Direct-Sequence Spread Spectrum in a Shadowed Rician Fading Land-Mobile Satellite Channel

Richard D. J. van Nee, Howard S. Misser, *Member, IEEE,* and Ramjee Prasad, *Senior Member, IEEE*

***Abstract*—The performance of a direct sequence spread-spectrum land-mobile satellite transmission system, using binary phase shift keying (BPSK) modulation, is analyzed. The satellite channel is modeled as having shadowed Rician fading characteristics. The bit error probability is evaluated, considering both the envelope and the phase variation. Assuming a Gaussian approximation for the interference, numerical results are obtained for both spread-spectrum and narrowband land-mobile satellite communication systems with BPSK modulation. A comparison of the two systems is made for light, average, and heavy shadowing.**

I. Introduction

THE application of spread-spectrum modulation in the field of land-mobile satellite communications offers code division multiple access (CDMA), resistance to multipath fading, and low peak-to-average power ratio. In addition, the properties of low probability of interception (LPI), antijam resistance, and message privacy and security are attractive in some applications. Furthermore, as shown in [1] and [2], spread-spectrum CDMA systems can provide greater capacity than FDMA for mobile satellite communications. In general, land-mobile satellite systems allow a wide range of services, including voice, data, position-finding, and paging services, interconnection to the public switched telephone network, and the possibility of private networks [3]. Accordingly, numerous research papers (e.g., [1]–[10]) have been published recently to study the land–mobile satellite communication channel and its effects on such systems.

This paper presents a performance analysis of land–mobile satellite communications using direct sequence spread-spectrum with BPSK modulation, in terms of bit error, outage, and message success probability. The channel model adopted in the analysis is characterized by the combined effect of Rician fading and lognormal shadowing [4]–[6]. Section II formulates and extends the channel model to the use of spread-spectrum modulation. The receiver model and performance analysis follow in Sections III and IV, respectively. Numerical results are presented in Section V. Finally, our conclusions are given in Section VI.

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R. D. J. van Nee and R. Prasad are with Telecommunications and Traffic Control Systems Group, Delft University of Technology, 2600 GA Delft, The Netherlands.

H. S. Misser is currently with PTT Research Dr. Neher Laboratories, Leidschendam, The Netherlands.

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II. Channel Model

A statistical propagation model for a narrowband channel in rural and suburban environments was developed in [4]–[6], assuming that the line-of-sight signal strength is lognormally distributed and the composite multipath signal is Rayleigh distributed. The resulting probability distribution of the received signal envelope r is given by:

$$p_\beta(r) = \frac{r}{b_o\sqrt{2\pi d_o}} \int_0^\infty \exp\left(-\frac{(\ln(z) - \mu_o)^2}{2d_o} - \frac{(r^2 + z^2)}{2b_o}\right) \frac{I_o(rz/b_o)}{z}\, dz \tag{1}$$

where $I_n(\cdot)$ is the modified Bessel function of the first kind and nth order, b_o is the average scattered power due to multipath, μ_o is the mean value due to shadowing, and d_o is the variance due to shadowing.

The probability density function of the received signal phase ϕ was found to be approximately Gaussian [6]:

$$p_\phi(\phi) = \frac{1}{\sqrt{2\pi\sigma_\phi^2}} \exp\left[-\frac{(\phi - \mu_\phi)^2}{2\sigma_\phi^2}\right] \tag{2}$$

where μ_ϕ and σ_ϕ^2 are the mean and variance of the received signal phase, respectively.

The above model is valid for a narrowband system. If spread-spectrum modulation is used with a chip duration less than the delay spread of the channel, the multipath power is partially reduced by the correlation operation in the receiver. The envelope and the phase distribution functions remain the same, but the values for b_o and σ_ϕ are reduced.

The impulse response of the channel can be written as:

$$h(t) = \sum_{m=1}^{M} \beta_m\, \delta(t - \tau_m) e^{j\theta_m} \tag{3}$$

where β, τ, and θ are the gain, time delay, and phase of the mth path, respectively. The first path is the line-of-sight and therefore its propagation statistics are described by (1) and (2). The other paths have a Rayleigh path gain distribution and uniformly distributed phase, since the direct line-of-sight is suppressed by the correlation operation. The parameters of the various path distributions can be found if the power-delay profile is known. For the present study, the power-delay profile is

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considered as:

$$P(\tau) = cb_o \exp(-c\tau) \tag{4}$$

where $1/c$ is the delay spread. For a rural environment, a typical value of $1/c$ is 0.65 μs [7].

Due to the correlation operation, the multipath power ($= b_o$) is reduced. For path m, it can be approximated as:

$$b_{mo} = b_o[1 - \exp(-cT_c)] \exp[-c(m-1)T_c] \tag{5}$$

where T_c is the chip duration.

Also, the phase variance of the first path will be decreased, since it is determined by the amount of multipath power and the statistics of the line-of-sight propagation. Using (4.5-19) of [11], the phase distribution function $p(\phi)$ of a lognormally shadowed Rician signal can be derived and given as:

$$p(\phi) = \frac{1}{\sqrt{8\pi^3 d_o}} \int_0^\infty \exp\left(\frac{-z^2}{2b_o} - \frac{(\ln(z) - \mu_o)^2}{2d_o}\right) \cdot \frac{(1 + G\sqrt{\pi}\exp(G^2)[1 + \operatorname{erf}(G)])}{z}\, dz \tag{6}$$

where

$$G \triangleq \frac{z\cos(\phi)}{\sqrt{2b_o}}$$

and erf $(\cdot)$ is the error function [12]: erf $(G) \triangleq 2/\sqrt{\pi} \int_0^G \exp(-t^2)\, dt$. Now, the phase variance in the case of spread-spectrum modulation can be determined by:

$$\sigma_\phi^2 = \int_{-\pi}^{\pi} \phi^2 p(\phi)\, d\phi. \tag{7}$$

Note that the mean phase μ_ϕ is zero because $p(\phi) = p(-\phi)$.

Equations (6) and (7) are used only to determine the effect of spread-spectrum modulation on the phase variance. The combined effect of phase variation and envelope fading [6] is evaluated by using (2) as an approximation for (6).

III. Receiver Model

The spread-spectrum receiver model is shown in Fig. 1. The total received signal is:

$$r(t) = \sum_{k=1}^{K} \sum_{m=1}^{M} A\beta_{mk} a_k(t - \tau_{mk})\, b_k(t - \tau_{mk}) \cdot \cos((\omega_c + \omega_{mk})t + \phi_{mk}) + n(t) \tag{8}$$

where m and k denote the path and user number, respectively, and A is the transmitted signal amplitude, which is assumed to be constant and identical for all users. For user k, $\{a_k\}$ is the spread-spectrum code, $\{b_k\}$ is the data sequence, $\omega_c + \omega_{mk}$ is the carrier plus doppler angular frequency, ϕ_{mk} is the carrier phase, and $n(t)$ is white Gaussian noise with two-sided power spectral density $N_o/2$.

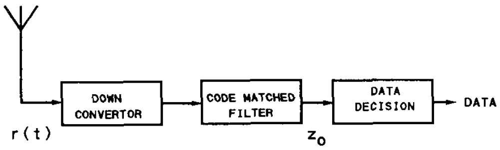

Fig. 1. Block diagram of the spread-spectrum receiver.

This signal is converted to baseband and correlated with a particular user code. If the receiver locks on the first path of user one, a signal sample of the correlation output can be written as:

$$z_o = A\beta_{11}\cos\phi_{11} T_b b_1^0 + \sum_{k=1}^{K} A(b_k^{-1}X_k + b_k^0\hat{X}_k) + \eta_1 \tag{9}$$

where b_k^{-1} and b_k^0 are the previous and current data bit, respectively, and

$$X_1 = \sum_{m=2}^{M} R_{11}(\tau_{m1})\beta_{m1}\cos\phi_{m1}$$

$$\hat{X}_1 = \sum_{m=2}^{M} \hat{R}_{11}(\tau_{m1})\beta_{m1}\cos\phi_{m1}$$

$$X_k = \sum_{m=1}^{M} R_{1k}(\tau_{mk})\beta_{mk}\cos\phi_{mk}$$

$$\hat{X}_k = \sum_{m=1}^{M} \hat{R}_{1k}(\tau_{mk})\beta_{mk}\cos\phi_{mk}$$

$$R_{1k}(\tau) = \int_0^\tau a_k(t-\tau)a_1(t)\, dt$$

$$\hat{R}_{1k}(\tau) = \int_\tau^{T_b} a_k(t-\tau)a_1(t)\, dt.$$

In (9), the carrier phase ϕ_{mk} is Gaussian distributed for $m = k = 1$, but uniformly distributed for the other users and paths, because the transmitters are assumed to have arbitrary phases. β_{mk} has a shadowed Rician distribution for $m = 1$, and a Rayleigh distribution otherwise. η_1 is a zero-mean Gaussian variable with variance $N_o T_b$.

IV. Performance Analysis

The bit error probability and outage probability are considered to be the two basic performance measures of digital systems. The message success probability is also an important consideration. All three performance measures are discussed here.

A. Bit Error Probability

Assuming that the data bits -1 and 1 are equiprobable, the bit error probability p_e can be expressed as:

$$p_e = P(z_o < 0 | b_1^0 = 1). \tag{10}$$

The interference can be approximated by Gaussian noise if KM is large, so p_e can be written as:

$$p_e = \int_{-\infty}^{\infty} p_e(x) p_u(x)\, dx. \tag{11}$$

Here,

$$p_e(x) = P\,(\text{Gaussian noise} < -x) = \frac{1}{2}\,\text{erfc}\left(\frac{x}{\sigma\sqrt{2}}\right). \tag{12}$$

erfc $(\cdot)$ is the complementary error function [12], p_u is the pdf of the first term of z_o, and σ^2 is defined as the total variance of the Gaussian noise:

$$\sigma^2 \triangleq N_o T_b + \sigma_i^2 \tag{13}$$

with σ_i^2 as the interference power.

1) Gaussian Approximation: Using the Gaussian approximation, only the mean and variance of the interference term $\Sigma_{k=1}^{K} A(b_k^{-1} X_k + b_k^0 \hat{X}_k)$ need to be evaluated. Since all terms of the summation are independent and symmetrically distributed, the mean reduces to zero. For the same reason, all the cross terms in the calculation of the second moment become zero, so the variance is:

$$\begin{aligned} \sigma_i^2 = & \sum_{m=2}^{M} A^2 E([b_1^{-1} R_{11}(\tau_{m1}) \\ & + b_1^0 \hat{R}_{11}(\tau_{m1})]^2) E([\beta_{m1} \cos \phi_{m1}]^2) \\ & + \sum_{k=2}^{K} \sum_{m=1}^{M} A^2 E([b_k^{-1} R_{1k}(\tau_{mk}) \\ & + b_k^0 \hat{R}_{1k}(\tau_{mk})]^2) E([\beta_{mk} \cos \phi_{mk}]^2). \end{aligned} \tag{14}$$

Note that the first and second term of the right-hand side of (14) represent the variance for $k = 1$ and $k > 1$, respectively.

The second moment of $\beta \cos \phi$ can be calculated as follows: If $m = 1$, then β has a shadowed Rician distribution, which can be viewed as a Rician distribution with a variable Rician parameter $s^2/2b_o$. The product of this Rician variable with the cosine of a uniformly distributed variable gives a Gaussian variable with a mean of $s/\sqrt{2}$ and a variance of b_o. Thus $E((\beta_{mk} \cos \phi_{mk})^2) = b_o + s^2/2$, on the condition that s is constant. This condition can be removed by integrating over the independent lognormal distribution of s:

$$E((\beta_{1k} \cos \phi_{1k})^2) = \int_0^{\infty} (b_{1o} + s^2/2) p(s)\, ds. \tag{15}$$

After integration, (15) reduces to:

$$E((\beta_{1k} \cos \phi_{1k})^2) = b_{1o} + \tfrac{1}{2} \exp\,(2d_o + 2\mu_o). \tag{16}$$

If $m > 1$, then β has a Rayleigh distribution. In that case, $s = 0$ and:

$$E((\beta_{mk} \cos \phi_{mk})^2) = b_{mo}. \tag{17}$$

Referring to [14], one obtains the second moment of the cross correlation for any user k and for any spread-spectrum code:

$$\begin{aligned} & E[R_{1k}^{2q}(\tau_{1k}) \hat{R}_{1k}^{2(p-q)}(\tau_{1k})] \\ & = \frac{T_c^{2p+1}}{T_b} \sum_{n=0}^{N-1} \sum_{i=1}^{2q} \frac{(-1)_i}{i+1} \frac{\binom{2q}{i}}{\binom{2(p-q)+i+1}{i+1}} \frac{B_{n1k}^i}{\hat{B}_{n1k}^{i+1}} \\ & \quad [(A_{n1k} + B_{n1k})^{2q-i} (\hat{A}_{n1k} + \hat{B}_{n1k})^{2(p-q)+i+1} \\ & \quad - A_{n1k}^{2q-i} \hat{A}_{n1k}^{2(p-q)+i+1}] \end{aligned} \tag{18}$$

where

$$A_{n1k} = C_{1k}(n - N) \qquad \hat{A}_{n1k} = C_{1k}(n)$$

$$B_{n1k} = C_{1k}(n + 1 - N) - C_{1k}(n - N)$$

$$\hat{B}_{n1k} = C_{1k}(n + 1) - C_{1k}(n)$$

$$C_{1k}(n) = \begin{cases} \sum_{j=0}^{N-1+n} a_k^j a_1^{j+n} & 0 \le n \le N - 1 \\ \sum_{j=0}^{N-1+n} a_k^{j-n} a_1^j & 1 - N \le n < 0 \end{cases}$$

and p and q are dummy variables.

For Gold codes, a simplified technique for evaluating the variance of the cross correlations can be used using [13], [16]. Now, one gets:

$$E([b_k^{-1} R_{1k}(\tau_{mk}) + b_k^0 \hat{R}_{1k}(\tau_{mk})]^2) = 2T_b^2/3N \tag{19}$$

where N is the Gold code length, which is assumed here to be equal to T_b/T_c. Thus, (19) simplifies the computation of (14) for Gold codes.

2) Distribution of the Wanted Signal: The first term of z_o in (9) consists of β with a shadowed Rician distribution, multiplied by the cosine of a Gaussian distributed phase. Using [11], the total distribution function of β cos (ϕ) can be written as:

$$p_u(x) = 2 \int_{\text{abs}(x)}^{\infty} p_\beta(r) p_\phi\,(\arccos\,(x/r)) \frac{dr}{\sqrt{r^2 - x^2}}. \tag{20}$$

Substituting (1) and (2) in (20) and using (11) and (12), p_e is expressed as:

$$p_e = 0.5 \int_0^{\infty} \text{erfc}\left(\frac{xT_b}{\sigma\sqrt{2}}\right) p_u(x)\, dx \tag{21}$$

with

$$\begin{aligned} p_u(x) = & \int_{\text{abs}(x)}^{\infty} \int_0^{\infty} \frac{r}{b_o \sqrt{2\pi^2 d_o \sigma_\phi^2}} \exp\left(-\frac{(\ln\,(z) - \mu_o)^2}{2d_o} \right. \\ & \left. - \frac{(r^2 + z^2)}{2b_o} - \frac{\arccos^2\,(x/r)}{2\sigma_\phi^2}\right) \\ & \cdot \frac{I_o(rz/b_o)}{z\sqrt{r^2 - x^2}}\, dz\, dr. \end{aligned} \tag{22}$$

Equation (22) is valid in the case that the receiver locks on the line-of-sight signal. This requires the bandwidth of the carrier tracking loop to be much smaller than the fading bandwidth of the received signal. If the bandwidth of the tracking loop is larger than the fading bandwidth, ϕ_{11} is approximately zero because the receiver will lock on the phase of the total signal. In that case, the distribution function p_u is equal to the shadowed Rician distribution p_β. However, with a larger tracking bandwidth the loop noise increases, so there is always a certain phase error. Therefore, in the bit error probability calculations one gets an upper bound by using (22), and a lower bound by using $p_u = p_\beta$.

B. Outage Probability

The outage probability P_{out} is the probability that the instantaneous bit error probability exceeds a certain threshold, denoted by p_o, and can be written as:

$$P_{\text{out}} = P(p_e \geq p_o) = P(x \leq x_o) = \int_{-\infty}^{x0} x p_u(x)\, dx. \tag{23}$$

Here x_o is the value of the amplitude x at which the instantaneous bit error probability is equal to p_o.

C. Message Success Probability

The message success probability p_s is defined as the probability that, in a received message of L bits, all possible errors can be corrected.

$$p_s = \sum_{j=0}^{n_e} p_e^j (1 - p_e)^{L-j} \binom{L}{j}. \tag{24}$$

In (24), p_e is the bit error probability and n_e is the number of errors that can be corrected by the error correcting code.

V. Numerical Results

A. Bit Error Probability

By calculating (22) using the Gaussian integration technique, the bit error probability (21) is evaluated using the Newton–Cotes technique [12]. Fig. 2 shows the bit error probability for light, average, and heavy shadowing, according to the measured values found in [8] and reproduced in Table I. The mean phase μ_ϕ is zero in all cases.

The plots in Fig. 2 are for the narrowband case, i.e., without the use of spread-spectrum modulation. The major difference between these plots and those in [8] is that, in [8], the effects of phase and envelope fading were calculated separately, while in this paper their combined effect is shown. Further, in [8] an upperbound for the envelope fading was calculated by an approximation of the complementary error function, while we present an exact analysis.

If we compare Fig. 2 with the Fig. 2 in [8], where only envelope fading is considered, it appears that for low signal-to-noise ratio Fig. 2 gives higher values for the bit

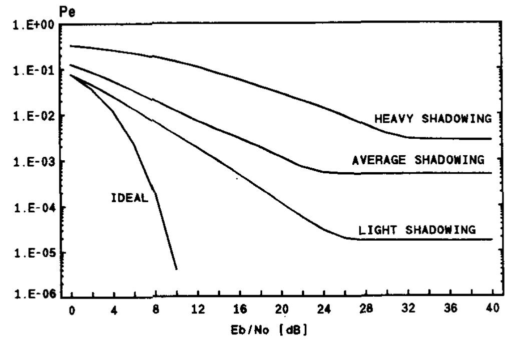

Fig. 2. Bit error probability for ideal BPSK and for narrowband BPSK with light, average, and heavy shadowing.

TABLE I
CHANNEL MODEL PARAMETERS

	Light	Average	Heavy
b_o	0.158	0.126	0.0631
μ	0.155	−0.115	−3.91
$\sqrt{d_o}$	0.115	0.161	0.806
σ	0.36	0.45	0.52

error probability. However, this is due to a mathematical error in the derivation of the upperbound of the bit error probability in [8]. Equation (25) is the correct upperbound of the bit error probability, which should replace (17) in [8].

$$p_e \leq \frac{1}{\sqrt{8\pi d_o}} \frac{\sigma^2}{b_o + \sigma^2} \int_0^\infty \frac{1}{z} \cdot \exp\left(\frac{-(\ln(z) - \mu)^2}{2d_o} - \frac{z^2}{2(b_o + \sigma^2)}\right) dz. \tag{25}$$

In addition, with the data in Table I, the power of the received signal in the case of light shadowing is greater than one, which results in a smaller bit error probability than theoretically possible at low signal-to-noise ratios. To repair this, we scaled the noise power by multiplying it with 1.585, which is the signal power for light shadowing. Without this scaling, all plots for light shadowing would shift 2 dB to the left.

In all three cases of shadowing, the bit error probability converges to the irreducible error probability p_{irr}, caused by the phase variation. A rather complicated expression for this error is given in (27) in [8]. By considering p_{irr} as the chance that the received bit in the absence of noise is negative while the transmitted bit is positive, p_{irr} can be simply given as:

$$p_{\text{irr}} = \int_{-\infty}^{0} p_u(x)\, dx, \qquad b_1^0 = 1. \tag{26}$$

The irreducible error probabilities in Fig. 2 do correspond with those found in [8].

Next, the effect of spread-spectrum modulation is com-

TABLE II
CALCULATED CHANNEL MODEL PARAMETERS

	σ_ϕ Narrowband	σ_ϕ Spread-Spectrum	b_o Spread-Spectrum
Light	0.40	0.14	0.023
Average	0.47	0.16	0.018
Heavy	1.55	1.42	0.009

puted. First, the bit error probability is shown for $T_c = 0.1$ μs, $1/T_b = 2400$ b/s, $N = T_b/T_c = 4095$ and $K = 1$. The fading parameters are adjusted according to (5) and (7). In Table II, the modified values for b_o and σ_ϕ are shown. Also, the calculated values for σ_ϕ for narrowband operation are given, which show a considerable difference with the measured values [6], [8] in the case of heavy shadowing. This difference may be due to the filtering of the received signal. It can be expected that σ_ϕ should be almost $\pi/2$ for heavy shadowing, because the distribution function approaches a Rayleigh pdf. This implies an almost uniformly distributed phase. However, in order to compare with [8], we have used the measured value for heavy shadowing ($\sigma_\phi = 0.52$) in all calculations. Assuming that, for heavy shadowing, σ_ϕ decreases by the same amount as for light and average shadowing, it is given by $(0.52/2.9)$ in the presence of spread-spectrum modulation. Comparing Figs. 2 and 3, it is seen that spread-spectrum modulation yields better performance than narrowband modulation for light and average shadowing. For heavy shadowing, the performance is worse at signal-to-noise ratios below 36 dB. The reason for this is that for light and average shadowing most of the signal power is received via the line-of-sight, so if the multipath power is reduced by the use of spread-spectrum modulation, a less perturbed signal will be obtained. In the case of heavy shadowing, however, the direct line-of-sight power is much smaller than the multipath power, resulting in an approximately Rayleigh faded signal. The use of spread-spectrum modulation now decreases the total signal power considerably, with the result that the resulting bit error probability increases. In this case, diversity techniques can be used to improve the bit error probability [14], [15], which is not considered here.

If the receiver is able to lock onto the total phase of the first path, then only the envelope fading has to be considered, which is investigated in Figs. 4–6, where the bit error probability is compared to the narrowband and spread-spectrum modulation. It can be seen from Figs. 2–6 that the bit error probability for narrowband modulation changes considerably by removing the phase variation, while in the case of spread-spectrum modulation there is much less change. The phase variance, which is reduced by approximately a factor 9 by the use of spread-spectrum modulation (7), exerts a nonnegligible influence only near the irreducible error probability level ($\leq 10^{-9}$ for spread-spectrum modulation).

It is seen from Fig. 7 that the light and average shadowed Rician distributions can be represented by the normal Rician distribution with a Rice factor of 5.3 and 4.1 dB, respectively, with an error of less than 5% for bit error probability values between 10^{-1} and 10^{-5}. Outside this interval, the error tends to increase. It can be expected that matching will be possible only in a certain range, since the two distributions functions are different in nature because of the shadowing. For heavy shadow-

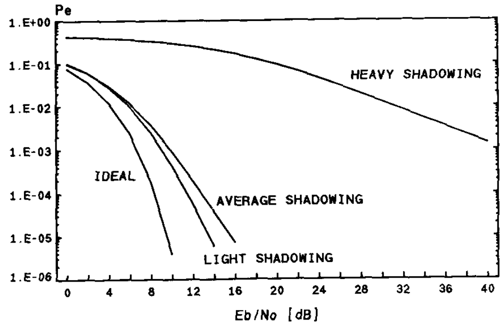

Fig. 3. Bit error probability for spread-spectrum modulation with $K = 1$ user, chip length $T_c = 0.1$ μs, Gold code length $N = 4095$, and bit rate $1/T_b = 2400$ b/s.

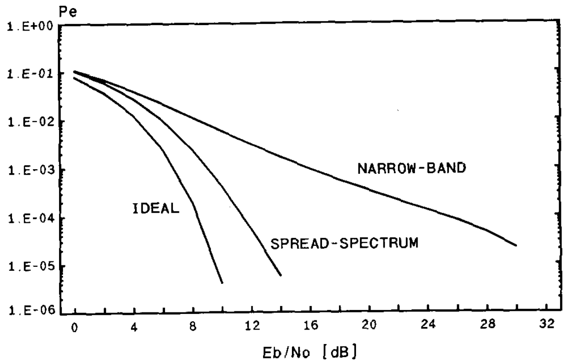

Fig. 4. Comparison of the bit error probability with narrowband and spread-spectrum modulation with light shadowing and envelope fading only for $K = 1$ user, chip length $T_c = 0.1$ μs, Gold code length $N = 4095$, and bit rate $1/T_b = 2400$ b/s.

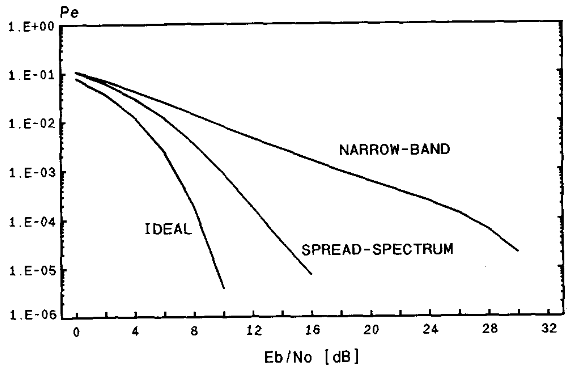

Fig. 5. Comparison of the bit error probability with narrowband and spread-spectrum modulation with average shadowing and envelope fading only for $K = 1$ user, chip length $T_c = 0.1$ μs, Gold code length $N = 4095$, and bit rate $1/T_b = 2400$ b/s.

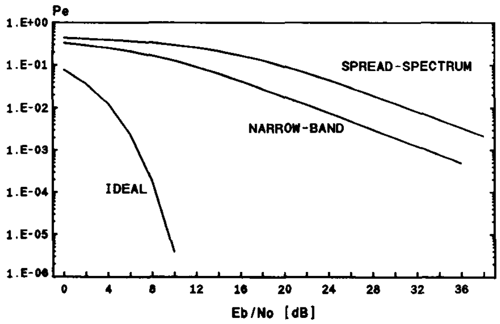

Fig. 6. Comparison of the bit error probability with narrowband and spread-spectrum modulation with heavy shadowing and envelope fading only for $K = 1$ user, chip length $T_c = 0.1\ \mu s$, Gold code length $N = 4095$, and bit rate $1/T_b = 2400$ b/s.

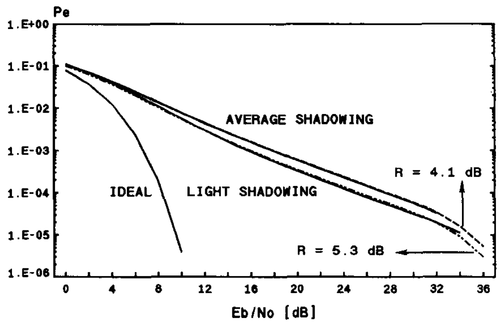

Fig. 7. Comparison of the shadowed Rician distribution with the normal Rician distribution for light and average shadowing with narrowband transmission.

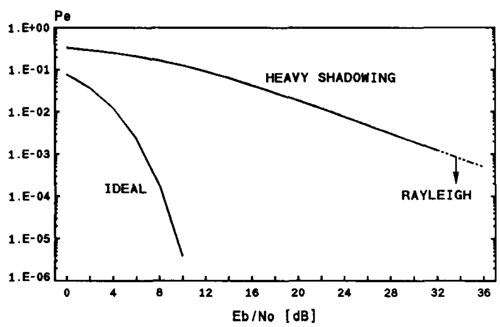

Fig. 8. Comparison of the shadowed Rician distribution with the Rayleigh distribution for heavy shadowing with narrowband transmission.

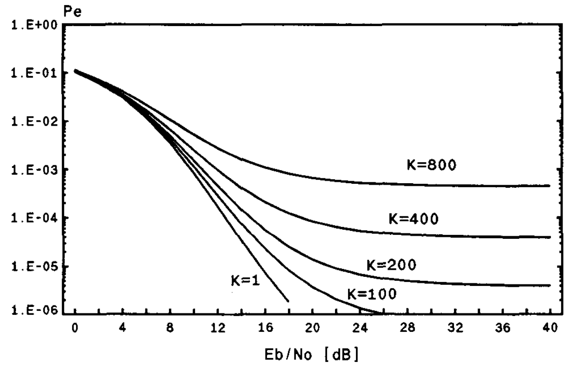

Fig. 9. Bit error probability for spread-spectrum modulation with average shadowing for chip length $T_c = 0.1\ \mu s$, Gold code length $N = 4095$, bit rate $1/T_b = 2400$ b/s, and K as a parameter.

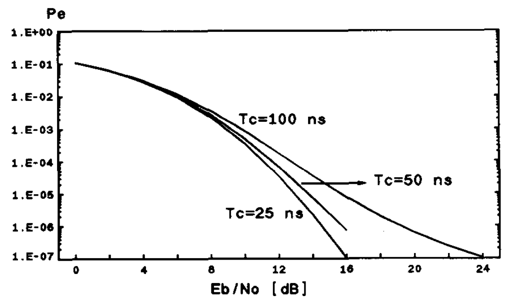

Fig. 10. Bit error probability for spread-spectrum modulation with average shadowing for $K = 1$, $N = 4095$, $T_b = T_cN$, and T_c as a parameter.

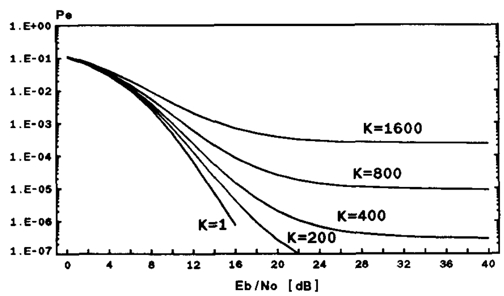

Fig. 11. Bit error probability for spread-spectrum modulation with average shadowing for $T_b = 1/2400$ s, $N = 8191$, $T_c = T_b/N$, and K as a parameter.

ing, however, the shadowing results in a negligible line-of-sight signal power. Accordingly, we see a perfect match with the curve of Rayleigh fading in Fig. 8, where the Rice factor is equal to zero.

Fig. 9 shows the results for spread-spectrum modulation with average shadowing and the number of users as a parameter. To maintain a bit error probability of 10^{-3}, the signal-to-noise ratio has to be increased by about 0.5 dB for $K = 100$ users, 1 dB for $K = 200$, and 2 dB for $K = 400$ as compared to the signal-to-noise ratio for a single user ($K = 1$). Note that E_b/N_o is the average signal-to-noise ratio, which does not include interference power. So increasing E_b/N_o decreases the bit error probability, until the irreducible bit error probability caused by the interference power is reached.

Fig. 10 shows that as T_c decreases, the bit error probability decreases because of the decreased multipath power.

Fig. 11 shows the bit error probability for $N = 8191$, $T_b = 1/2400$ s, $T_c = T_b/N$ and K as a parameter. The plot for $K = 1$ in Fig. 11 is the same as the one for $T_b = 1/4800$ s in Fig. 10, because they have the same values

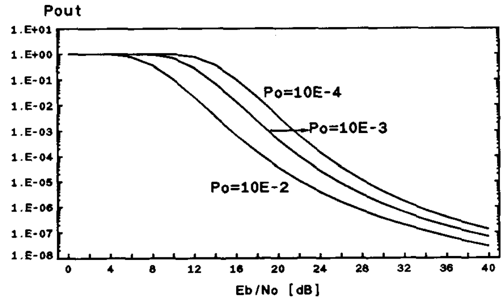

Fig. 12. Outage probability for spread-spectrum modulation with average shadowing for $K = 1$, $T_b = 1/2400$ s, and $T_c = T_b/4095$.

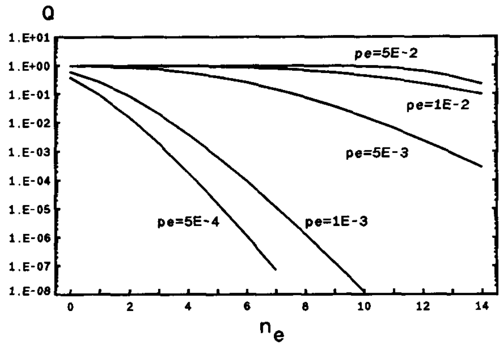

Fig. 13. Message error probability as a function of n_e for $K = 1$, $T_c = 0.1$ μs, $N = 4095$, $T_b = 1/2400$ s, and $L = 1024$.

for T_c. Comparing Figs. 9 and 11, it is seen that for particular values of p_e and E_b/N_o the number of users increases more than twice by doubling N, due to the reduced cross-correlation power and the reduced multipath power.

B. Outage Probability

The outage probability (23) is shown in Fig. 12 for $T_c = 0.1$ μs, $T_b = 1/2400$ s, $K = 1$ and average shadowing. To achieve an outage probability of 10^{-2}, for instance, the signal-to-noise ratio has to be 13, 16.5, or 19 dB for a threshold p_o of 10^{-2}, 10^{-3}, or 10^{-4}, respectively.

C. Message Success Probability

Instead of the message success probability, we have presented the computational results for the message error probability Q which is defined as $Q \triangleq 1 - p_s$.

Fig. 13 depicts the message error probability as a function of n_e with parameter p_e for $L = 1024$, $T_c = 0.1$ μs, $T_b = 1/2400$ s, $N = 4095$, $K = 1$ and for average shadowing.

In Fig. 14, Q is plotted as a function of the number of users K at $E_b/N_o = 10$ dB for $K = 1$. The other parameters are the same as in Fig. 13. If a maximum of 512 users, for instance, must be supported with a message success probability of at least 0.99, then the error correcting code must be able to correct 8 errors per 1024 bits.

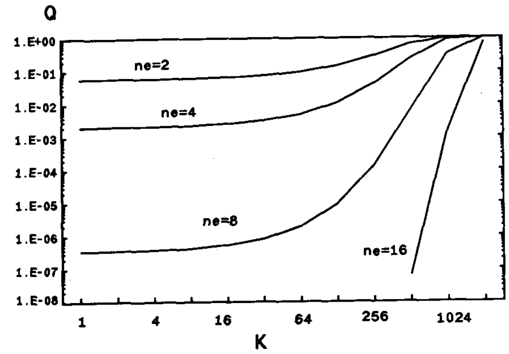

Fig. 14. Message error probability as a function of K for $E_b/N_o = 10$ dB, $T = 0.1$ μs, $N = 4095$, $T_b = 1/2400$ s, and $L = 1024$.

VI. Conclusions

The shadowed Rician channel model given in [4]–[6] has been expanded to a wideband model. The performance of a land–mobile satellite channel has been evaluated in terms of the bit error, outage and message success probability for light, average, and heavy shadowing, using a direct-sequence spread-spectrum transmission system with BPSK modulation. The main conclusion that can be drawn from this analysis is that the spread-spectrum system yields better performance than the narrowband transmission if the line-of-sight path is dominant.

It is shown that for bit error probability calculations, the shadowed Rician distribution can be approximated in a limited but useful range of the bit error probability by a normal Rician distribution with a Rice factor of 5.3, 4.1, and $-\infty$ dB for light, average, and heavy shadowing, respectively. When spread-spectrum modulation is used, these Rice factors will increase because of the multipath rejection capability of the correlation operation in the receiver. As a result, spread-spectrum modulation yields better performance than narrowband transmission, except for heavy shadowing, i.e., where the line-of-sight signal power is smaller than the multipath power. In the latter case, diversity techniques can be added to improve the performance. Further, it is recommended to study the influence of very low rate convolutional codes [17] on the performance of spread-spectrum multiple access land-mobile satellite channels.

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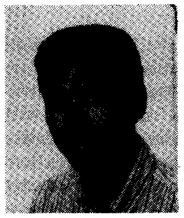
Richard D. J. van Nee was born in Schoonoord, the Netherlands, on January 17, 1967. He received the M.Sc. degree in electrical engineering (with distinction) from the University of Twente, Twente, The Netherlands, in 1990.

He is currently working towards the Ph.D. degree at Delft University, The Netherlands, where he studies the influence of channel fading on satellite communication and navigation.

Mr. van Nee received the Ashtech Award for the best student paper at the Institute of Navigation GPS-91 conference, held in Albuquerque, New Mexico, in September 1991, for his paper on the effects of multipath propagation on spread-spectrum code tracking errors.



Howard S. Misser (S'89–M'91) was born in Paramaribo, Surinam, on May 11, 1968. He received the M.Sc. degree in electrical engineering from Delft University, Delft, The Netherlands, in 1990.

He worked at the Telecommunications and Traffic-Control Systems Group of the same university as a Research Fellow. He is currently with Dr. Neher Laboratories of the Royal Dutch PTT in Leidschendam, The Netherlands. His research interests are in the fields of broadband network technologies and architectures, radio and mobile communication, and spread-spectrum communication.

Ramjee Prasad (M'89–SM'90) was born in Babhnaur (Gaya), Bihar, India, on July 1, 1946. He received the B.Sc. (Eng.) degree from the Bihar Institute of Technology, Sindri, India, and the M.Sc. (Eng.) and Ph.D. degrees from the Birla Institute of Technology (BIT), Ranchi, India, in 1968, 1970, and 1979, respectively.

He joined BIT as a Senior Research Fellow in 1970 and became Associate Professor in 1980. From 1983 to 1988, he was with the University of Dar es Salaam (UDSM), Tanzania, where in 1986 he became Professor in Telecommunications at the Department of Electrical Engineering. Since February 1988, he has been with the Telecommunications and Traffic Control Systems Group, Delft University of Technology, The Netherlands, where he is actively involved in the area of mobile and indoor radio communications. While he was with BIT, he supervised many research projects in the area of microwave and plasma engineering. At UDSM, he was responsible for the collaborative project "Satellite Communications for Rural Zones" with the Eindhoven University of Technology. He has published over 80 technical papers. His current research interests are in packet communications, adaptive equalizers, spread-spectrum systems, and telematics. He has served as a member of advisory and program committees of several IEEE international conferences. He has also presented tutorials on mobile and indoor radio communications at various universities, technical institutions, and IEEE conferences. He is also a member of a Working Group of the European Co-Operation in the field of scientific and technical research for Project (COST-231) as an expert for the Netherlands.

Prof. Prasad is a Fellow of the Institution of Electronics and Telecommunication Engineers.