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FIRST ARRIVAL DETECTION BASED ON CHANNEL ESTIMATION FOR POSITIONING IN WIRELESS OFDM SYSTEMS

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ABSTRACT

Based on the estimated channel, this paper presents a new method for first arrival estimation for positioning application in OFDM mobile communication systems. In the new method, the characteristics of the information theoretic criteria is exploited to estimate the time of arrival (TOA). The information theoretic criteria is established on the basic of the different statistical characteristics of noise and the mobile channel. In the proposed algorithm, the calculation of the autocorrelation matrix and their eigenvalues are not required. Therefore, the complexity of the proposed method is low. Simulation results show that the performance of system in terms of the detection rate is very high. An accurate estimation of the first arrival path (or the time of arrival) can be obtained even though the first arrival path is strongly attenuated and the system suffers from strong additive noise.

1. INTRODUCTION

An accurate estimates of the time of arrival (TOA) for the first arrival path (FAP) is very important for place determination in mobile communication and proper estimation acquires special relevance [1]. Under the assumption that receiver and transmitter are perfectly synchronized, the propagation delay reveals the length of the corresponding path. The tap with the shortest propagation delay relates to the shortest transmission route. If this route is the line of sight (LOS) connection between the receiver and the transmitter, then the length of this transmission route is the distance between the receiver and the transmitter [2]. From this point of view, one can state that the distance between the transmitter and the receiver can be calculated on the basis of the estimated channel.

An optimal detection scheme for subscriber location based on Generalized Likelihood Ratio Test (GLRT) for detection the first arrival path from the estimated channel has been introduced in [1]. The application of super resolution techniques for TOA estimation for indoor positioning application has been introduced in [3].

To estimate the First Arrival Path (FAP), the information theoretic criteria is established on the basis of the different statistical characteristics of additive noise and the mobile channel. Whereby, the variance of the true Channel Impulse Response (CIR) is distributed in the area of the channel window, whereas variance of noise per channel tap is uniformly distributed on the whole length of the estimated CIR. The channel window is the area of the estimated CIR which is limited by the first arrival path and the last arrival path of the channel.

The purpose of this paper is to improve the algorithm presented in [4] to determine the FAP for positioning application. The scheme proposed in this paper is based on the concept that proposed in [2]. This criteria is applied directly to the estimated Delay Power Spectrum (DPS) without the requirements of the eigenvalues determination of the channel autocorrelation matrix. Therfore, the computation complexity of the proposed algorithm is significantly reduced. However, how accurate is the estimated arrival path depends also on how small is the sampling interval. To improve the case that the sampling interval is not small enough to get an accurate time of arrival, the interpolation solution can be applied.

The paper is organized as follows: Section 2 explains the OFDM system model. The principles of the proposed FAP estimation algorithm are presented in Section 3. The performance of the FAP estimator is shown in Section 4. Finally, concluding remarks are given in Section 5.

2. OFDM CHANNEL AND TOA ESTIMATION SYSTEM MODEL

For coherent communication systems, the channel must be estimated in order to reconstruct the received signal. The channel parameters to be estimated are the CIR and the FAP. The position of the first arrival path depends mainly on the environment transmission, and varies more slowly than its coefficients. The estimated CIR consists of several taps. Each tap corresponds to a transmission route and the corresponding propagation delay. If we use not only the information of the shortest tap, but also the whole information of all taps including the channel coefficients, the time delay, and the arrival angle, then this information give us more precise location of the receiver[2]. This concept is illustrated in Fig.1.

It is obvious that in each location the channel parameters are different. Therefore, these channel parameters give a good basis for position determination. For example, when the arrival angles of the transmitted signals to the receivers are known at different times, the movement direction of the receiver can be detected. Moreover, the maximal Doppler frequency f_D can be evaluated from the estimated channel. This information helps us to determine the relative velocity between the receiver and the transmitter. We denote the estimated channel coefficient $\hat{h}_{k,i}$, which contains the true channel coefficient $h_{k,i}$ and noise component added to the channel

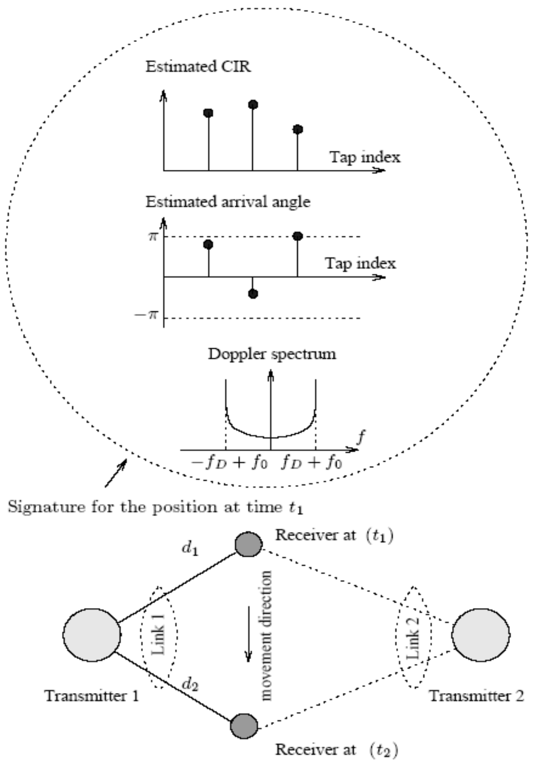

Figure 1: position determination based on the estimated channel [2]

coefficient in time domain $w^{cir}_{k,i}$, as follows [5]:

$$\hat{h}_{k,i} = h_{k,i} + w^{cir}_{k,i} \tag{1}$$

where k, i are respectively the time delay index and the instance time index. If T_a is the sampling interval of the system and T_s is the length of an estimated CIR, then the time delay is $\tau_k = kT_a$, and the instance time $t_i = iT_s$.The channel estimation in OFDM systems can be performed by using pilot symbols. These pilots are inserted into data streams at the transmitter, and removed afterwards at the receiver. The investigated system is depicted in Fig. 2. Assuming that the channel is time invariant over one OFDM symbol, the estimated Channel Transfer Function (CTF) $H'_{p,i}$ at the position of the pilot symbol$S_{p,i}$ is obtained by dividing the received pilot symbol $(R_{p,i} + W_{p,i})$ by the known pilot symbol $S_{p,i}$

$$H'_{p,i} = \frac{R_{p,i}}{S_{p,i}} + \frac{W_{p,i}}{S_{p,i}} \tag{2}$$

where p and i indicate the p^{th} subcarrier and the i^{th} OFDM symbol, respectively, and $W_{p,i}$ is the additive white Gaussian noise (AWGN) component. An average estimate for the CTF is obtained by time-averaging the first estimated CTF over I OFDM symbols [6]. The averaging procedure is repeated M times to reduce the effect of noise [5]. The associated estimated CTF at the positions of pilot symbols per one OFDM symbol can be represented in vector form as

$$\vec{\hat{H}}_i = \left[\hat{H}_{0,i}\hat{H}_{1,i}, ..., \hat{H}_{N_p-1,i}\right]^T \tag{3}$$

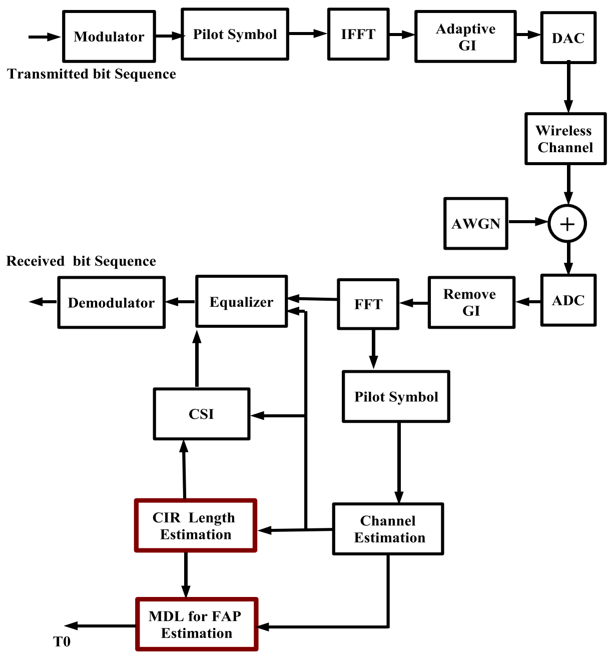

Figure 2: OFDM System

The matrix formulation of the CTF could be written as

$$\vec{\hat{H}}_i = \mathbf{F}\vec{\hat{h}}_i + \vec{\hat{W}}_i \tag{4}$$

where $\mathbf{F}$ is an $(N_p \times N_p)$ discrete-Fourier-transform (DFT) matrix . It is given as

$$\mathbf{F} = \begin{bmatrix} F_{0,0} & F_{0,1} & \cdots & F_{0,N_p-1} \\ F_{1,0} & F_{1,1} & \cdots & F_{1,N_p-1} \\ \vdots & \vdots & \ddots & \vdots \\ F_{N_p-1,0} & F_{N_p-1,0} & \cdots & F_{N_p-1,N_p-1} \end{bmatrix} \tag{5}$$

The element (n,k) of $\mathbf{F}$ reads

$$F_{n,k} = e^{j2\pi nk/N_p}, \ ,n,k \in \{0,...,N_p\}, \tag{6}$$

where N_p is the number of pilot symbols per one OFDM symbol and T denotes the transpose operation. The vector $\vec{\hat{h}}$ which represents N_p samples of the CIR at the i^{th} OFDM symbol is the counterpart vector of $\vec{\hat{H}}_i$ in the time domain. It is given by

$$\vec{\hat{h}}_i = [\hat{h}_{0,i}, \hat{h}_{1,i}, \ldots, \hat{h}_{N_p-1,i}]^T \tag{7}$$

The distance between the channel tap indices k and k+1 is assumed to be equal to the sampling interval T_a of the system. Although the above procedure results in an estimation for the CIR, the related CIR length may be overestimated due to the additive noise, which produces some additional taps beyond the noise-free CIR length. Recently a new method for prediction the CIR length L has been developed [4]. The predicted CTF and the corresponding CIR are given by

$$\vec{\hat{H}}_i = \left[\hat{H}_{0,i}\hat{H}_{1,i}, ..., \hat{H}_{L-1,i}\right]^T \tag{8}$$

$$\vec{\hat{h}}_i = [\hat{h}_{0,i}, \hat{h}_{1,i}, \ldots, \hat{h}_{L-1,i}]^T \tag{9}$$

3. NEW FAP ESTIMATION ALGORITHM

The FAP path may not necessarily be the one with highest power, in the non line of sight NLOS case the first arrival path may suffer attenuation higher than the other later arrival and some times the first arrival belong to the noise (see Fig.3). In this case it is necessary to predict the position of the first

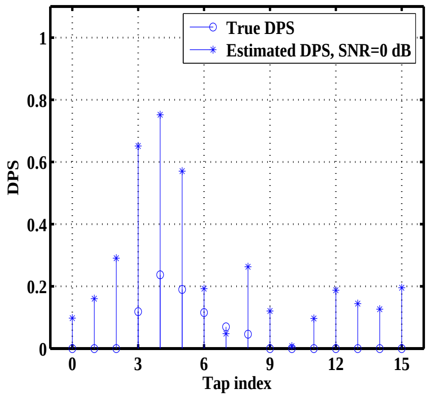

Figure 3: Estimated power delay spectrum distorted by additive noise.

arrival in the presence of noise. The estimated DPS in such case is represented by

$$\vec{\check{P}}_i = [\sigma_1{}^2, \cdots, \sigma_{N_s}{}^2, \hat{P}_{N_s+1,i}, \ldots, \hat{P}_{L-1,i}]^T \qquad (10)$$

where N_s is the noise subspace length and N_s+1 is the position of the FAP. Without lose of generality the CTF could be written as

$$\vec{\check{H}}_i = \left[\hat{H}_{L-1,i}\hat{H}_{L-2,i}, ..., \hat{H}_{0,i}\right]^T \qquad (11)$$

Since the channel is assumed to be represented by an uncorrelated scattering process, the autocorrelation matrix $\mathbf{R}_{HH}$ of the CTF can be written as

$$\mathbf{R}_{HH} = \mathbf{E}[\vec{\check{H}}, \vec{\check{H}}^{\mathbf{H}}] \qquad (12)$$

where H denotes the Hermitian transpose, a Singular Value Decomposition (SVD) of $\mathbf{R}_{HH}$ results in

$$\mathbf{R}_{HH} = \mathbf{U}^{\mathbf{H}}\Lambda\mathbf{U} \qquad (13)$$

where Λ is $(L \times L)$ diagonal matrix with elements [7]

$$P_k = E[\check{h}_k.\check{h}_k^*] + \sigma^2 \qquad (14)$$

that represent the singular values of $\mathbf{R}_{HH}$. The parameter σ^2 represents the noise variance and $k \in [L-1, \cdots, 0]$. The elements of Λ are then L samples of the estimated noisy DPS $P(\tau)$. The L-dimentional space to which $\mathbf{R}_{HH}$ belongs consists generally of two orthogonal spaces: A signal space and a noise space. The matrix Λ can be partitioned as

$$\Lambda = \begin{bmatrix} \Lambda_{CIR} & 0 \\ 0 & \Lambda_N \end{bmatrix} \qquad (15)$$

where Λ_{CIR} is $(L-N_s) \times (L-N_s)$ diagonal matrix containing the $L-N_s$ elements P_k=E$[h_k.h_k^*]$ +$\sigma_k{}^2$, $k \in [L-1, \ldots, L-N_s-1]$, and Λ_N is $N_s \times N_s$ diagonal matrix containing the (N_s) elements $\sigma_k{}^2$, $k \in [N_s, \ldots, 0]$. The latter characterizes the additive noise. Since the additive noise is white, the noise terms $\sigma_k{}^2$ are approximately equal, i.e.,

$$\sigma_1^2 \approx \cdots \approx \sigma_{N_s}^2 \approx \cdots \approx \sigma_L^2 \qquad (16)$$

Let us consider now the structural and statistical properties of $\mathbf{R}_{HH}$. The conditional probability density function of the statistically independent complex Gaussian random vector $\vec{\check{H}}$ is given by [8]

$$f(\vec{\check{H}}|\phi) = \frac{1}{\pi^L det(\mathbf{R}_{H,H})} \mathrm{e}^{-\vec{\check{H}}[\mathbf{R}_{HH}]^{-1}\vec{\check{H}}} \qquad (17)$$

where ϕ =$[P_1, \ldots, \quad P_k, \sigma^2, \mathbf{f}_1^T, \ldots, \mathbf{f}_k^T\ ]^T$ is the parameter vector of the model, $\mathbf{f}_1, \ldots, \mathbf{f}_k$ are the columns of the $\mathbf{F}$-matrix, and *det* denotes the determinant of the matrix. The corresponding log-likelihood function is given by

$$L(\phi) = -L\log(det(\mathbf{R}_{HH})) - tr[\mathbf{R}_{HH}]^{-1}\mathbf{R}_{HH} \qquad (18)$$

where *tr* represent the trace of the matrix. The close expression of the maximum log-likelihood function (ML) in (18) has been obtained in [9], and is given by

$$L(\phi) = -(L-k)M_B \log \frac{G(P_{k+1}, \cdots, P_L)}{A(P_{k+1}, \cdots, P_L)} \qquad (19)$$

where the functions $G(.)$ and $A(.)$ are the geometric and arithmetic mean of their arguments, respectively [10]. The value of M_B needs to be selected to provide a balance between resolution and stability of the criteria. Equation (19) shows that there is no need to find the autocorrelation matrix or its eigenvalues. This significantly reduces the computation complexity of the proposed algorithm.

Utilizing the ML-function for estimating N_s results in its maximum allowable value. On the other hand, utilizing information theoretic criteria for the model selection, includes an additional bias correction term (penalty function). It is added to the log-likelihood function in order to bias the over model estimation. This results in the MDL-function, which is given in [9] as

$$MDL(k) = L(\phi) + f_p(k, L) \qquad (20)$$

where the penalty function $f_p(k,L)$ is given by[9, 4]

$$f_p(k,L) = \frac{1}{4}k(2L-k)\log(M_B) + k \qquad (21)$$

According the properties of MDL, the noise space length N_s is now taken to be the value of $k \in \{0, 1, \ldots .L-1\}$ for which MDL(k) is minimum. The starting of signal space (N_s+1) is the position of the FAP. The performance of MDL as a function of M and I are found in [4]

4. SIMULATION

The simulated OFDM system parameters of Fig. 2 have been selected similar to that of HiperLAN/2 [5, 11]. These read:

- Bandwidth of the system B = 20 MHz,

- Sampling interval: $T_a = 1/B$=50 ns,
- FFT length: $N = 64$,
- OFDM symbol duration: $T_s = N.T_a$=3.2 μs.

In multipath propagation the amplitudes of the paths change rapidly due to fast fading, but their angles and delays are usually fairly constant over several time slots, for reasonable geometries and speeds. The detection of first arrival path is tested under critical case when the channel is time varying and the system suffers from strong additive noise. The channel model in this paper for each transmission rout is based on a typical indoor channel model A in [11]. Each subchannel is modeled by Monti Carlo method [12, 13]

$$h(\tau_k,t) = \frac{1}{\sqrt{N_h}} \sum_{k=1}^{L} p[k] \sum_{i=1}^{N_h} e^{j(2\pi f_{k,i}t+\theta_{k,i})}\delta(\tau-\tau_k) \qquad (22)$$

where $f_{k,i} = f_D \sin(2\pi u_{k,i}), \theta_{k,i} = 2\pi u_{k,i}$ and N_h are called the discrete Doppler frequencies, the Doppler phases, and the number of harmonic functions respectively. The propagation delay τ_k relates to the k^{th} channel path. The qualities $u_{k,i}$ are independent random variables, each with a uniform distribution in the range (0,1]. They are independently generated for each sub-channel. The maximum Doppler frequency f_D is selected to be 50 Hz. The number of harmonic functions N_h is chosen to be 40. Fig. 4 illustrate the time varying channel transfer function $H(f,t)$. The CIR length L is first predicted

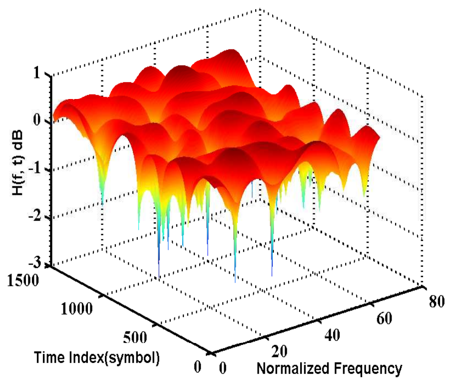

Figure 4: Time variant channel transfer function

using the new method that introduced in [4]. The important factor for the detection of first arrival is the relative power with respect to the 2^{nd} arrival path. Fig. 5 shows the probability of detection as a function of relative power of first arrival with respect to the 2nd one. Generally MDL gives best accuracy close to 100% at low SNR and the relative amplitude is 0.2 or more. At higher values of SNR the probability of detection is not sensitive to the relative amplitude of the FAP. The shape of MDL for the prediction of CIR is shown in Fig. 6 for low SNR, where the noise apace length N_s and FAP position N_s+1 are found by the index k that gives minimum value for MDL. Even in the condition of low SNR, an accurate estimate FAP can be achieved.The stability of the MDL criteria for different simulations and at SNR of 0dB is shown in Fig.7, where the FAP of the simulated channel corresponds to the time delay index 5.

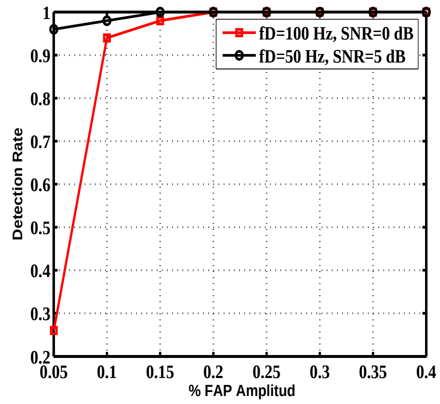

Figure 5: Probabilty of detection performance

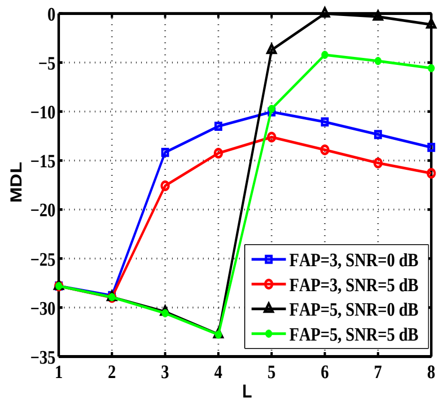

Figure 6: Simulation results of MDL in different cases SNR

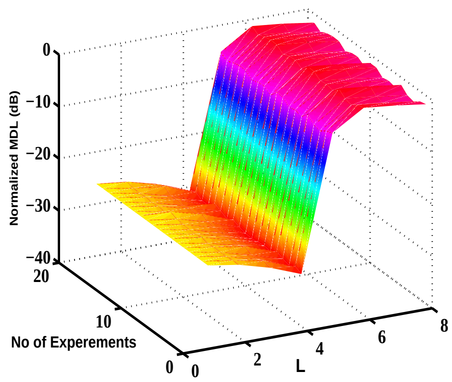

Figure 7: Stability of the estimation criteria

5. CONCLUSIONS

In this paper a new method for determination the FAP in OFDM wireless systems has been proposed. The algorithm is based on the principle of information theoretic criteria and MDL of the estimated channel impulse response. The performance analysis demonstrate the functionality of the first arrival detection considering HiperLAN/2 system parameters with time varying channel and at very low SNR. The proposed algorithm has low complexity and requires no significant computation effort.

REFERENCES

[1] Játiva, E., Vidal, J. “First Arrival Detection for positioning in Mobile Channels,” In Proc. *13th IEEE International Symposium on Personal, Indoor and Mobile Radio Communications (PIMRC)*, vol.4, 5-18 Sept. 2002,p. 1540 - 1544,.

[2] K. Kyamakya , V. D. Nguyen ” Cross-layer Optimization, Especially Combination of Channel Estimation and Position Determination in Multihop Wireless Networks (Cellular and Adhoc),” Proc. *58th IEEE Semiannual Vehicular Technology Conference, IEEE VTC2003 Fall*, Orlando, USA, 69. Oct. 2003, pp. 15371543

[3] Xinrong Li, Kaveh Pahlavan ” Super-Resolution TOA Estimation With Diversity for Indoor Geolocation,” *IEEE Transactions on Wireless Communications*, vol.3 No.1.,pp 224-234, 2004

[4] A. Aassie-Ali, V. D. Nguyen, K. Kyamakya, A. S. Omar ” Estimation of the Channel-Impulse-Response Length for Adaptive OFDM Systems Based on Information Theoretic Criteria,” Accepted for publication in *63rd IEEE Vehicular Technology Conference, IEEE VTC-Spring 2006*, Melbourne, Australia,7-10 May 2006.

[5] V. D. Nguyen, *Channel Impulse Response Length Estimation and Interference Cancellation for OFDM Systems*, Germany, Shaker Verlag, Aachen

[6] B. Yang, K. B. Letaief, R. S. Cheng, and Z. Cao, ”Channel Estimation for OFDM Transmission in Multipath Fading Channels Based on Parametric Channel Modeling,” *IEEE Transactions on Communications*, Vol. 49, No. 3, pp 467-478, March 2001.

[7] S. V. Vaseghi, *Advanced Digital Signal Processing and Noise Reduction,* John Wiley & Sons , 2000.

[8] S. Valaee, and P. Kabal ”An Information Theoretic Approach to Source Enumeration in Array Signal Processing,” *IEEE Trans. Signal Processing*, vol. 52, no. 5, pp. 1171-1178, May 2004.

[9] M. Wax and T. Kailath ” Detection of signals by information theoretic criteria,” *IEEE Trans. Acoust., Speech, Signal Processing*, vol.ASSP-33, pp. 387-392, April 1985.

[10] A. P. Liavas, and P A. Regalia ”On the Behavior of Information Theoretic Criteria for Model Order Selection,” *IEEE Trans. Signal Processing*, vol. 49, no. 8,pp. 1689-1695, Aug. 2001

[11] ETSI DTS/BRAN-0023003, HIPERLAN Type 2 Technical Specification; Physical (PHY) layer. 1999.

[12] H. Schulze, ”Stochastische Modelle und digitale Simulation von Mobilfunkkanalen,” in U. R. S. I./ITG Conf. in Kleinheubach 1988, Germany (FR), Proc. *Kleinheubacher Reports of the German PTT*, Darmstadt, Germany, 1989, pp. 473-483, in German.

[13] P. Höher, ”A Statistical Discrete-Time Model for the WSSUS Multipath Channel,” *IEEE Trans. Veh. Technology*, vol. 41, no. 4, pp. 461468, Nov. 1992.