On the Performance Analysis of Composite Multipath/Shadowing Channels Using the $\mathcal{G}$-Distribution

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***Abstract*—Composite multipath fading/shadowing environments are frequently encountered in different realistic scenarios. These channels are generally modeled as a mixture of Nakagami-m multipath fading and log-normal shadowing. The resulting composite probability density function (pdf) is not available in closed form, thereby making the performance evaluation of communication links in these channels cumbersome. In this paper, we propose to model composite channels by the $\mathcal{G}$-distribution. This pdf arises when the log-normal shadowing is substituted by the Inverse-Gaussian one. This substitution will prove to be very accurate for several shadowing conditions. In this paper we conduct a performance evaluation of single-user communication systems operating in a composite channel. Our study starts by deriving an analytical expression for the outage probability. Then, we derive the moment generating function of the $\mathcal{G}$-distribution, hence facilitating the calculation of average bit error probabilities. We also derive analytical expressions for the channel capacity for three adaptive transmission techniques, namely, i) optimal rate adaptation with constant power, ii) optimal power and rate adaptation, and iii) channel inversion with fixed rate.**

***Index Terms*—Adaptive transmission techniques, fading channels, information rates, log normal, Nakagami distribution and composite distributions, outage probability.**

I. Introduction

MIXTURES of multipath fading and shadowing are frequently encountered in several scenarios. This is particularly the case for communication systems with low mobility or stationary users. In such configurations, the receiver can not mitigate the multipath fading effect by averaging and is subject to the instantaneous composite multipath/shadowed signal [1, Sect.2.4.2.], [2, Sect.2.2.3]. A composite distribution arises therefore as the perfect model for this type of channels. Several composite models were presented in the literature (see [3] and the references therein), maybe one of the most known is the shadowed Nakagami fading channel [4], which is a generalization of the Rayleigh-Lognormal model (called also the Suzuki model) [5]-[6], and consists of a mixture of Nakagami-m multipath fading and log-normal shadowing.

The main drawback of the shadowed Nakagami fading model is that the composite probability density function (pdf) is not in closed-form thereby making the performance evaluation (such as average error probabilities, outage probabilities and channel capacity) of communication links in these channels cumbersome[1]. Serious attempts have been made to obtain a practical composite distribution. We can site, for instance, the well known $\mathcal{K}$-distribution [7] and its generalized version [10]. The $\mathcal{K}$-distribution is obtained by substituting the gamma shadowing to the log-normal one. This distribution proved to be particularly useful in evaluating the performance of composite channels [12]-[15]. Recently, the Inverse-Gaussian pdf was proposed as a substitute to the log-normal one [16]. The authors proved that a composite Rayleigh-Inverse Gaussian distribution approximates the Suzuki distribution more accurately than the $\mathcal{K}$-distribution.

In this paper, we consider the more general Nakagami-Inverse Gaussian model. We demonstrate that this combination gives birth to a composite distribution called the $\mathcal{G}$-distribution. This distribution was first proposed in [17] in the context of Synthetic Aperture Radar (SAR) image modeling. In this paper, we derive several tools for the performance evaluation of single-user communication systems in such channels. Our study starts by deriving an analytical expression for the outage probability. Then, we derive the moment generating function (MGF) of the $\mathcal{G}$-distribution, hence making the average bit error probabilities in this type of channels (with and without diversity combining in an uncorrelated environment) easy to compute. We also derive analytical expressions for the channel capacity with different adaptive transmission techniques. All the mathematical formulae that we develop are expressed in terms of special functions that are available in standard Mathematical packages such as Mathematica.

The remainder of the paper is organized as follows. In Section II, we present the $\mathcal{G}$-distribution and some of its properties. In Section III, we give the expression for the

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[1]Indeed, since the pdf is an infinite integral, then all the expressions for the outage probability, BER and capacity will be under the form of a double infinite integral. In this paper, by approximating the log-normal shadowing by the Inverse-Gaussian one, we will obtain expressions that involve Bessel and Hermite functions which are readily computable with the majority of mathematical software packages.

outage probability and derive the MGF and average bit error probabilities for some modulation schemes. In Section IV, we give an analytical expression for the capacity of three adaptive transmission techniques namely, i) optimal rate adaptation with constant power, ii) optimal power and rate adaptation, and iii) channel inversion with fixed rate. Section V provides some selected numerical results to illustrate the derived formulae and validates the newly developed analytical expressions via computer simulations. Section VI concludes the paper with a summary of the main results.

II. The $\mathcal{G}$-distribution

A. The Probability Density Function of the Composite Envelope

In a composite Nakagami-lognormal channel, the probability density function of the envelope X is

$$f_X(x) = \int_0^{+\infty} f_{X/Y}(x/Y=y) f_Y(y) dy, \quad (1)$$

where $f_{X/Y}$ is the Nakagami-m multipath fading distribution and is given by

$$f_{X/Y}(x/Y=y) = \frac{2m^m x^{2m-1}\exp(-\frac{mx^2}{y})}{\Gamma(m)y^m}, \quad x>0, \quad (2)$$

where $\Gamma(\cdot)$ is the gamma function [30] and m is generally an arbitrary number superior to 0.5. However, in our performance study, in order to derive the forthcoming mathematical expressions, we will restrict this parameter to integer values. Note that this is however not a severe limitation. Indeed, if m is not an integer, then we can apply the formulas that we will derive for $\lfloor m \rfloor$ and $\lfloor m \rfloor + 1$ to obtain practical upper and lower bounds.

In (1), $f_Y(y)$ is the log-normal shadowing distribution, i.e.,

$$f_Y(y) = \frac{1}{\sqrt{2\pi}\sigma y}\exp\left(-\frac{(\ln(y)-\mu)^2}{2\sigma^2}\right), \quad y>0, \quad (3)$$

where μ and σ are, respectively, the mean and the standard deviation of $\ln(y)$. The resulting composite distribution $f_X(\cdot)$ is unfortunately not available in closed-form, hence making the performance evaluation of communication links over such channels very challenging. In order to obtain a more tractable composite distribution, and as it was done in [16], the log-normal shadowing is approximated[2] by the Inverse-Gaussian (IG) distribution which is given by

$$f_Y(y) = \sqrt{\frac{\lambda}{2\pi}} y^{-\frac{3}{2}}\exp\left(-\frac{\lambda(y-\theta)^2}{2\theta^2 y}\right), \quad y>0. \quad (4)$$

The parameters λ and θ are linked to μ and σ by matching the first and second moments of the log-normal distribution with the first and the second moments of the IG distribution as follows[3]

$$\begin{cases} \exp\left(\mu + \frac{\sigma^2}{2}\right) = \theta, \\ \exp\left(2(\mu+\sigma^2)\right) = \theta^2\left(\frac{\theta}{\lambda}+1\right), \end{cases}$$

leading to

$$\begin{cases} \lambda = \frac{\exp(\mu)}{2\sinh(\frac{\sigma^2}{2})}, \\ \theta = \exp\left(\mu + \frac{\sigma^2}{2}\right). \end{cases}$$

Substituting (2) and (4) in (1) and using [30, Eq. (3.471.9)], we find that this substitution results in the following composite distribution given by

$$f_X(x) = \left(\frac{\lambda}{\theta^2}\right)^{m+\frac{1}{2}} \sqrt{\frac{\lambda}{2\pi}} \frac{4m^m \exp(\frac{\lambda}{\theta}) x^{2m-1}}{\Gamma(m)\left(\sqrt{g(x)}\right)^{m+\frac{1}{2}}} K_{m+\frac{1}{2}}\left(\sqrt{g(x)}\right), \quad (5)$$

where $g(x) = \frac{2\lambda}{\theta^2}(mx^2 + \frac{\lambda}{2})$ and $K_\nu(\cdot)$ is the modified bessel function of the second kind of order ν [30]. This pdf was first discovered in [17] where it was called the $\mathcal{G}$-distribution (in [17], it is referred to as $\mathcal{G}_A$). If $m=1$, this distribution reduces to the Rayleigh-Inverse Gaussian model that was considered in [16]. Previously, the gamma pdf was used as a substitute to the log-normal one. The resulting composite pdf is the generalized $\mathcal{K}$-distribution. In [16], the authors demonstrated that a composite Rayleigh-Inverse Gaussian distribution can better describe a composite Rayleigh-lognormal channel. This fact is further confirmed later in our numerical examples section for different shadowing environments.

B. The Probability Density Function of the Instantaneous Composite SNR

With the change of variable $\gamma = \frac{\overline{\gamma}}{E[x^2]}x^2$ in (5) [2, Eq. (2.3)], the pdf of the composite instantaneous signal-to-noise power ratio (SNR) $f_\gamma(\gamma)$, can be easily deduced as

$$f_\gamma(\gamma) = A\frac{\gamma^{m-1}}{\left(\sqrt{\alpha+\beta\gamma}\right)^{m+\frac{1}{2}}} K_{m+\frac{1}{2}}\left(b\sqrt{\alpha+\beta\gamma}\right), \quad (6)$$

where the following constants have been used

$$\begin{cases} A = \frac{(\lambda\overline{\gamma})^{\frac{1+2m}{4}}}{\Gamma(m)}\sqrt{\frac{2\lambda}{\pi\theta}}\exp(\frac{\lambda}{\theta})\left(\frac{m}{\overline{\gamma}}\right)^m, \\ b = \frac{1}{\theta}\sqrt{\frac{\lambda}{\overline{\gamma}}}, \\ \alpha = \lambda\overline{\gamma}, \\ \beta = 2m\theta. \end{cases}$$

Note that if a maximum ratio combiner with M i.i.d. branches is used at the receiver, then the distribution of the instantaneous SNR at the output of this combiner can be readily obtained from (6) by substituting m with Mm and $\overline{\gamma}$ with $M\overline{\gamma}$ (refer to [11] for a similar analysis treating the $\mathcal{K}$-distributed fading), i.e.,

$$f_\gamma(\gamma) = \left(\frac{m\sqrt{\lambda}\gamma}{\sqrt{\overline{\gamma}}h(\gamma)}\right)^{Mm+\frac{1}{2}} \frac{K_{Mm+\frac{1}{2}}(bh(\gamma))}{\gamma\Gamma(Mm)}\sqrt{\frac{2\lambda\overline{\gamma}}{\pi m\theta\gamma}}\exp\left(\frac{\lambda}{\theta}\right), \quad (7)$$

[2]Note that this approximation is not based on field measurements but rather stems from the fact that compared to the Gamma pdf, the IG distribution can provide a better substitute for the log-normal pdf.

[3]Note that in [16], the authors use different matching equations. Indeed, they match the moments of the Suzuki distribution with the moments of the Rayleigh-Inverse Gaussian one. Our method has the advantage of leading to simpler matching equations between (λ,θ) and (μ,σ) and as it will be shown in the simulations, this matching scheme allows for an accurate approximation.

where $h(\gamma) = \sqrt{\alpha + \beta\gamma}$. Consequently, all the following performance study applies also if maximum ratio combining (in i.i.d. fading) is employed at the receiver.

C. Moments and Amount of Fading

The average SNR ([2], p. 4) (or equivalently, the first moment of the instantaneous received SNR) and the amount of fading ([2], p. 12) are two measures used to assess the severity of the fading experienced by a communication system. Unlike other performance criteria, like the average BER, these two measures are relatively easy to compute. Indeed, using [30, Eq. (6.596.3)], the kth moment of the output SNR can be found to be given by

$$E[\gamma^k] = A\int_0^{+\infty} \frac{\gamma^{m+k-1}}{\left(\sqrt{\alpha+\beta\gamma}\right)^{m+\frac{1}{2}}} K_{m+\frac{1}{2}}\left(b\sqrt{\alpha+\beta\gamma}\right) d\gamma \quad (8)$$

$$= \sqrt{\frac{2\lambda}{\pi\theta}} e^{\frac{\lambda}{\theta}} \left(\frac{\overline{\gamma}}{m}\right)^k \frac{\Gamma(m+k)}{\Gamma(m)} K_{k-\frac{1}{2}}\left(\frac{\lambda}{\theta}\right), \quad (9)$$

which yields the following Amount of Fading (AF)

$$AF = \frac{E[\gamma^2]}{E[\gamma]^2} - 1 = \left(\frac{1}{m}+1\right)\left(\frac{\theta}{\lambda}+1\right) - 1. \quad (10)$$

The AF ranges from $\frac{\theta}{\lambda}$ (for $m = +\infty$) to $3\frac{\theta}{\lambda}+2$ (for $m = \frac{1}{2}$).

III. Outage Probability and Average Bit Error Probability

A. Outage Probability

For a reliable communication, the received signal level has to exceed a certain threshold. In a wireless environment, the received signal fluctuates according to the multipath fading and shadowing. The communication system will therefore experience outages when the strength of the perceived signal is insufficient. In this context, the outage probability is an important performance measure of communication links operating over fading channels. Here we define the outage probability as the probability that the instantaneous error probability increases above a certain error rate, say η, i.e., $P_{\text{out}} = P(P_b(\gamma) > \eta)$. This is equal to the probability that the output SNR falls below a given threshold $\gamma_{\text{th}} = P_b^{-1}(\eta)$. It should be noted that in this paper the threshold on the SNR is obtained from the instantaneous BEP. The interested reader is referred to [8], for a case where the threshold is obtained from the mean BEP (averaged over the fast fading). Hence

$$P_{\text{out}} = \int_0^{\gamma_{\text{th}}} f_\gamma(\gamma) d\gamma = F_\gamma(\gamma_{\text{th}}) - F_\gamma(0), \quad (11)$$

where $F_\gamma(\cdot)$ is the primitive of the instantaneous SNR's pdf and is defined as

$$F_\gamma(\gamma) = \int f_\gamma(\gamma) d\gamma = A I_{m,m}(\gamma, \beta, \alpha), \quad (12)$$

where

$$I_{p,q}(x,y,z) = \int x^{q-1} \frac{K_{p+\frac{1}{2}}(b\sqrt{xy+z})}{(\sqrt{xy+z})^{p+\frac{1}{2}}} dx. \quad (13)$$

Using the results of Appendix I, $F_\gamma(\gamma)$ can be written as

$$F_\gamma(\gamma) = -A\Gamma(m) \sum_{k=1}^{m} \frac{2^k \gamma^{m-k}}{(\beta b)^k (m-k)!} \frac{K_{m-k+\frac{1}{2}}(b\sqrt{\alpha+\beta\gamma})}{\left(\sqrt{\alpha+\beta\gamma}\right)^{m+\frac{1}{2}-k}}. \quad (14)$$

Consequently, the outage probability is given by

$$P_{\text{out}} = 1 - A\Gamma(m) \sum_{k=1}^{m} \frac{2^k \gamma_{\text{th}}^{m-k}}{(\beta b)^k (m-k)!} \frac{K_{m-k+\frac{1}{2}}(b\sqrt{\alpha+\beta\gamma_{\text{th}}})}{\left(\sqrt{\alpha+\beta\gamma_{\text{th}}}\right)^{m-k+\frac{1}{2}}}. \quad (15)$$

B. Average Bit Error Probability

The average bit error probability (BEP) constitutes an important performance measure of a digital communication system. Unfortunately, the average BEP is generally not easy to compute. However, it was shown in [2] that the MGF can play a key role in evaluating the average BEP for several kinds of modulation (with and without diversity in an uncorrelated environment). For instance, for differentially coherent detection of binary phase-shift-keying (DPSK) or noncoherent detection of binary orthogonal frequency-shift-keying (FSK), the average BEP can be written as [2]

$$P_b(E) = C_1 M(-a_1), \quad (16)$$

where $M(\cdot)$ is the MGF and C_1 and a_1 are constants that depend on the modulation.

The MGF is therefore a key tool that needs to be derived. In Appendix II, we prove that the MGF corresponding to the $\mathcal{G}$-distribution can be written as

$$M(-s) = 1 + m \sum_{k=0}^{m-1} \frac{(-1)^{k+1} C_{m-1}^k}{(k+1)!} \sum_{p=0}^{k} C_k^p \left(2\sqrt{\frac{\alpha s}{\beta}}\right)^{k+1-p} \times \Gamma[k+p+1] H_{-(k+p+1)}\left(\frac{b}{2}\sqrt{\frac{\beta}{s}} + \sqrt{\frac{s\alpha}{\beta}}\right), \quad (17)$$

where C_k^p is the binomial coefficient and $H_\nu(x)$ is the Hermite function of order ν and is given by [31]

$$H_\nu(x) = 2^\nu \sqrt{\pi} \left(\frac{{}_1F_1(-\frac{\nu}{2}; \frac{1}{2}; x^2)}{\Gamma(\frac{1-\nu}{2})} - \frac{2x\,{}_1F_1(\frac{1-\nu}{2}; \frac{3}{2}; x^2)}{\Gamma(-\frac{\nu}{2})} \right),$$

where ${}_1F_1(\cdot;\cdot;\cdot)$ is the confluent hypergeometric function of the first kind [30].

IV. Channel Capacity

Spectral efficiency of adaptive transmission techniques has attracted a rising concern in the last decade. This interest stems from the fact that Shannon's channel capacity represents the upper bound for the data rate achievable in a transmission with an arbitrary small error probability, and as such represents an important performance measure of communication systems. The evaluation of the capacity of fading channels mainly started with Lee's paper [18], in which he analyzed the capacity of Rayleigh fading channels under the optimal rate constant power policy. Since then, several results on wireless channel capacity became available. In [19], Alouini and Goldsmith extended the work of Lee by examining the capacity of Rayleigh fading channels under different adaptive transmission techniques and different configurations. Other fading channels like Rician, Hoyt, Nakagami, Weibull and $\mathcal{K}$ fading channels were studied in [20], [21], [22] and [15]. Here, we present analytical expressions for the capacity with different adaptive transmission techniques for the $\mathcal{G}$-distributed fading channels. Note finally that in this paper we assume that the receiver reacts to the combined effects of the fast and the slow fading. For an analysis of the spectral efficiency of adaptive modulations when the receiver reacts to the shadowing only, the interested reader is referred to [9] where a comparison between the performance of adaptive modulation techniques tracking shadowing with that tracking composite fading and shadowing is provided.

A. Optimal Rate Adaptation with Constant Transmit Power

Under the optimal rate constant power (ora) policy, only the receiver has access to the Channel State Information (CSI). The transmitter does not know the channel fade level and hence, regardless of the channel conditions, keep sending information with a constant power. The capacity for this scheme is known to be given by [23], [24]

$$< C >_{\text{ora}} = \int_0^{+\infty} \ln(1+\gamma) f_\gamma(\gamma) d\gamma = -\int_0^{+\infty} \frac{F_\gamma(\gamma)}{1+\gamma} d\gamma. \tag{18}$$

Substituting $F_\gamma(\gamma)$ with the expression obtained above, then using the change of variable $v = 1+\gamma$ and applying the binomial expansion, we obtain

$$\frac{< C >_{\text{ora}}}{A\Gamma(m)} = \sum_{k=1}^{m} \frac{2^k \int_0^{+\infty} \frac{\gamma^{m-k}}{1+\gamma} \frac{K_{m-k+\frac{1}{2}}(bh(\gamma))}{(h(\gamma))^{m-k+\frac{1}{2}}} d\gamma}{(\beta b)^k (m-k)!} \tag{19}$$

$$= \sum_{k=1}^{m} \frac{2^k}{(\beta b)^k (m-k)!} \sum_{j=0}^{m-k} C_{m-k}^{j} (-1)^{m-k-j} \times \int_1^{+\infty} \frac{v^{j-1} K_{m-k+\frac{1}{2}}(b\sqrt{\alpha-\beta+\beta v})}{\left(\sqrt{\alpha-\beta+\beta v}\right)^{m-k+\frac{1}{2}}} dv \tag{20}$$

$$= \sum_{k=1}^{m} \frac{2^k}{(\beta b)^k (m-k)!} \Big((-1)^{m-k} R_{m-k}(\beta, \alpha-\beta) - \sum_{j=1}^{m-k} C_{m-k}^{j} (-1)^{m-k-j} I_{m-k,j}(1,\beta,\alpha-\beta) \Big), \tag{21}$$

where $R_n(\cdot,\cdot)$ is given by

$$R_n(x,y) = \int_1^{+\infty} \frac{K_{n+\frac{1}{2}}(b\sqrt{xv+y})}{v\left(\sqrt{xv+y}\right)^{n+\frac{1}{2}}} dv, \tag{22}$$

and is computed in Appendix III.

B. Optimal Simultaneous Power and Rate Adaptation

For optimal power and rate adaptation (opra), the CSI is accessible to both the transmitter and the receiver. The transmitter therefore adapts both its power and rate according to the channel conditions, i.e., higher transmission power and rate for high SNR and lower power and rate for low SNR. The capacity for this scheme is known to be given by [24]

$$< C >_{\text{opra}} = \int_{\gamma_0}^{+\infty} \ln\left(\frac{\gamma}{\gamma_0}\right) f_\gamma(\gamma) d\gamma = -\int_{\gamma_0}^{+\infty} \frac{F_\gamma(\gamma)}{\gamma} d\gamma. \tag{23}$$

By replacing $F_\gamma(\gamma)$ with its expression, we obtain

$$\frac{< C >_{\text{opra}}}{A\Gamma(m)} = \sum_{k=1}^{m} \frac{2^k \int_{\gamma_0}^{+\infty} \gamma^{m-k-1} \frac{K_{m-k+\frac{1}{2}}(bh(\gamma))}{(h(\gamma))^{m+\frac{1}{2}-k}} d\gamma}{(\beta b)^k (m-k)!} \tag{24}$$

$$= -\sum_{k=1}^{m-1} \frac{2^k I_{m-k,m-k}(\gamma_0,\beta,\alpha)}{(\beta b)^k (m-k)!} + \frac{2^m}{(\beta b)^m} R_0(\gamma_0\beta, \alpha). \tag{25}$$

γ_0 is the optimal cutoff under which the channel conditions are judged to be unfavorable by the transmitter and so no data is sent. Note that this will result in an outage probability equal to $P[\gamma \le \gamma_0]$. The optimal cutoff is the solution of the following equation [24]

$$\int_{\gamma_0}^{+\infty} \left(\frac{1}{\gamma_0} - \frac{1}{\gamma}\right) f_\gamma(\gamma) d\gamma = 1, \tag{26}$$

which, for $m > 1$, amounts to

$$\frac{\gamma_0}{A\Gamma(m)} = \sum_{k=1}^{m} \frac{2^k \gamma_0^{m-k}}{(\beta b)^k (m-k)!} \frac{K_{m-k+\frac{1}{2}}(b\sqrt{\alpha+\beta\gamma_0})}{\left(\sqrt{\alpha+\beta\gamma_0}\right)^{m+\frac{1}{2}-k}} - \frac{1}{m-1} \times \sum_{k=1}^{m-1} \frac{2^k \gamma_0^{m-k}}{(\beta b)^k (m-1-k)!} \frac{K_{m-k+\frac{1}{2}}(b\sqrt{\alpha+\beta\gamma_0})}{\left(\sqrt{\alpha+\beta\gamma_0}\right)^{m+\frac{1}{2}-k}}. \tag{27}$$

The optimal cutoff γ_0 can not be solved in closed-form and the last equation has to be evaluated numerically. For $m = 1$, an equation involving $R_1(\gamma_0\beta, \alpha)$ can be easily obtained.

C. Channel Inversion with Fixed Rate

1) Total Channel Inversion: This is a suboptimal policy in which the transmitter exploits his knowledge of the CSI by preserving a constant received power. In other words the transmitter inverts the channel, i.e., higher transmitted power for low SNR and lower transmitted power for high SNR. Probably the main advantage of this scheme is its simple implementation. Indeed, from an encoder-decoder perspective, the channel becomes equivalent to an AWGN channel, and hence these components can be designed independently of the fading channel characteristics. The capacity for this policy is known to be given by [24]

$$< C >_{\text{cifr}} = \ln\left(1 + \frac{1}{\int_0^{+\infty} \frac{1}{\gamma} f_\gamma(\gamma) d\gamma}\right). \tag{28}$$

For $m \ge 2$, the integral in the logarithm can be expressed in terms of $I_{m,m-1}(x,y,z)$, yielding the following expression

$$< C >_{\text{cifr}} = \ln\left(1 - \frac{1}{A I_{m,m-1}(0,\beta,\alpha)}\right) = \ln\left(1 + \overline{\gamma}\frac{(m-1)}{m}\frac{\lambda}{\lambda+\theta}\right). \tag{29}$$

Note that, for Rayleigh fading, i.e. $m = 1$, the capacity with total channel inversion tends to zero. This is because the integral inside the logarithm will diverge.

$$< C >_{\text{tcifr}} = A\Gamma(m)\ln\left(1+\frac{m-1}{A\Gamma(m)\sum_{k=1}^{m-1}\frac{2^k\gamma_0^{m-1-k}}{(\beta b)^k(m-1-k)!}\frac{K_{m-k+\frac{1}{2}}(b\sqrt{\alpha+\beta\gamma_0})}{\sqrt{\alpha+\beta\gamma_0}^{\,m-k+\frac{1}{2}}}}\right)\sum_{k=1}^{m}\frac{2^k\gamma_0^{m-k}}{(\beta b)^k(m-k)!}\frac{K_{m-k+\frac{1}{2}}(b\sqrt{\alpha+\beta\gamma_0})}{\left(\sqrt{\alpha+\beta\gamma_0}\right)^{m+\frac{1}{2}-k}}. \quad (31)$$

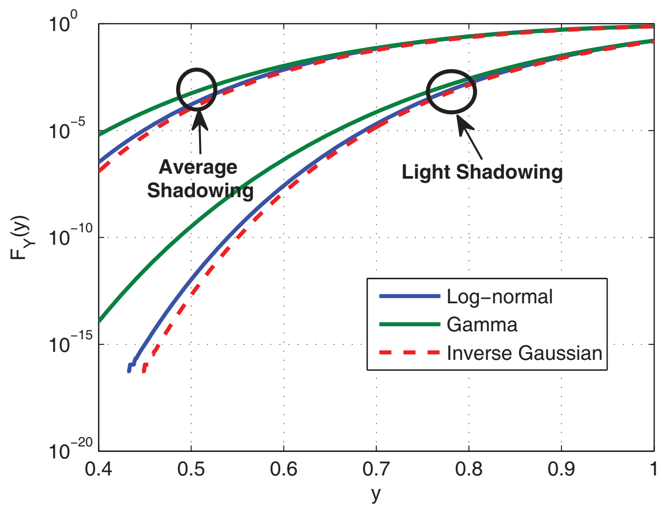

Fig. 1. Cdf of the lognormal, the inverse-Gaussian and the gamma distribution for infrequent light shadowing ($\mu = 0.115$ and $\sigma = 0.115$) and average shadowing ($\mu = -0.115$ and $\sigma = 0.161$).

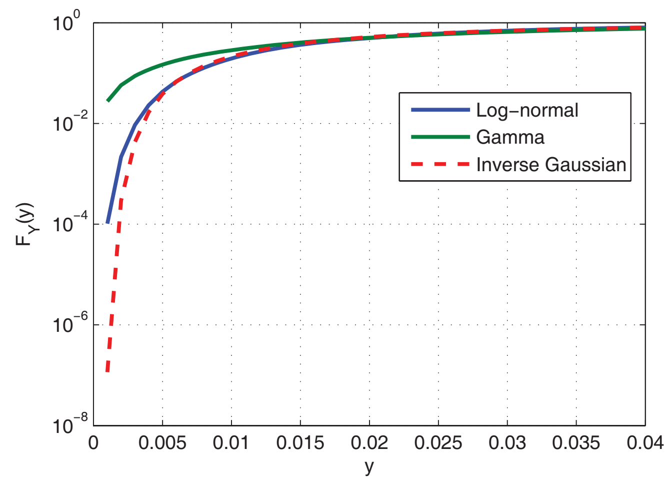

Fig. 2. Cdf of the lognormal, the inverse-Gaussian, and the gamma distribution for frequent heavy shadowing ($\mu = -3.914$ and $\sigma = 0.806$).

2) Truncated Channel Inversion: Despite its simplicity, total channel inversion can suffer a large capacity penalty for certain types of fading. A solution to this problem would be to suspend the transmission when the SNR falls below a certain threshold, say γ_0, and to resume transmission when the SNR is larger than γ_0. Note, however, that this solution is not perfect since the system will suffer from an outage probability equal to $P[\gamma \leq \gamma_0]$. This policy is called truncated channel inversion and its capacity is given by [24]

$$< C >_{\text{tcifr}} = \ln\left(1+\frac{1}{\int_{\gamma_0}^{+\infty}\frac{1}{\gamma}f_\gamma(\gamma)d\gamma}\right)(1-P_{\text{out}}), \quad (30)$$

which is readily obtained using the previously derived results. For instance, if $m > 1$, the capacity is given by (31) at the bottom of this page. For $m = 1$, the expression for the capacity will involve $R_1(\gamma_0\beta, \alpha)$.

V. Numerical Results

Figs. 1 and 2 show the Cumulative Distribution Functions (CDF) of the log-normal, the Inverse-Gaussian, and the Gamma distributions for three different shadowing scenarios. The shadowing environments that we consider were introduced by Loo in [25] and [26] and are referred to as infrequent light shadowing ($\mu = 0.115$ and $\sigma = 0.115$, this corresponds to a sparse tree cover), average shadowing ($\mu = -0.115$ and $\sigma = 0.161$, this corresponds to an average tree cover) and frequent heavy shadowing ($\mu = -3.914$ and $\sigma = 0.806$, this corresponds to a dense tree cover). These models were widely used in the literature like for instance in [27]-[29]. For more details about these models, the interested reader is referred to [3], [25] and [26]. Figs. 1 and 2 show clearly that the IG-distribution can be used as a precise substitute to the log-normal one for several shadowing environments. In addition, it can be seen that the IG pdf is a better substitute for the log-normal distribution than the Gamma pdf is. This is especially true for the frequent heavy shadowing environment. Note also that the Kullback-Leibler distance (also called the relative entropy) can be used to compare these distributions, as in [16] which confirms the results obtained in this paper.

In the next examples, we compare the analytical expressions that we have developed (referred to in the figures as analytical formulae) with numerical integration (referred to as simulation). Numerical integration refers to the approximate computation of integrals using quadratures techniques which are available in mathematical packages. The integrals that were evaluated numerically here are given by (11), (16), (18), (23), (28) and (30), where $f_\gamma(\gamma)$ was replaced by (6). The comparison is conducted for two shadowing scenarios, namely the average shadowing environment and the frequent heavy shadowing one. Throughout our simulations the parameter m is arbitrarily set to 5. Fig. 3 depicts the outage probability versus the average SNR $\overline{\gamma}$ for $\gamma_0 = 5$ dB. As expected, the outage probability increases as the shadowing becomes more pronounced. The average bit error probability for DPSK ($C_1 = \frac{1}{2}$ and $a_1 = 1$ in (16)) and noncoherent frequency shift keying ($C_1 = \frac{1}{2}$ and $a_1 = \frac{1}{2}$ in (16)) is also illustrated in Fig. 4. Here also the performance is degraded in the frequent heavy shadowing environment.

Figs. 5 and 6 depict the capacity of the different adaptation policies. Note the adequacy between the capacity given by

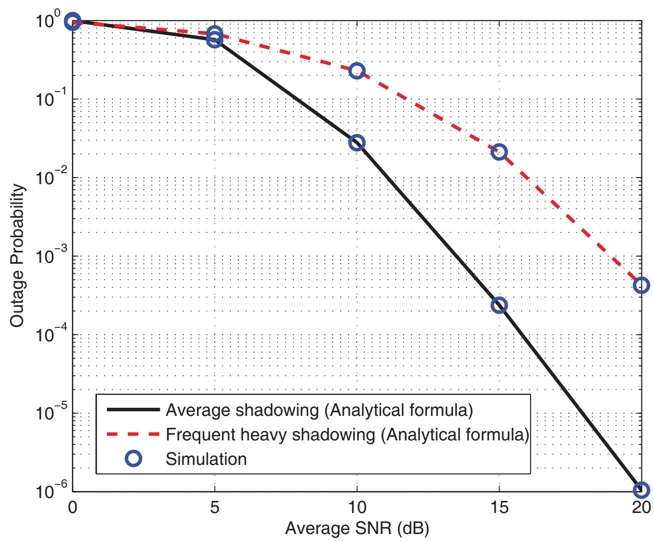

Fig. 3. Outage probability for $\gamma_{\text{th}} = 5$ dB in average ($\mu = -0.115$ and $\sigma = 0.161$) and frequent heavy shadowing environments ($\mu = -3.914$ and $\sigma = 0.806$).

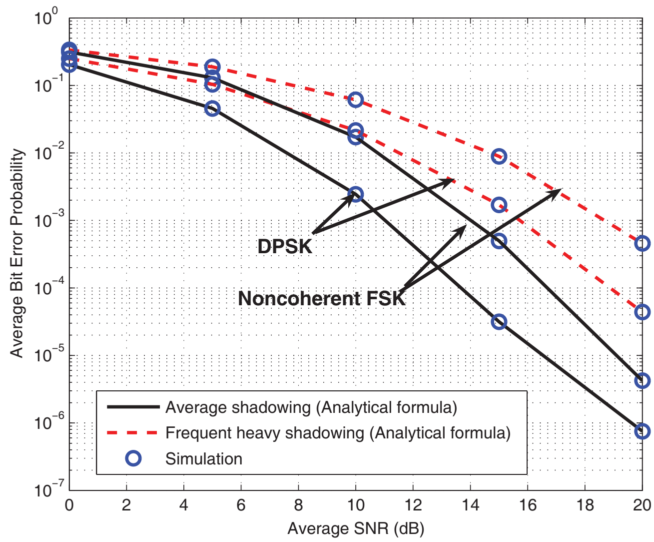

Fig. 4. Average bit error probability for DPSK and noncoherent FSK in average ($\mu = -0.115$ and $\sigma = 0.161$) and frequent heavy shadowing environments ($\mu = -3.914$ and $\sigma = 0.806$).

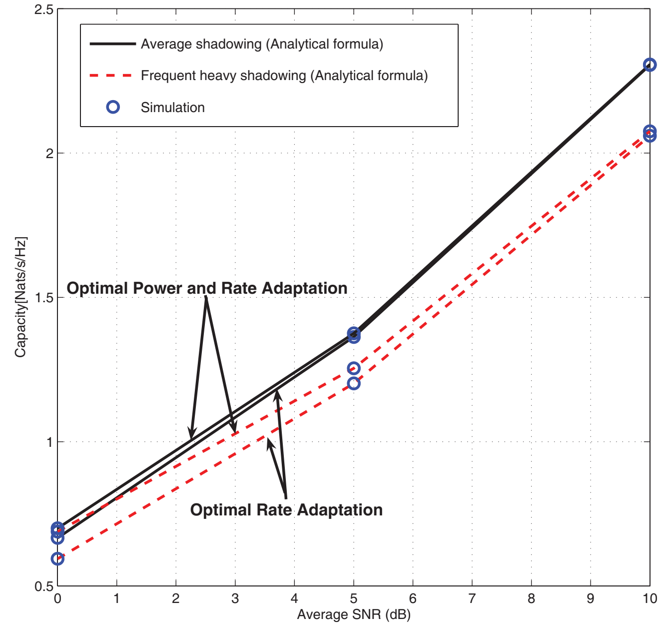

Fig. 5. Capacity for ORA and OPRA policies versus SNR in average ($\mu = -0.115$ and $\sigma = 0.161$) and frequent heavy shadowing environments ($\mu = -3.914$ and $\sigma = 0.806$).

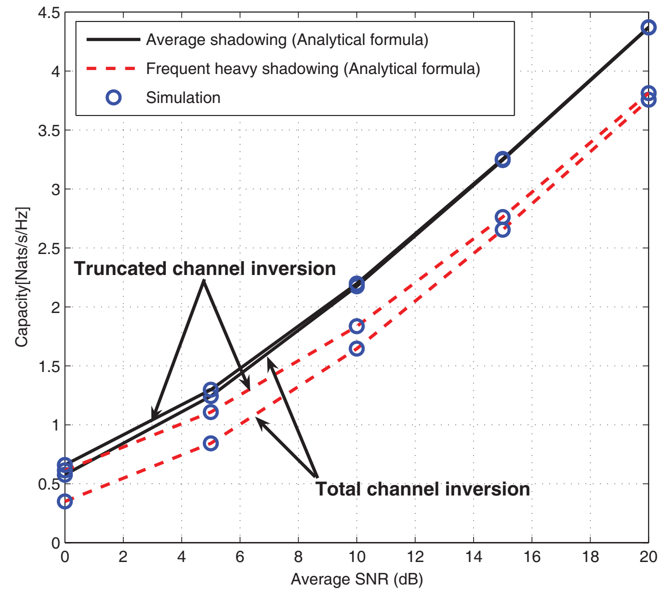

Fig. 6. Capacity for total channel inversion and truncated channel inversion policies versus SNR in average ($\mu = -0.115$ and $\sigma = 0.161$) and frequent heavy shadowing environments ($\mu = -3.914$ and $\sigma = 0.806$).

the theoretical formulae and the one obtained by simulations. Fig. 5 shows also that compared to optimal power and rate adaptation, transmission with optimal rate adaptation suffers capacity penalty at low SNR only. However, as $\overline{\gamma}$ increases, the two policies will provide the same capacity. Therefore, the information provided by the CSI at the transmitter has only a small impact on the capacity. As illustrated in Fig. 6, this comparison holds also for truncated channel inversion and total channel inversion. Finally, Fig. 7 shows the dependence of the capacity with truncated channel inversion on γ_0 for different $\overline{\gamma}$ values. It can be seen that the capacity is maximized for an optimal γ_0^* which increases as $\overline{\gamma}$ increases.

VI. Conclusion

Composite multipath fading/shadowing environments are frequently encountered in several realistic scenarios. In this paper, we have proposed to use the Nakagami-Inverse Gaussian composite model as a substitute for the log-normally shadowed Nakagami fading. The resulting distribution (the $\mathcal{G}$-distribution) has an analytical expression that facilitates the performance evaluation of single-user communication systems over composite channels. In this study, several key results have been presented, including the outage probability, average error probabilities, and the capacity for different adaptive transmission techniques. The expressions that we have provided can be used to assess the performance of communication systems over composite channels.

$$M(-s) = 1 - \frac{1}{\Gamma(m)}\sqrt{\frac{\pi\alpha}{\beta}}\frac{d^{m-1}}{ds^{m-1}}\left(s^{m-\frac{1}{2}}\,\mathrm{erfcx}\left(\frac{b}{2}\sqrt{\frac{\beta}{s}}+\sqrt{\frac{s\alpha}{\beta}}\right)\right), \tag{47}$$

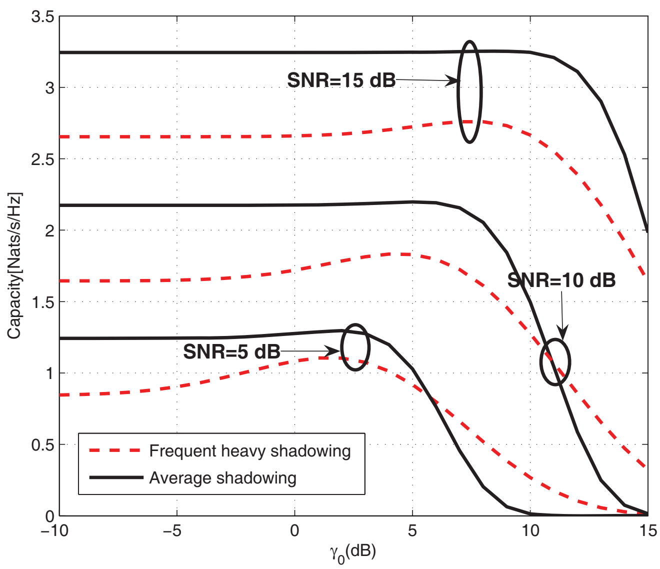

Fig. 7. Capacity for truncated channel inversion versus γ_0 in average ($\mu = -0.115$ and $\sigma = 0.161$) and frequent heavy shadowing environments ($\mu = -3.914$ and $\sigma = 0.806$).

Appendix A
Evaluation of $I_{p,q}(x,y,z)$

Define $I_{p,q}(x,y,z)$ as

$$I_{p,q}(x,y,z) = \int x^{q-1}\frac{K_{p+\frac{1}{2}}(b\sqrt{xy+z})}{(\sqrt{xy+z})^{p+\frac{1}{2}}}dx. \tag{32}$$

Using the fact that [30, Eq. 8.486.15]

$$\frac{d(x^{-\mu}K_\mu(x))}{dx} = -x^{-\mu}K_{\mu+1}(x), \tag{33}$$

and integrating by part we obtain

$$\begin{aligned} I_{p,q}(x,y,z) &= -\frac{2x^{q-1}}{yb}\frac{K_{p-\frac{1}{2}}(b\sqrt{xy+z})}{(\sqrt{xy+z})^{p-\frac{1}{2}}} \\ &+ \frac{2(q-1)}{yb}I_{q-1,p-1}(x,y,z). \end{aligned} \tag{34}$$

Iterating over this equation, we finally find that

$$\frac{I_{p,q}(x,y,z)}{(q-1)!} = -\sum_{k=1}^{q}\frac{2^k x^{q-k}}{(yb)^k(q-k)!}\frac{K_{p-k+\frac{1}{2}}(b\sqrt{xy+z})}{(\sqrt{xy+z})^{p+\frac{1}{2}-k}}. \tag{35}$$

Appendix B
Derivation of the Moment Generating Function

By definition, we have

$$\begin{aligned} M(-s) &= E_\gamma[e^{-s\gamma}] & (36)\\ &= A\int_0^{+\infty} e^{-s\gamma}\gamma^{m-1}\frac{K_{m+\frac{1}{2}}\left(b\sqrt{\alpha+\beta\gamma}\right)}{\left(\sqrt{\alpha+\beta\gamma}\right)^{m+\frac{1}{2}}}d\gamma & (37)\\ &= (-1)^{m-1}A\frac{d^{m-1}G_m(s)}{ds^{m-1}}. & (38) \end{aligned}$$

where $G_m(s)$ is defined as follows

$$G_m(s) = \int_0^{+\infty}\exp(-s\gamma)\frac{K_{m+\frac{1}{2}}\left(b\sqrt{\alpha+\beta\gamma}\right)}{\left(\sqrt{\alpha+\beta\gamma}\right)^{m+\frac{1}{2}}}d\gamma. \tag{39}$$

Applying an integration by part, we find that $G_m(s)$ satisfies the following recursion formula

$$G_m(s) = \frac{2}{b\beta}\frac{K_{m-\frac{1}{2}}(b\sqrt{\alpha})}{(\sqrt{\alpha})^{m-\frac{1}{2}}} - \frac{2s}{b\beta}G_{m-1}(s). \tag{40}$$

Iterating over this equation, we obtain

$$\begin{aligned} G_m(s) &= \sum_{k=1}^{m}(-1)^{k-1}\frac{2^k s^{k-1}}{(b\beta)^k}\frac{K_{m-k+\frac{1}{2}}(b\sqrt{\alpha})}{(\sqrt{\alpha})^{m-k+\frac{1}{2}}} \\ &+ (-1)^m\frac{2^m s^m}{(b\beta)^m}G_0(s), \end{aligned} \tag{41}$$

where $G_0(s)$ is given by

$$G_0(s) = \sqrt{\frac{\pi}{2b}}\int_0^{+\infty}\exp(-s\gamma)\frac{e^{-b\sqrt{\beta\gamma+\alpha}}}{\sqrt{\beta\gamma+\alpha}}d\gamma. \tag{42}$$

By plugging the expression of $G_m(s)$ in $M(-s)$, the latter will be given by

$$M(-s) = 1 - \frac{\Gamma(m+1)A2^m}{(b\beta)^m}\sum_{k=0}^{m-1}C_{m-1}^k\frac{s^{k+1}}{(k+1)!}G_0^{(k)}(s). \tag{43}$$

The kth derivative of G_0 can be calculated as follows

$$G_0^{(k)}(s) = (-1)^k\sqrt{\frac{\pi}{2b}}\int_0^{+\infty}\gamma^k\exp(-s\gamma)\frac{e^{-b\sqrt{\beta\gamma+\alpha}}}{\sqrt{\beta\gamma+\alpha}}d\gamma, \tag{44}$$

which, after some manipulations, leads to

$$\begin{aligned} G_0^{(k)}(s) &= \frac{(-1)^k}{\beta^{k+1}}\sqrt{\frac{2\pi}{b}}e^{-b\sqrt{\alpha}}\sum_{p=0}^{k}C_k^p(2\sqrt{\alpha})^{k-p} \\ &\times \int_0^{+\infty}y^{k+p}e^{-\frac{sy^2}{\beta}}e^{-y(b+\frac{2s\sqrt{\alpha}}{\beta})}dy. \end{aligned} \tag{45}$$

Using [30, Eq. (3.462.1)] with [31], the last equation can be written as

$$\begin{aligned} G_0^{(k)}(s) &= \frac{(-1)^k}{\beta^{k+1}}\sqrt{\frac{2\pi}{b}}e^{-b\sqrt{\alpha}}\sum_{p=0}^{k}C_k^p(2\sqrt{\alpha})^{k-p}\left(\frac{s}{\beta}\right)^{-\frac{k+p+1}{2}} \\ &\times \Gamma[k+p+1]H_{-(k+p+1)}\left(\frac{b}{2}\sqrt{\frac{\beta}{s}}+\sqrt{\frac{s\alpha}{\beta}}\right), \end{aligned} \tag{46}$$

where $H_\nu(x)$ is the Hermite function. Note that the $M(-s)$ can be also expressed in a more compact form[4] as given in (47), where $\mathrm{erfcx}(\cdot)$ is the scaled complementary error function and is given by

$$\mathrm{erfcx}(x) = e^{x^2}\mathrm{erfc}(x). \tag{48}$$

[4]Note that even though this expression seems simpler than (17), (17) does not involve successive derivatives and thus has the advantage of being directly computable using mathematical software packages.

APPENDIX C
EVALUATION OF $R_n(x, y)$

Define $R_n(x, y)$ as

$$R_n(x,y) = \int_1^{+\infty} \frac{K_{n+\frac{1}{2}}(b\sqrt{xv+y})}{v\left(\sqrt{xv+y}\right)^{n+\frac{1}{2}}} dv. \quad (49)$$

Using the change of variable $X = \sqrt{xv+y}$, we obtain

$$R_n(x,y) = 2\int_{\sqrt{x+y}}^{+\infty} \frac{K_{n+\frac{1}{2}}(bX)}{(X^2-y)X^{n-\frac{1}{2}}} dX. \quad (50)$$

Using the fact that $K_{n+\frac{1}{2}}(bX)$ can be written as

$$K_{n+\frac{1}{2}}(bX) = \sqrt{\frac{\pi}{2bX}} \exp(-bX) \sum_{l=0}^{n} \frac{\Gamma(n+1+l)(2bX)^{-l}}{\Gamma(n+1-l)\Gamma(l+1)}, \quad (51)$$

we obtain

$$R_n(x,y) = \sqrt{\frac{2\pi}{b}} \sum_{l=0}^{n} \frac{\Gamma(n+1+l)(2b)^{-l}}{\Gamma(n+1-l)\Gamma(l+1)} T_{n+l}(\sqrt{y}, \sqrt{x+y}), \quad (52)$$

where

$$T_l(a,c) = \int_c^{+\infty} \frac{\exp(-bX)}{(X^2-a^2)X^l} dX. \quad (53)$$

This integral can be solved by applying partial fraction decomposition

$$\frac{a^{l+1}}{(X^2-a^2)X^l} = \frac{1}{2}\left[\frac{1}{X-a} - \frac{(-1)^l}{X+a}\right] - \sum_{j=1}^{\frac{l+t}{2}} \frac{1}{X^{2j-t}a^{-2j+t+1}}, \quad (54)$$

where $t = \frac{1-(-1)^l}{2}$. Here two cases must be distinguished:

- $a \neq 0$: in this case, $T_l(a, c)$ will be given by

$$T_l(a,c) = \frac{e^{-ab}\Gamma[0, b(c-a)] - (-1)^l e^{ab}\Gamma[0, b(c+a)]}{2a^{l+1}} - \sum_{j=1}^{\frac{l+t}{2}} \frac{E_{2j-t}(bc)}{c^{2j-t-1}a^{l-2j+t+2}}, \quad (55)$$

- $a = 0$: and in this case, $T_l(0, c)$ reduces to

$$T_l(0,c) = \frac{E_{l+2}(bc)}{c^{l+1}}, \quad (56)$$

where $\Gamma(\cdot,\cdot)$ is the incomplete Gamma function [30] and $E_n(z) = \int_1^{+\infty} \frac{e^{-zt}}{t^n} dt$ is the nth order exponential integral function [30].

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