^2|y|^2}(L)}}, \qquad m > L. \quad (17)$$

Through the estimated maximum Doppler, we can calculate the mobile speed via $\nu = f_d c/f_c$, where c is the speed of light and f_c is the carrier frequency.

Based on the theoretical analysis shown above, we can see that the modified auto-covariance given by (17) can be used for mobile speed estimation. Compared to existing techniques, this new method has a few advantages. First, this method avoids the noise effect which was significant for large Rice factor K for many existing algorithms. Second, the new method does not need to estimate the Rice factor K. Third, the new method does not need to know the fading channel coefficients. Fourth, the new method even does not need to directly estimate the channel's statistics, it only needs to calculate the auto-covariance of the received signal power. Therefore, the new method is computationally efficient and robust to intersymbol interference and noise.

We leave this section with one final remark on equation (17). The auto-covariance of the received signal power is a theoretical function of time lag. It is derived by ensemble average and usually calculated by time average for ergodic fading channels. This time-average calculation requires very large time duration of data, and therefore, leads to a large latency of the estimation method. In next section, we will present a practical algorithm for mobile speed estimation with very short latency in the range of less than one second. We will show in the simulation section, the new algorithm provides accurate estimation on the mobile speed corresponding maximum Doppler up to 500 Hz.

IV. PRACTICAL ALGORITHMS FOR BROADBAND WIRELESS SYSTEMS

In this section, we take single-carrier broadband wireless systems as an example to demonstrate how to utilize the theoretical equation (17) to develop a practical mobile speed estimation algorithm.

For single-carrier broadband wireless systems, the transmitted data symbols $\{x(k)\}$ are commonly partitioned into blocks of length M data symbols, and each block is added a cyclic prefix of M_{cp} symbols with $M_{cp} \geq L$. At the receiver, the prefix is discarded to eliminate inter-block interference and to assist data symbol detection via frequency-domain channel equalization [29], [30].

Let $s_p(k) = |y_p(k)|^2$ be the kth received symbol's instantaneous power of the pth block after discarding the prefix. In principle, we can estimate the mobile speed via computing the auto-covariance of $s_p(k)$ with N blocks data. However, N has to be very large to achieve a reliable estimation of mobile speeds, and estimation latency will be in the order of tens of seconds if not minutes, which is too large to be acceptable for mobiles changing speeds in seconds. To keep the estimation latency small (less than one or two seconds) while achieving satisfactory estimation reliability, we pass $s_p(k)$ through a low-pass filter $f(l)$ to suppress noise and the power fluctuation caused by data symbols $x(k)$, and the filtered power signal is given by

$$\hat{s}_p(k) = \sum_{l=0}^{L_f} s_p(k-l) f(l) \tag{18}$$

where L_f is the order of the low-pass filter.

The auto-covariance of N block filtered signal $\hat{s}_p(m)$ is given by

$$V_N^s(pT_s) = \sum_{k=1}^{N}\sum_{n=1}^{M-p} [\hat{s}_k(n+p)-\bar{s}][\hat{s}_k(n)-\bar{s}]\,,\ p \geq L+L_f \tag{19}$$

$$V_N^b(qT_b) = \sum_{k=1}^{N-q}\sum_{n=1}^{M} \left[\hat{s}_{k+q}(n)-\bar{s}\right][\hat{s}_k(n)-\bar{s}]\,,\ q \geq 1 \tag{20}$$

where T_s is the symbol interval, $T_b = (M + M_{cp})T_s$ is the time duration of a block, and $\bar{s}$ is given by

$$\bar{s} = \frac{1}{MN}\sum_{k=1}^{N}\sum_{n=1}^{M}\hat{s}_k(n). \tag{21}$$

It is noted here that p is chosen to be not smaller than $L+L_f$, then the noise effect and the intersymbol interference caused by the fading channel impulse response and the low-pass filter $f(l)$ will be eliminated for $V_N^s(pT_s)$, moreover, L_f is chosen to have $(L+L_f)T_s \cong 30\mu s$, then $J_0\left[\omega_d\left(L+L_f\right)T_s\right] \doteq 1$ even if $f_d = 500$Hz, which corresponds to the mobile speed of 270km/h at carrier frequency of 2GHz.

We are now in a position to present our practical algorithm as follows:

Algorithm: Mobile speed estimation for frequency selective Rician fading channels.

Step 1. Choose a set values for $\{N, q, p\}$ to have $NT_b \leq 1s$, $T_F \triangleq qT_b \approx 5ms$, and $T_f \triangleq pT_s \approx 30\mu s$.

Step 2. Compute $V_N^s(T_f)$ and $V_N^b(T_F)$. Classify the maximum Doppler as “low”, “medium” and “high” as follows:

$$\frac{V_N^b(T_F)}{V_N^s(T_f)} \begin{cases} > 0.95, & \text{low} \\ < 0.3, & \text{high} \\ \text{otherwise}, & \text{medium} \end{cases} \tag{22}$$

where the coefficients 0.95 and 0.3 are classification thresholds, they can be changed to other values for different level of “low”, “high” and “medium”.

Step 3. After classifying Doppler into one of the three categories, maximum Doppler can be estimated as follows:

$$f_d = \begin{cases} \frac{1}{\pi 10 T_F}\sqrt{\frac{V_N^s(T_f)-V_N^b(10T_F)}{2V_N^s(T_f)}}, & \text{low} \\ \frac{1}{\pi T_F}\sqrt{\frac{V_N^s(T_f)-V_N^b(T_F)}{2V_N^s(T_f)}}, & \text{medium} \\ \frac{1}{\pi 0.1 T_F}\sqrt{\frac{V_N^s(T_f)-V_N^b(0.1T_F)}{2V_N^s(T_f)}}, & \text{high.} \end{cases} \tag{23}$$

Step 4. Estimate the mobile speed via computing $v = \frac{f_d c}{f_c}$, where c is the speed of light and f_c is the carrier frequency.

We described our idea and algorithm for broadband wireless systems as an example, however, it is noted that this method can be applied to other multiple access wireless protocols including TDMA, CDMA and OFDM systems with various modulation schemes.

V. SIMULATION RESULTS

The performance evaluation of the proposed algorithm has been carried out by extensive computer simulation with various system parameters and wireless fading channels. In this paper, we consider a broadband wireless system with symbol interval $T_s = 0.25\mu s$, data block size $M = 512$ symbols. The transmit filter $p_T(\tau)$ and receive filter $p_R(\tau)$ are normalized square root raised cosine filter with roll-off factor being 0.3. The wireless fading channel has 120 taps with half a T_s spacing between every two consecutive taps. The average power of the first 40 taps ramps up linearly and the last 80 taps ramps down linearly, the total power of the fading channel is

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normalized to unity. The T_s-spaced composite channel impulse response has $L = 63$ taps with inter-tap correlations. The cyclic prefix length is chosen to be $M_{cp} = L - 1 = 62$. We designed a low-pass filter with $L_f = 50$ taps with T_s spaced.

We employ the fading simulator presented in [31] to generate frequency selective Rayleigh and Rician fading coefficients, we also generate the random 8PSK or 64QAM modulated signal $x(n)$, partition it and add cyclic prefix, then we compose the received signal $y(n)$ based on (2). At the receiver, we discard the prefix and use the proposed algorithm to estimate the mobile speed.

Fig. 2 and Fig. 3 show the estimation accuracy of the algorithm for frequency selective Rayleigh fading channels with 8PSK modulated signals and 64QAM modulated signals, respectively. As can be seen, the proposed estimation algorithm is insensitive to the modulation schemes. Both figures show that the proposed algorithm provides good estimation accuracy even if the SNR is as low as 0dB, and when SNR is 5dB or higher, the standard deviation of the estimated Doppler is small.

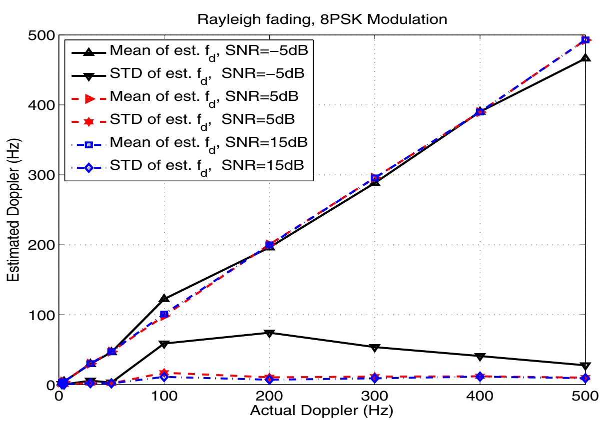

Fig. 2. Estimation accuracy for frequency selective Rayleigh fading channel with 8PSK modulated signals.

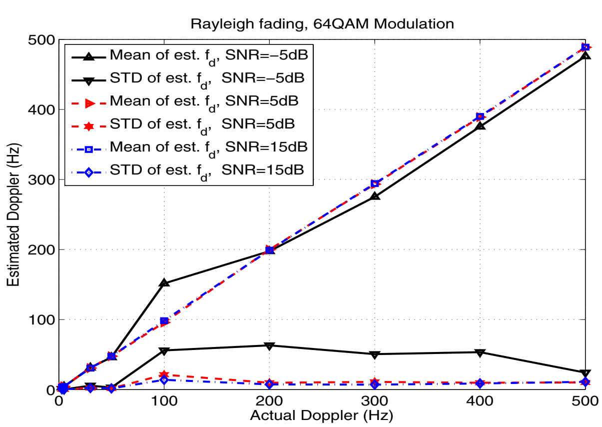

Fig. 3. Estimation accuracy for frequency selective Rayleigh fading channel with 64QAM modulated signals.

Fig. 4 and Fig. 5 show the estimation accuracy of the algorithm for frequency selective Rician fading channels (K=5) with 8PSK modulated signals and 64QAM modulated signals, respectively. As can be seen, the proposed estimation algorithm is also insensitive to the modulation schemes. Both figures show that the proposed algorithm provides good estimation accuracy when SNR is 5dB or higher.

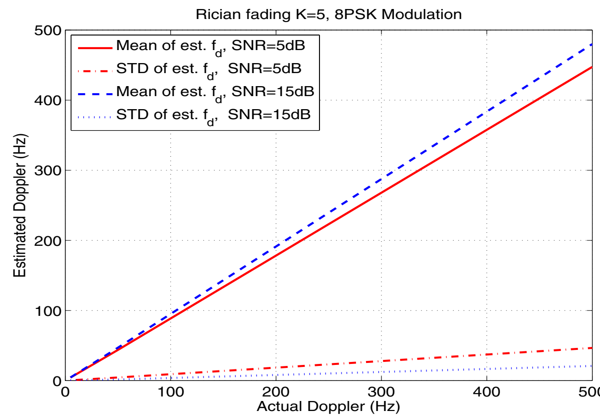

Fig. 4. Estimation accuracy for frequency selective Rician fading channel with 8PSK modulated signals.

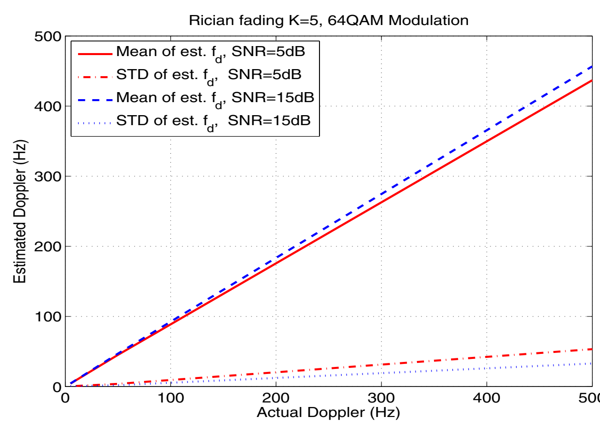

Fig. 5. Estimation accuracy for frequency selective Rician fading channel with 64QAM modulated signals.

As can be seen from these figures, our new algorithm provides reliable results for estimating mobile speeds for broadband wireless communications over frequency selective Rayleigh and/or Rician fading channels.

It is noted that when we increase SNR, our algorithm will give better estimation results. If we increase the slot number N, our method will also give better estimation results. However, if we decrease the slot number N and/or SNR, then the estimation accuracy will decrease. The estimator starts to report mobile speed estimation results within one second after the communication is established.

It is also noted that the thresholds 0.95 and 0.3 are chosen for illustration purpose only, they can be chosen to other values to get better estimation accuracy in favor of high speed estimation or low speed estimation or a compromise for both.

VI. Conclusion

In this paper, we analyzed the statistical properties of the received signals which contain unknown transmitted data, unknown frequency selective Rician fading coefficients, and additive white Gaussian noise. Based on the received signal's statistics, we proposed a mobile speed estimation algorithm. The new algorithm employed modified auto-covariance to first classify "slow", "medium" and "fast" mobiles, then estimate the maximum Doppler frequency and calculate the mobile speed. Extensive simulations have shown that our new algorithm provides very reliable estimation results for broadband wireless communications over various fading channel conditions, which include Rician fading, frequency selective channel with severe multipath spread, and low signal-to-noise ratio scenario, etc. This method is computationally efficient, and it only need simple arithmetic operations such as multiplications, additions and subtractions. As a by-product, our theoretical analysis explains why many existing algorithms fail on certain channel conditions including frequency selective Rayleigh fading, and/or realistic SNR values.

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References

[1] K. Kawabata, T. Nakamura, and E. Fukuda, "Estimating velocity using diversity reception," in *Proc. IEEE Veh. Tech. Conf.*, vol.1, pp.371-374, June 1994.

[2] M.D. Austin and G.L. Stuber, "Velocity adaptive handoff algorithms for microcellular systems," *IEEE Trans. Veh. Technol.*, vol.43, pp.549-561, Aug. 1994.

[3] T.L. Doumi and J.G. Gardiner, "Use of base station antenna diversity for mobile speed estimation," *Electronics Letters*, vol.30, no.22, pp.1835-1836, Oct. 1994.

[4] J.M. Holtzman and A. Sampath, "Adaptive averaging methodology for handoffs in cellular systems," *IEEE Trans. Veh. Technol.*, vol.44, pp.59-66, Feb. 1995.

[5] L. Wang, M. Silventoinen, and Z. Honkasalo, "A new algorithm for estimating mobile speed at the TDMA-based cellular system," in *Proc. IEEE Veh. Tech. Conf.*, pp.1145-1149, April 1996.

[6] M. Hellebrandt, R. Mathar, and M. Scheibenbogen, "Estimating position and velocity of mobiles in a cellular radio network," *IEEE Trans. Veh. Technol.*, vol.46, pp.65-71, Feb. 1997.

[7] D. Mottier and D. Castelain, "A doppler estimation for UMTS-FDD based on channel power statistics," in *Proc. IEEE Veh. Tech. Conf.*, pp.3052-3056, Spet. 1999.

[8] K.D. Anim-Appiah, "On generalized covariance-based velocity estimation," *IEEE Trans. Veh. Technol.*, vol.48, pp.1546-1557, Sept. 1999.

[9] R. Narasimhan and D.C. Cox, "Speed estimation in wireless systems using wavelets," *IEEE Trans. Commun.*, vol.47, pp.1357-1364, Sept. 1999.

[10] M. Turkboylari and G.L. Stuber, "Eigen-matrix pencil method-based velocity estimation for mobile cellular radio systems," *Proc. IEEE ICC'00*, pp.690-694, June 2000.

[11] L. Krasny, H. Arslan, D. Koilpillai, and S. Chennakeshu, "Doppler spread estimation in mobile radio systems," *IEEE Commun. Lett*, vol.5, pp.197-199, May 2001.

[12] C. Xiao, K.D. Mann, and J.C. Olivier, "Mobile speed estimation for TDMA-based hierarchical cellular systems," *IEEE Trans. Veh. Technol.*, vol.50 pp.981-991, July 2001.

[13] C. Tepedelenlioglu and G.B. Giannakis, "On velocity estimation and correlation properties of narrow-band mobile communication channels," *IEEE Trans. Veh. Technol.*, vol.50, pp.1039-1052, July 2001.

[14] R. Narasimhan and D.C. Cox, "Estimation of mobile speed and average received power in wireless systems using best basis method," *IEEE Trans. Commun.*, vol.49, pp.2172-2183, Dec. 2001.

[15] C. Xiao, "Estimating Velocity of Mobiles in EDGE Systems," in *Proc. IEEE ICC'02*, pp.3240-3244, April 2002.

[16] H. Schober and F. Jondral, "Velocity estimation for OFDM based communication systems," in *Proc. IEEE VTC'02*, pp.715-718, Sept. 2002.

[17] C. Juncker, P. Toft, and N. Merch, "Speed estimation for WCDMA based on the channel envelope derivative," in *Proc. IEEE SPAWC'03*, pp.527-531, June 2003.

[18] G. Azemi, B. Senadji, and B. Boashash, "Mobile unit velocity estimation based on the instantaneous frequency of the received signal," *IEEE Trans. Veh. Technol.*, vol.53, pp.716-724, 2004.

[19] M. Porretta, *et al*, "Estimating position and velocity of mobile terminals in a microcellular network using an adaptive linear regression setup," in *Proc. IEEE PIMRC'04*, pp.561-565, 2004.

[20] O. Mauritz, "A hybrid method for Doppler spread estimation," in *Proc. IEEE Veh. Technol. Conf.*, pp.962-965, Sept. 2004.

[21] L. Zhao and J.W. Mark, "Mobile speed estimation based on average fade slope duration," *IEEE Trans. Commun.*, vol.52, pp.2066-2069, Dec. 2004.

[22] G. Park, S. Nam, T. Yu, D. Hong, and C. Kang,"A modified covariance-based mobile velocity estimation method for Rician fading channels," *IEEE Commun. Lett.*, vol.9, pp.706-708, Aug. 2005.

[23] S. Mohanty, "VEPSD: a novel velocity estimation algorithm for next-generation wireless systems," *IEEE Trans. Wireless Commun.*, vol.4, pp.2655-2660, Nov. 2005.

[24] K.E. Baddour and N.C. Beaulieu, "Robust Doppler spread estimation in nonisotropic fading channels," *IEEE Trans. Wireless Commun.*, vol.4, pp.2677-2682, Nov. 2005.

[25] P.A. Bello, "Characterization of randomly time-variant linear channels," *IEEE Trans. Commun. Sys.*, pp.360-393, Dec. 1963.

[26] C. Xiao, J. Wu, S.Y. Leong, Y.R. Zheng, and K.B. Letaief, "A discrete-time model for triply selective MIMO Rayleigh fading channels," *IEEE Trans. Wireless Commun.*, vol.3, pp.1678-1688, Sept. 2004.

[27] R.H. Clarke, "A statistical theory of mobile-radio reception," *Bell Syst. Tech. J.*, pp.957-1000, Jul.-Aug. 1968.

[28] A.J. Paulraj, R. Nabar, and D. Gore, *Introduction to Space-Time Wireless Communications*, Cambridge University Press, 2003.

[29] M. V. Clark, "Adaptive frequency-domain equalization and diversity combining for broadband wireless communications," *IEEE JSAC*, vol.16, pp.1385-1395, Oct. 1998.

[30] D. Falconer, S. L. Ariyavisitakul, A. Benyamin-Seeyar, and B. Eidson, "Frequency domain equalization for single-carrier broadband wireless systems," *IEEE Commun. Mag.*, vol.40, no.4, pp.58-66, Apr. 2002.

[31] C. Xiao, Y.R. Zheng, and N.C. Beaulieu, "Novel sum-of-sinusoids simulation models for Rayleigh and Rician fading channels," *IEEE Trans. Wireless Commun.*, vol.5, pp.3667-3679, Dec. 2006.