

Accurate Simulation of Nonisotropic Fading Channels with Arbitrary Temporal Correlation

Amirhossein Alimohammad
Dept. of Elec. & Comp. Eng.

San Diego State University
San Diego, CA, 92182, U.S.A
Email: aalimohammad@mail.sdsu.edu

Saeed Fouladi Fard
Ukalta Engineering

4344 Enterprise Square, 10230 Jasper Ave
Edmonton, AB, T5J 4P6, Canada
Email: saeed@ukalta.com

Bruce F. Cockburn

Dept. of Elec. & Comp. Eng.
University of Alberta
Edmonton, AB, T6G 2V4, Canada
Email: cockburn@ualberta.ca

Abstract—The accurate simulation of wireless channels is important since it permits the realistic and repeatable performance measurement of wireless systems. A new technique is proposed for simulating Rayleigh fading channels with isotropic or non-isotropic scattering and with arbitrary temporal correlation. Fading samples are generated by passing Gaussian samples through a spectrum shaping filter. A new iterative algorithm is then presented for designing stable complex infinite impulse response (IIR) filters with quantized coefficients. The algorithm utilizes a least-squares cost function augmented with a barrier function to ensure filter stability and to reduce quantization noise. The performance of the proposed filter design algorithm is verified with 18-bit fixed-point simulations of different fading channel scenarios including isotropic and nonisotropic scattering and the IEEE 802.11n Model F fading spectrum.

I. MOTIVATION AND BACKGROUND

The accurate design and verification of wireless communication systems needs faithful modeling and accurate simulation of realistic radio propagation channels. Prior to the availability of standardized channel models, wireless products needed to be verified using extensive and expensive field testing. A far less costly approach is to model the behavior of radio channels on a fading channel simulator. Two major techniques have been widely used for simulating Rayleigh fading channels. In the first approach, the so-called sum-of-sinusoids (SOS) model, the Rayleigh fading process is modeled as the superposition of a sufficiently large number of sinusoidal waves. This approach was originally proposed by Clarke [1] and later simplified by Jakes [2]. Over the past four decades several modified SOS-based models have been proposed (e.g., [3], [4]). This technique is widely used in the COST 259, COST 273, and IST-Winner fading channel models [5], [6].

The second well-known approach for fading channel simulation is to first generate a complex, zero-mean, Gaussian random process with independent samples. Then the fading process is obtained by passing the variates through a suitable spectrum-shaping filter (SSF). This technique is used to simulate the TGN fading channel models proposed for the 802.11n wireless standard [7]. The filtering process can be carried out in the time or frequency domains with digital or analog filters. In particular, one can generate the fading variates by multiplying the Gaussian samples in the frequency domain with suitable coefficients and then taking the inverse Fourier

transform (IDFT) [8]. It has been shown that generating fading samples with a single Fast Fourier Transform (FFT) operation requires a relatively large memory and, hence, results in an inefficient implementation [9]. An autoregressive (AR) modeling approach has also been proposed for generating fading processes by passing the white noise samples through an all-pole infinite impulse response filter (IIR) [9]. To produce samples with accurate statistics, the AR model needs a large filter order, which greatly increases the number of required multiplications. Also, implementation of the AR fading simulator requires highly accurate variables, which makes it unappealing for compact fixed-point implementations.

While the above techniques have been used to simulate isotropic scattering channels, they might not be the best candidates for simulating nonisotropic Rayleigh fading channels. In nonisotropic fading, the power spectral density (PSD) of the fading samples is asymmetric, implying an SSF with complex-valued coefficients. In conventional practice, filters are designed with standard tools (e.g., `fdatool` or `iirlpnorm` in MATLAB). While this approach is applicable to SSF design for isotropic fading channels, it is not appropriate for designing the stable complex filters required for shaping the spectrum of nonisotropic channels. Similarly, the IIR filter design procedure in [10] can be only applied to filters with real coefficients, and hence, it is not suitable for designing the complex filters required in nonisotropic channels.

Our previous work focused on designing fading channel filters for the isotropic scenario (when the angle of arrival of the received multipath signals is uniformly distributed and we assume the use of an omni-directional antenna at the receiver) [11]–[15]. This work considers the more general scenario of non-isotropic scattering due to real-world antenna directivity and possible selective attenuation along some propagation directions. In this article we address the problem of designing IIR filters for the accurate simulation of nonisotropic Rayleigh fading channels with arbitrary time correlation properties. Several algorithms exist for designing stable IIR filters with real coefficients [16]–[20]. However, designing stable IIR filters with complex coefficients has been less well studied. The main contributions of this work are as follows:

- A new algorithm for designing general stable IIR filters with complex and real coefficients is proposed. The maximum

radius of the poles and/or zeros is limited to ensure numerical stability. The least-squares cost function is formulated in polar coordinates and augmented with a barrier function that keeps the poles (and potentially the zeros) within the unit circle. This method poses a sufficient (but not necessary) stability condition on the filter.

- To minimize the computational requirements, filters are quantized for fixed-point implementation and, hence, variables are implemented with the minimum possible fixed-point word-length. Since reducing the word-length can greatly impact the response (and potentially also the stability) of the designed filter, our proposed filter design technique searches for the best filters with fixed-point coefficients that meet our word-length budget.

- Several examples are provided for designing fixed-point fading channel simulators with 18-bit IIR filters and 16-bit coefficients. Both isotropic and nonisotropic fading are simulated and the fixed-point results show a close match between several generated statistical properties and the theoretical references.

The rest of this article is organized as follows. Section II reviews the nonisotropic channel model and related work on complex filter design. The new filter design technique is presented in Section III. We used our proposed complex filter design algorithm for the simulation of two different fading channel scenarios. Numerical results of these simulators are presented in Section IV. Finally, Section V makes some concluding remarks.

II. FADING CHANNEL CHARACTERISTICS AND THE RELATED WORK ON COMPLEX FILTER DESIGN

In wireless communication systems, the received signal strength varies significantly due to destructive and constructive interference resulting from multipath propagation [21]. Isotropic scattering refers to the case in which the incident direction of the received multipath signals, or the angle of arrival (AOA), is uniformly distributed. Assuming two-dimensional isotropic scattering with an omni-directional antenna at the receiver [2], the PSD functions associated with both the in-phase or quadrature components of a complex fading signal have the well-known Jakes' U-shaped band-limited form [22] with independent in-phase and quadrature samples. However, such assumptions have been challenged [23] and experimentally demonstrated [24]–[26] to be inaccurate due to the selective attenuation of some propagation directions as well as antenna directivity [27]. Such effects produce a nonuniform probability density function (pdf) for the AOA at the receiver. The pdf of the AOA has a great impact on the second-order statistics of the fading process including the correlation functions, the level crossing rate (LCR) and the average fade duration (AFD) [26].

Several nonuniform pdfs have been proposed in the literature for the angle of arrival including geometrically-based AOA pdfs [28], the Gaussian pdf [29], the quadratic pdf [30], the Laplace pdf [24], the cosine pdf [31], and the von Mises pdf [32]. The von Mises pdf, which includes the uniform AOA distribution as a special case, is supported experimentally with empirical measurements of narrowband fading channels [32].

The authors in [32] argue that the von Mises pdf is preferable because it can approximate other nonuniform pdfs and it is mathematically convenient for analysis.

In this article, we assume narrowband fading in which the complex envelope of the fading process is given by

$$c(t) = c_I(t) + j c_Q(t) \\ = \lim_{N \rightarrow \infty} \frac{1}{\sqrt{N}} \sum_{n=1}^N \alpha_n e^{j(2\pi f_D \cos(\psi_n) + \varphi_n)}, \quad (1)$$

where N is the number of sinusoids, f_D is the Doppler frequency, the $\{\psi_n\}$ are independent and identically distributed (i.i.d.) angles of arrival of the incoming wave at the receiver antenna with distribution $p_\Psi(\psi)$, the $\{\varphi_n\}$ are i.i.d. phases with uniform distribution over $[-\pi, \pi)$, and the $\{\alpha_n\}$ are deterministic complex constants that are normalized to satisfy $\sum_{n=1}^N |\alpha_n|^2 = N$. When the scattering encountered in the propagation environment is nonisotropic, the PSD function associated with $c(t)$ is given by [32]

$$S_c(f) = \frac{e^{\kappa \cos(\tilde{\psi}) \frac{f}{f_D}} \cosh\left(\kappa \sin(\tilde{\psi}) \sqrt{1 - \left(\frac{f}{f_D}\right)^2}\right)}{\pi I_0(\kappa) \sqrt{1 - \left(\frac{f}{f_D}\right)^2}} \quad (2)$$

where $\kappa \geq 0$ controls the beamwidth, $\tilde{\psi}$ denotes the average AOA of the scattered component, and $I_m(\cdot)$ is the m -th order modified Bessel function of the first kind. To obtain (2) it is assumed that the AOA of the scattered component is distributed with the von Mises/Tikhonov distribution [33] as follows

$$p_\Psi(\psi) = \frac{\exp[\kappa \cos(\psi - \tilde{\psi})]}{2\pi I_0(\kappa)}, \quad \psi \in [-\pi, \pi). \quad (3)$$

Note that when the beamwidth parameter $\kappa = 0$, the AOA has a uniform distribution over $[-\pi, \pi)$ and (2) reduces to Jakes' U-shaped symmetric spectrum $S_c(f) = (\pi \sqrt{1 - (f/f_D)^2})^{-1}$. The autocorrelation function (ACF) of $c(t)$ is obtained by taking the inverse Fourier transform of (2) as follows [32]

$$R_c(\tau) = \frac{I_0\left(\sqrt{\kappa^2 - 4\pi^2 f_D^2 \tau^2 + 4j\kappa \cos(\tilde{\psi})\pi f_D \tau}\right)}{I_0(\kappa)}. \quad (4)$$

Two other important statistical properties of $c(t)$ are the LCR and AFD. These properties can greatly affect the design and analysis of the wireless systems, not only at the physical layer, but also at the link and network layers [34]. The LCR of the envelope $|c(t)|$ is defined as the expected number of envelope crossings per second at a given level R with positive slope. It can be shown that the LCR of $c(t)$ is [35]

$$L(\rho) = \frac{\sqrt{I_0^2(\kappa) - I_1^2(\kappa) + \cos(2\tilde{\psi})[I_0(\kappa)I_2(\kappa) - I_1^2(\kappa)]}}{I_0(\kappa)} \times \\ \sqrt{2\pi} f_D \rho \exp(-\rho^2), \quad (5)$$

where for beamwidth $\kappa = 0$ (isotropic scattering), (5) reduces to the Rayleigh LCR function $L(\rho) = \sqrt{2\pi} f_D \rho \exp(-\rho^2)$.

AFD determines the average length of error bursts. For the case where the AOA is modeled by the von Mises pdf, the AFD is [35]

$$T(\rho) = \frac{I_0(\kappa)}{\sqrt{I_0^2(\kappa) - I_1^2(\kappa) + \cos(2\tilde{\psi})[I_0(\kappa)I_2(\kappa) - I_1^2(\kappa)]}} \times \frac{\exp(\rho^2) - 1}{\sqrt{2\pi}f_D\rho}. \quad (6)$$

For isotropic scattering ($\kappa = 0$), the AFD reduces to $T(\rho) = (\exp(\rho^2) - 1)/(\sqrt{2\pi}f_D\rho)$.

To mimic the behavior of a realistic wireless channel, a fading channel simulator should be able for generating a sequence of path gains $\{c(t)\}$ with statistics that accurately match the ideal reference models. An efficient technique to generate a sequence $\{c(t)\}$ with the desired statistics is to pass a stream of independent Gaussian samples through a suitably designed SSF. For the case of nonisotropic scattering, the SSF must have a magnitude response equal to the square root of the magnitude of (2). In this case, $S_c(f)$ is asymmetric and the SSF can be implemented using a complex IIR filter. While IIR filters have a reduced computational complexity and greater stopband attenuation compared to FIR filters of similar size, it is challenging to implement stable and accurate complex IIR filters in fixed-point arithmetic.

Various methods using nonlinear optimization [36]–[38], linear programming [39], and semidefinite programming [40] have been suggested for designing digital IIR filters in the complex Chebyshev sense. Least-squares methods have been applied extensively to design FIR and IIR filters [41]–[45]. However, ensuring the stability of a complex IIR filter whose coefficients are obtained by least-squares methods is not straightforward. Many authors appear to have neglected this topic and have concentrated instead on finding close approximations to the desired frequency response. When designing real filters, for which only the magnitude and not the phase response is important, all unstable factors in the filter transfer function $1/(1 - s_k e^{j\phi_k} e^{-j\omega})$ can be replaced with their stable counterparts $1/(e^{-j\omega} - s_k e^{j-\phi_k})$ with identical magnitude response, where $s_k e^{j\phi_k}$ is the k -th pole. If, however, the phase response is also important, such as when the filter coefficients are complex, this method cannot be applied.

When optimization techniques are used for designing a filter, various approaches exist to ensure stabilization. One method proposed in [36] is to start the optimization from a stable point and then control the step size so that the solution never leaves the stable region. However, this method is computationally expensive and is not easy to implement with traditional optimization procedures. A second method is to make sure that the target response is minimum-phase [16] or that it has a “large-enough” group delay [17]. These conditions cannot be met in many situations since the flexibility to modify the target frequency response is restricted. In a third method, explicit constraints are imposed on the coefficients of the denominator of the transfer function [18], [19]. This

technique, however, has some limitations and affects the filter quality [20]. Finally, in a fourth method, the least squares cost function is modified so that the minimum always falls in the stable region [20]. In [20] the authors suggest adding a barrier function to the original cost function to avoid filter instability. To design the barrier function, first they form an all-pole proxy transfer function consisting of all of the filter poles. The barrier function is basically the sum of the squared amplitude of a section of the impulse response of the proxy transfer function. If the filter is unstable, the tail will have (large) non-zero values.

III. SPECTRUM SHAPING FILTER DESIGN

To design a compact IIR spectrum shaping filter, the filter is represented as a product of first-order sections (FOSs)

$$H(e^{-j\omega}) = A \prod_{k=1}^{\Gamma} \frac{1 - r_k e^{j(\theta_k - \omega)}}{1 - s_k e^{j(\phi_k - \omega)}}, \quad (7)$$

where A is a positive scaling factor, $r_k e^{j\theta_k}$ and $s_k e^{j\phi_k}$ are the k -th complex zero and pole, respectively, and Γ is the filter order, i.e., the number of FOSs. When the beamwidth parameter is $\kappa = 0$, the poles and zeros of (7) appear as complex conjugate pairs and thus the shaping filter can be implemented using $\Gamma/2$ canonic second-order sections (SOSs), also known as biquads.

In practical applications, since f_D is much smaller than the sample rate F_s , the designed SSF would have an extremely narrow passband and a very sharp cut-off. We can reduce the complexity and increase the accuracy of the SSF by designing it instead at a lower sampling frequency. For an efficient implementation we propose to design the SSF at a sampling frequency F_1 , where $4f_D < F_1 \leq 8f_D$. At a later stage the generated samples are up-sampled to reach the required sampling rate F_s . To design the SSF, we first define $2 \times M$ frequency points in the interval $u_i \in [-0.5, 0.5]$ and then try to fit the frequency response of SSF to the sampled desired response as follows

$$y_i^d = \begin{cases} e^{\kappa \cos(\tilde{\psi}) \frac{F_1 u_i}{2f_D}} \frac{\sqrt{\cosh(\kappa \sin(\tilde{\psi})) \sqrt{1 - (\frac{F_1 u_i}{f_D})^2}}}{\sqrt{\pi I_0(\kappa) \sqrt{1 - (\frac{F_1 u_i}{f_D})^2}}}, & \text{if } |u_i| < \frac{f_D}{F_1} \\ \varepsilon, & \text{otherwise} \end{cases} \quad (8)$$

where $\varepsilon > 0$ is the attenuation in the stopband. We pick the frequency samples at $u_i = (2i + 1 - 2M)/(4M)$, $i = 0, \dots, 2M - 1$ to have appropriate samples at the corner frequencies. Similar to the work in [46], we express $H(e^{-j\omega}) = AF(\mathbf{x}; e^{-j\omega})$, where $F(\mathbf{x}; e^{-j\omega})$ represents the product of FOSs in (7) and the column vector $\mathbf{x}$ of length 4Γ containing r_k, s_k, θ_k and ϕ_k . Next, to find the filter parameters we define the cost function

$$q(A, \mathbf{x}) = \sum_{i=0}^{2M-1} v_i (\log(A|F(\mathbf{x}; e^{-j\omega})|) - \log(y_i^d))^2 + B(\vartheta; \varrho; \mathbf{x}), \quad (9)$$

where the weight vector $\mathbf{v} = [v_1, v_2, \dots, v_{2M}]^T$ is used to allow us to emphasize the error minimization for certain frequency bands. Since the target phase response of the designed filter does not have any impact on the correlation properties, only the magnitude of the filter response needs to be considered in (9). Note that the sum of squared errors on a logarithmic scale is augmented by the parametric barrier function $B(\vartheta; \varrho; \mathbf{x})$. The barrier function $B(\vartheta; \varrho; \mathbf{x})$ is included to keep the poles and zeros within the unit circle (i.e., minimum phase) and is defined as

$$B(\vartheta; \varrho; \mathbf{x}) = \sum_{k=1}^{4\Gamma} b(\vartheta; \varrho; x_k), \quad (10)$$

where

$$b(\vartheta; \varrho; \tau) = \begin{cases} 0 & \text{if } |\tau| \leq \varrho \text{ or } k > 2\Gamma, \\ \vartheta \left(\frac{|\tau| - \varrho}{1 - \varrho} \right)^2 & \text{if } \varrho < |\tau| \leq 1 \text{ and } k \leq 2\Gamma, \\ \frac{2\vartheta}{1 - \varrho} |\tau| - \vartheta \left(\frac{1 + \varrho}{1 - \varrho} \right) & \text{if } |\tau| > 1 \text{ and } k \leq 2\Gamma. \end{cases} \quad (11)$$

In (10), ϑ determines how fast the barrier function grows outside of the unit circle and the parameter ϱ determines an outer boundary for the poles and zeros. The barrier function (10) is especially useful when designing filters for fixed-point implementation since it can be parameterized to keep the poles and zeros at any desired safe distance from the unit circle. Controlling the maximum absolute value of the filter poles limits the quantization effects to an acceptable level.

The coefficients of the SSF are found iteratively. At each iteration, first the optimum scaling factor A^o is calculated as

$$A^o = \prod_{i=0}^{2M-1} \left(\frac{y_i^d}{|F(\mathbf{x}; e^{-j2\pi u_i})|} \right)^{\frac{v_i}{\sum_{i=0}^{2M-1} v_i}}. \quad (12)$$

This expression for A^o is found by differentiating (9) with respect to A and then setting the resulting expression to zero. Algorithm 1 summarizes the steps for our proposed iterative filter design. Here we utilized the Ellipsoid algorithm [47] for its simplicity, however, other techniques can be used for optimization. The inputs to Algorithm 1 include desired

Algorithm 1 Iterative calculation of the filter coefficients

Require: $\varrho; \vartheta; \Omega; \mathbf{v} = [v_i]; \mathbf{y}^d = [y_i^d]; i = 0, \dots, 2M - 1;$

Initialize $k = 0; \mathbf{x}_0, \mathbf{E}_0 = 20 \mathbf{I}_{4\Gamma \times 4\Gamma};$

while $|\mathbf{x}_{k+1} - \mathbf{x}_k| \geq \epsilon$ **do**

 find A_k^o from (12);

 find $\mathbf{g}_k = \mathbf{g}(A_k^o; \mathbf{x}_k);$

$\chi_k = \sqrt{\mathbf{g}_k^T \mathbf{E}_k \mathbf{g}_k};$

$\tilde{\mathbf{g}}_k = \mathbf{g}_k / \chi_k;$

$\mathbf{x}_{k+1} = \mathcal{Q} \left[\Omega, \mathbf{x}_k - \frac{1}{4\Gamma+1} \mathbf{E}_k \tilde{\mathbf{g}}_k \right];$

$\mathbf{E}_{k+1} = \frac{(4\Gamma)^2}{(4\Gamma)^2 - 1} \left(\mathbf{E}_k - \frac{2}{4\Gamma+1} \mathbf{E}_k \tilde{\mathbf{g}}_k \tilde{\mathbf{g}}_k^T \mathbf{E}_k \right);$

$k = k + 1;$

end while

response $\mathbf{y}^d = [y_i^d]$, the fixed-point format Ω used for the filter coefficients, a weight vector $\mathbf{v}$ and the outer boundary ϱ for the zeros and poles. The filter design procedure can

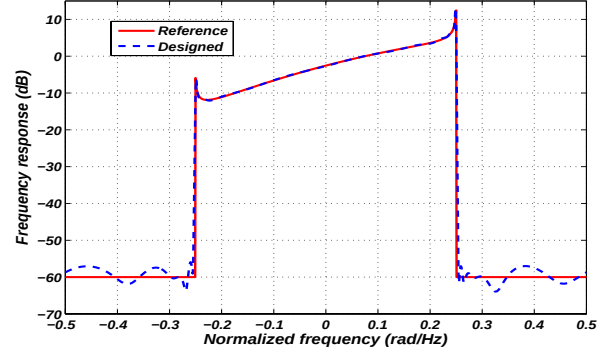


Fig. 1. Frequency response of the designed SSF with $f_D/F_1 = 0.25$, $\kappa = 3$, and $\tilde{\psi} = \pi/4$.

start with a reasonable order Γ for the initial approximation. The filter order can be increased gradually if the desired filter characteristics are not met. The algorithm starts from an arbitrary point $\mathbf{x}_0$ contained within the unit sphere and a relatively large (20 times in this algorithm) initial ellipsoid matrix $\mathbf{E}_0 = 20 \mathbf{I}_{4\Gamma \times 4\Gamma}$, where $\mathbf{I}$ denotes the identity matrix. The algorithm then searches for the optimal solution within the present ellipsoid of feasible points. This algorithm then converges on the optimal solution by successively reducing the size of the ellipsoid by χ_k until it is small enough (i.e., the algorithm has converged) or when $|\mathbf{x}_{k+1} - \mathbf{x}_k|$ is within a chosen accuracy ϵ . The function $\mathcal{Q}[\Omega, \mathbf{x}]$ represents the quantization effects that affect each element of $\mathbf{x}$ in the Cartesian coordinate system (coefficients are transferred to Cartesian coordinates, quantized and then transferred back to polar coordinates). Note that stable IIR filters with real coefficients can also be designed with the above algorithm. To design such filters, the sample update is only performed for half of the poles and zeros, and the other half are simply the complex conjugates of the updated samples.

To demonstrate the performance of our filter design procedure, we designed an example SSF with $\Gamma = 10$ first-order sections. We begin by setting the parameters $\vartheta = 5$ and $\varrho = 0.99$, i.e., the poles and zeros are bounded within a circle of radius 0.99. For all of the FOSs, the number of bits for representing each coefficient is set to $\Omega = 16$. Figure 1 shows the frequency response of the designed filter (with $\epsilon = 0.001$) as well as the desired response for the SSF with the given fading characteristics. As this figure shows, the designed filter accurately produces the desired response within the passband. In the stopband, the designed filter provides more than 55 dB attenuation. Figure 2 shows the position of the poles and zeros for the designed filter. Note that all the poles and zeros are located within a circle of radius 0.99.

IV. FADING SIMULATION

To demonstrate the performance of our filter design procedure, we simulated different fading scenarios. All of the simulations were performed in fixed-point arithmetic, where the filter coefficients are represented with $\Omega = 16$ bit variables

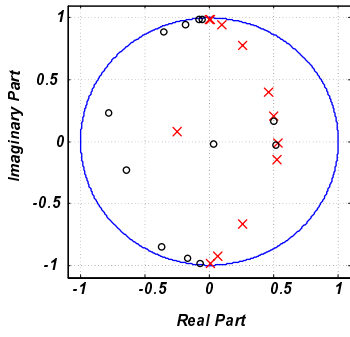


Fig. 2. Positions of the poles and zeros in the unit circle for the designed SSF with $f_D/F_1 = 0.25$, $\kappa = 3$, and $\psi = \pi/4$.

and 18-bit variables are used to store the intermediate signals. Also, all poles and zeros are constrained to lie within a circle of radius $\varrho = 0.99$ and the target sample rate for all scenarios is $F_s = 40$ MHz. As mentioned in Section III, we use an interpolator to up-sample the generated fading variates to reach the desired sampling rate. It would be more computationally-efficient if the up-sampling factors were to be powers of 2. We thus design the first interpolation stage with an up-sampling rate of $I_1 = 16$. Then the up-sampled stream can be efficiently filtered with an IIR lowpass filter designed with our procedure. After interpolation and considering our filter design assumption that $F_1 > 4f_D$, where F_1 is the sampling frequency of the SSF, the resulting signal has a bandwidth small enough to allow us to use fading-specific interpolation lowpass filters (SILPF) with the frequency response

$$D_P^{l_z}(e^{-j\omega}) = \left(\frac{1 - e^{-j\omega P}}{P - P e^{-j\omega}} \right)^{l_z}, \quad l_z \geq 1. \quad (13)$$

In particular, this response is equivalent to the frequency response of a cascade of l_z multiplication-free filters

$$d_P^1 = \underbrace{[P^{-1}, P^{-1}, \dots, P^{-1}]}_{P \text{ elements}}, \quad (14)$$

where P is the interpolation factor. The samples here are passed through T_g successive SILPFs. The i -th SILPF interpolates the signal 2^{k_i} times. Based on the processing architecture, the relation between F_1 and the target output sampling rate is $F_s = 16 \times F_1 \times \prod_{i=1}^{T_g} 2^{k_i}$. From this we have $F_1 = 2^{-(4+S_g)} F_s$, where $S_g = \sum_{i=1}^{T_g} k_i$ is an integer value within the range $\log_2(F_s/f_D) - 7 \leq S_g < \log_2(F_s/f_D) - 6$. Based on the maximum interpolation factor $2^{K_{max}}$, where $K_{max} = 4 + \max\{S_g\}$, each SILPF is assigned a specific interpolation factor.

A. Example 1

As a first example, we designed three different filters to simulate various fading scenarios. In scenario (a) we generated the complex path gains that simulate an isotropic fading channel with Doppler frequency $f_D = 9$ Hz. In (b) we simulated a nonisotropic fading system with Doppler frequency $f_D = -18.5$ Hz, beamwidth $\kappa = 1$, and AOA $\psi = \pi/4$ rad. Finally in (c) we have $f_D = 2.25$ Hz,

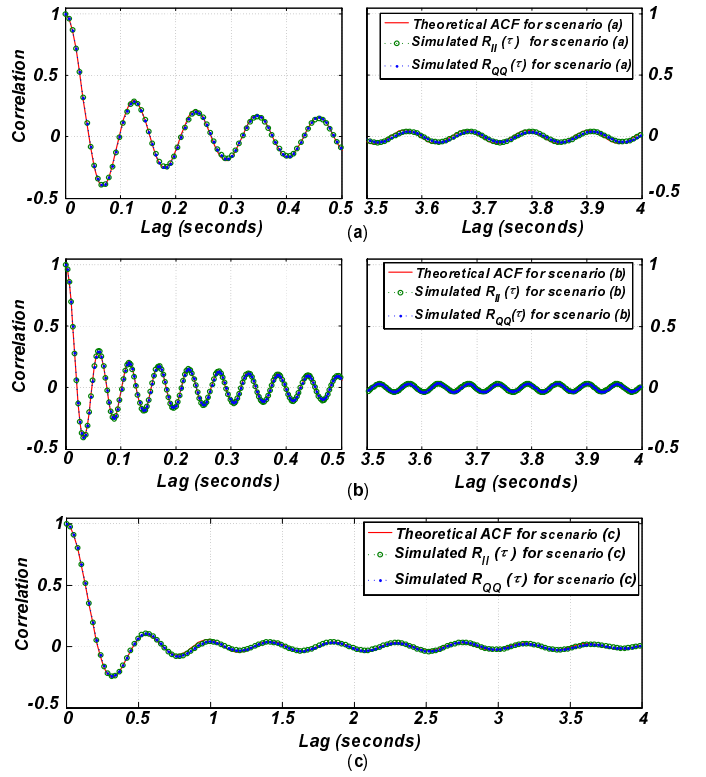


Fig. 3. Autocorrelation for the real and imaginary components of the generated fading processes: (a) $f_D = 9$ Hz, $F_s = 40$ MHz, $\kappa = 0$, and $\psi = 0$ rad, (b) $f_D = -18.5$ Hz, $F_s = 40$ MHz, $\kappa = 1$, and $\psi = \pi/4$ rad. (c) $f_D = 2.25$ Hz, $F_s = 40$ MHz, $\kappa = 5$, and $\psi = \pi/3$ rad.

beamwidth $\kappa = 5$, and AOA $\psi = \pi/3$ rad. Figure 3 compares the autocorrelation for the real and imaginary components of the generated fading samples with the theoretical references. Note that there is a close match between the desired and generated autocorrelations up to four seconds (160 million samples). Also, Fig. 4 plots the cross-correlation between the real and imaginary components of the generated fading samples and the reference curves for up to four seconds. This figure confirms a close match between the achieved and desired curves. Figure 5 compares the normalized LCR of the amplitude of generated complex path gains in scenarios (a), (b) and (c) with the theoretical LCR from (5). The LCR is normalized to $f_D \times T_s$. This figure again confirms excellent agreement between the theoretical and generated curves. The normalized AFD for scenarios (a), (b) and (c) are plotted in Fig. 6. This figure shows that the simulated AFD matches the reference curve with good accuracy over a wide range of normalized fading durations. Finally, in Fig. 7 the pdf of the amplitude of the generated samples is plotted and it can be observed that this pdf accurately mimics the Rayleigh pdf.

B. Example 2

As another example, we designed appropriate filters to simulate the PSD proposed for the IEEE 802.11n indoor wireless fading channel model [7]. In this model, the bell-

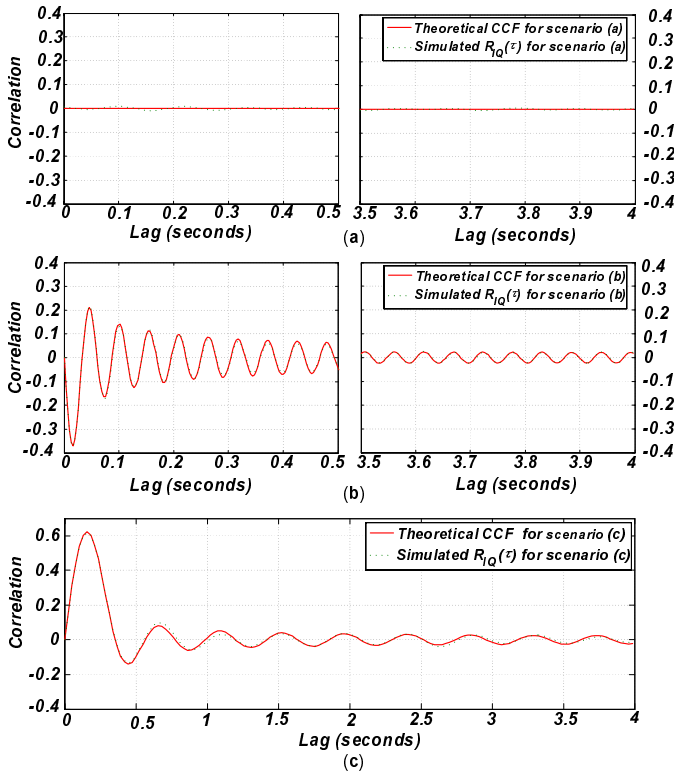


Fig. 4. Crosscorrelation between the real and imaginary components of the generated fading processes: (a) $f_D = 9$ Hz, $F_s = 40$ MHz, $\kappa = 0$, and $\psi = 0$ rad. (b) $f_D = -18.5$ Hz, $F_s = 40$ MHz, $\kappa = 1$, and $\psi = \pi/4$ rad. (c) $f_D = 2.25$ Hz, $F_s = 40$ MHz, $\kappa = 5$, and $\psi = \pi/3$ rad.

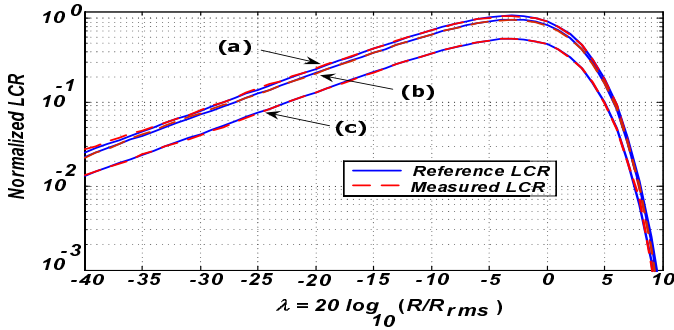


Fig. 5. Normalized level crossing rate of the envelope of the generated fading processes: (a) $f_D = 9$ Hz, $F_s = 40$ MHz, $\kappa = 0$, and $\psi = 0$ rad. (b) $f_D = -18.5$ Hz, $F_s = 40$ MHz, $\kappa = 1$, and $\psi = \pi/4$ rad. (c) $f_D = 2.25$ Hz, $F_s = 40$ MHz, $\kappa = 5$, and $\psi = \pi/3$ rad.

shaped PSD

$$S(f) = \begin{cases} \frac{1}{1+A(\frac{f}{f_D})^2} + \frac{B}{1+C(\frac{f-f_{spike}}{f_{spike}})^2} & f \leq f_{max} \\ 0 & f > f_{max} \end{cases} \quad (15)$$

is proposed for representing indoor propagation. The second term in (15) corresponds to a Doppler component that represents a reflection from a moving vehicle as described in model F of [7]. This component is identified with a spike in the PSD at frequency $f_{spike} = \nu_v/\lambda$, where ν_v is the vehicle speed and λ is the signal wavelength. Here, f_{max} is the maximum

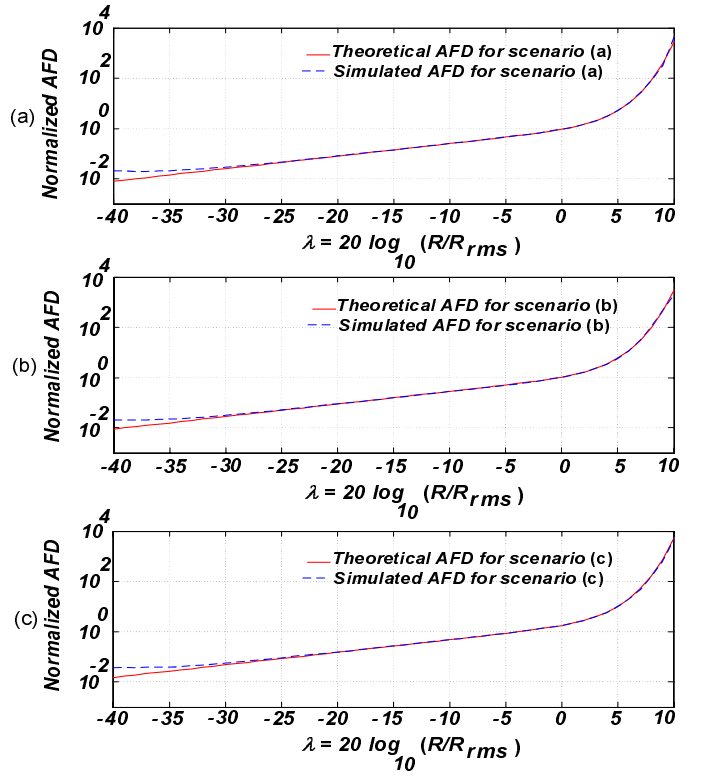


Fig. 6. Normalized average fade duration of the envelope of the generated fading processes: (a) $f_D = 9$ Hz, $F_s = 40$ MHz, $\kappa = 0$, and $\psi = 0$ rad. (b) $f_D = -18.5$ Hz, $F_s = 40$ MHz, $\kappa = 1$, and $\psi = \pi/4$ rad. (c) $f_D = 2.25$ Hz, $F_s = 40$ MHz, $\kappa = 5$, and $\psi = \pi/3$ rad.

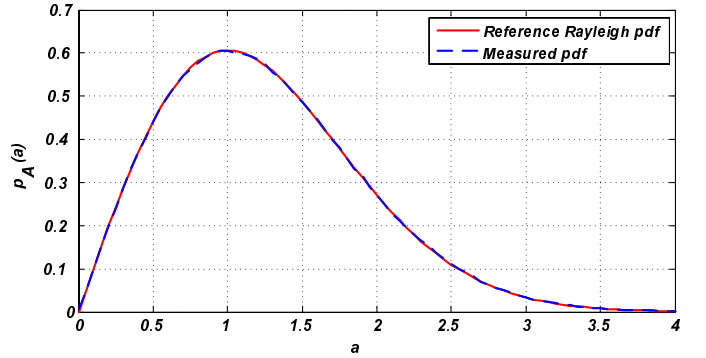


Fig. 7. Probability density function of the amplitude of the generated fading process for a system with $f_D = 2.25$ Hz, $F_s = 40$ MHz, $\kappa = 5$, and $\psi = \pi/3$ Rad. The measured pdf closely matches to the Rayleigh distribution.

frequency component of the Doppler spectrum, which can be set to several times the Doppler frequency [7].

We simulated an indoor fading channel with the above specifications for a system with carrier frequency $F_c = 2.4$ GHz, maximum Doppler frequency 3.0 Hz, and vehicle speed 40.0 KM, which corresponds to $f_{spike} = 88.9$ Hz. The values for the constants A , B , and C in (15) are 9, 0.5, and 90000, respectively [7]. The Bell-shaped Doppler spectrum of the designed filter is modeled with a complex filter of order $\Gamma = 4$ and is shown along with the matching reference spectrum in Fig. 8. Figure 9 shows the autocorrelation of the quadrature

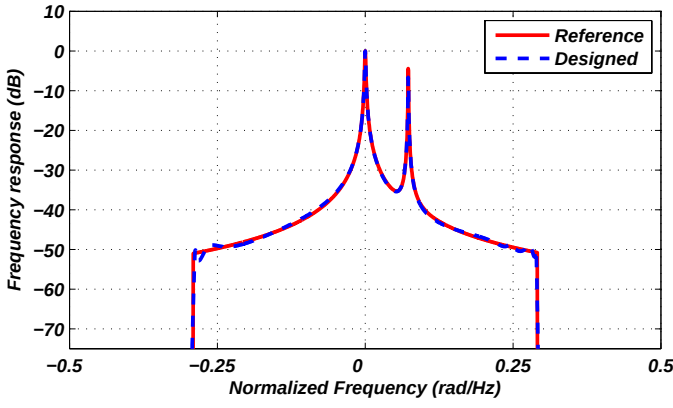


Fig. 8. Reference and designed Bell-shaped Doppler spectra for a moving vehicle.

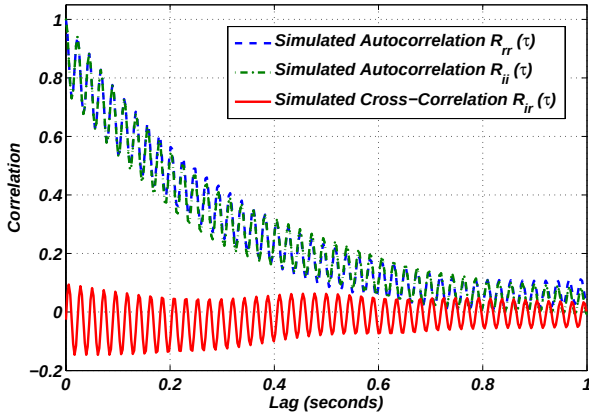


Fig. 9. Autocorrelation and cross-correlation of the quadrature components of the fading process of the IEEE 802n indoor model.

components of the simulated fading processes. The cross-correlation between quadrature components is also plotted in the same figure. Finally, the normalized LCR and AFD of this channel are plotted in Figs. 10 and 11, respectively. As for example 1, the LCR and AFD are normalized to $f_D \times T_s$ and the simulations are performed in fixed-point arithmetic. Compared to the three scenarios in the previous example, note that the simulated indoor 802.11n model has a higher LCR and a lower AFD.

V. CONCLUSION

While investigating the design of nonisotropic Rayleigh fading simulators, we developed an iterative design procedure for arbitrary complex IIR filters with quantized coefficients. The stability of the resulting complex filters is assured by keeping the filter poles within a circle with a radius that is slightly smaller than one. With this procedure it is possible to keep the quantization noise within manageable limits. The resulting filters efficiently generate complex path gains for several different nonisotropic fading channels. Simulation results show that with the proposed technique, the complex

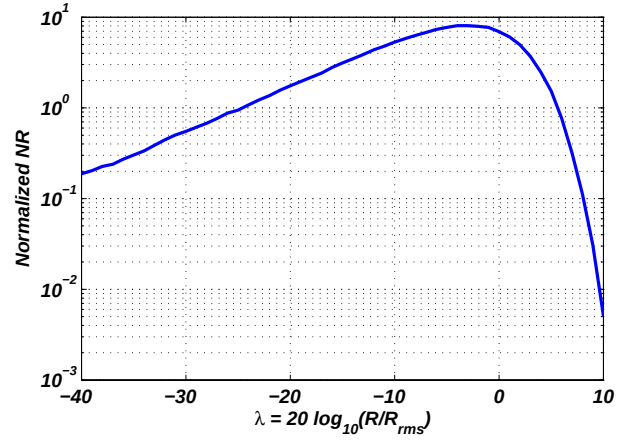


Fig. 10. Normalized level crossing rate of the simulated system.

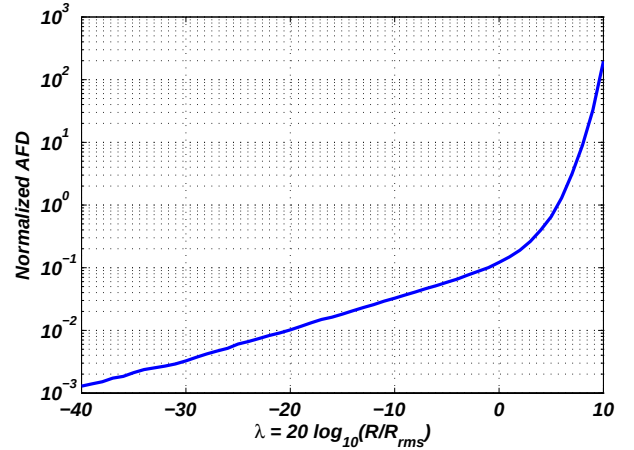


Fig. 11. Normalized average fade duration of the simulated system in Example 2.

path gains of a nonisotropic Rayleigh fading channel can be generated with high statistical accuracy.

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