Performance of Microcellular Mobile Radio in a Cochannel Interference, Natural, and Man-Made Noise Environment

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***Abstract*—A model is developed for performance analysis of a microcellular digital mobile radio system with Rayleigh-faded cochannel interference, Gaussian noise, and narrow-band impulsive noise (Middleton's Class A noise) using the differential phase shift keying modulation (DPSK) technique. The desired signal has been assumed to be Rician faded. The effects of selection diversity on the performance have also been investigated and compared to the nondiversity case. The performance is measured in terms of bit error probability and spectrum efficiency. The influence of reuse distance, i.e., cluster size, traffic intensity, i.e., blocking probability, impulsive index, Rician parameter, and turning point of the dual path loss law characteristic of microcells on the performance parameters have been investigated in detail.**

I. Introduction

As a result of the scarcity of available frequency bandwidth for radio applications in general, radio system designers are forced to develop communication systems that are very efficient with respect to the bandwidth. This constraint holds even more for the designers of large capacity mobile communications systems.

The application of digital cellular radio communications invokes a new range of techniques to increase this efficiency. So the key-words for such high capacity systems are digital and cellular. Evaluation of their performance have been reported in numerous papers, e.g., [1]–[5]. Most studies consider the interfering noise as Gaussian. But a realistic electromagnetic (EM) environment contains also man-made noise, that is partly impulsive. It would be interesting to analyze the effect of such noise on the system performance, in order to provide a more reliable tool for the systems designer.

The cellular system studied in [3]–[5] has relatively small cells, known as microcells; see also [6]. A microcell is a subject of major interest in science and industry at the moment. This is because the overall system capacity can be increased by decreasing the cell size, so channels can be reused more often.

The radio signal propagation in a microcell differs from that in macrocells (conventional cells) in two ways: i) a line-of-sight path may exist between the two antennas and ii) the path loss law follows dual path loss characteristics. Microcellular systems interfered by Gaussian noise only, are studied in [4]. In the present paper, the effects of cochannel interference, Gaussian, and man-made noise on a microcellular network are investigated. The modulation scheme considered is differential phase shift keying (DPSK). The performance criteria studied are the bit error probability and the spectrum efficiency. Selection diversity, a well-known technique to reduce the effects of multipath fading, is also considered.

In the future, a high capacity mobile radio system will consist of different cell types. Firstly, the traditional macrocells with a 1–20 km diameter and a power range from 0.6–10 W. Secondly microcells, covering high traffic density areas with 0.4–2 km diameter and powers that can be less than 20 mW. At the indoor end of the system, coverage will be provided by picocells with a cell diameter of 15–200 m.

Compared to conventional cells, called macrocells, microcells offer good signal quality in a low time dispersion and a low RF power arena, due to the potential line-of-sight transmission and the shorter radio paths. This is confirmed by several experimental measurements [7].

This paper is organized as follows, Section II briefly reviews the man-made noise. The propagation characteristics of a microcell are described in Section III. The performance parameters are derived and evaluated in Section IV. Section V presents the effects of selection diversity on the performance parameters. Finally, conclusions are drawn in Section VI.

II. Man-Made Noise

During the past two decades, man-made electromagnetic interference (or noise) has become a problem of increasing concern to the telecommunication community.

The deteriorating effects of man-made noise on system performance is now generally recognized. The sources of man-made noise are numerous: incidental radiation from electrical devices of all sorts, out-of-band modulation products from radio communication systems, automotive ignition systems, electric power lines; and so on.

This man-made contribution to the EM-environment is basically 'impulsive,' i.e., it has a highly structured form, characterized by significant probabilities of interference levels [8].

In a metropolitan area, man-made interference can be present in the 30 Hz to 7 GHz radio spectrum range. Above 100 MHz the spectrum is dominated by man-made noise with automotive ignition noise as a major contributor, but up to 100 MHz atmospheric noise and several man-made sources form a mixture. At lower frequencies the impulse widths and amplitudes of the impulsive emission patterns are greater, but the average

Manuscript received June 24, 1992.

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IEEE Log Number 9205016.

occurrence rate of the largest pulses may increase with the frequency [9].

These highly non-Gaussian random processes can have severe degrading effects on system performance, particularly on most conventional systems, which are designed for optimal or near optimal performance against normal noise.

Therefore, it is important that the true EM-environment (natural plus man-made noise) is modeled by a physical statistical model, providing a better basis for system design and comparison. A lot of research has been done on the development of statistical-physical models. The model proposed by Middleton [10], including both man-made noise and Gaussian noise, and classifying the EM-interference in three classes, is used in this paper and therefore briefly reviewed.

Three classes of EM-interference (including both man-made and Gaussian noise) are identified in Middleton's theory [10]:

- Class A: The spectral bandwidth of the noise entering the receiver is comparable to or less than the bandwidth of the receiving system. Transient effects are ignorable.
- Class B: The bandwith of the noise is greater than the bandwidth of the receiving system, i.e., the noise pulses produce transients in the receiver.
- Class C: A linear sum of Class A and Class B.

Middleton's Class A interference model is governed basically by three parameters (A, Γ, Ω), where [11]:

A = the impulsive index, also known as 'overlap' or 'unstructure' index, which has been defined as the average number of radiation 'events' per second times the mean duration of a typical source emission. The smaller A, the more the interference is structured (in time). When $A \to \infty$ the noise is Gaussian.

$\Gamma \equiv \sigma_G^2/\sigma_p^2$ = the Gaussian factor = the ratio of average intensity of the Gaussian component of the interference to that of the non-Gaussian component.

σ_p^2 = the mean intensity of the non-Gaussian (or 'impulsive') noise component of the interference.

From [12] we find that the probability density function for the instantaneous amplitude yields

$$P_{\text{ampl}}(X) \simeq e^{-A} \sum_{m=0}^{\infty} \frac{A^m}{m!\sqrt{2\pi\sigma_m^2}} \cdot e^{\frac{-X^2}{2\sigma_m^2}} \tag{1}$$

where

$$\sigma_m^2 \equiv \frac{\frac{m}{A} + \Gamma}{1+\Gamma}. \tag{2}$$

Note that $P_{\text{ampl}}(X)$ is in fact a weighted sum of Gaussian distributions with increasing variance. Also the distribution function of the envelope can be obtained from [12]:

$$P_1(E \geq E_0) \simeq e^{-A} \sum_{m=0}^{\infty} \frac{A^m}{m!} \cdot e^{\frac{-E_0^2}{\sigma_m^2}}. \tag{3}$$

This density function can be seen as a weighted sum of Rayleigh distributions with increasing variance.

III. Microcellular Mobile Radio Systems

Measured propagation results indicate that the received signal envelope is Rician distributed [7], [13]. Then the probability density function (PDF) for the ratio of instantaneous signal power to the sum of cochannel interference, Gaussian and impulsive noise can be written as

$$f(\rho) = \frac{1}{\sigma^2} \exp\left(-\frac{2\rho + S^2}{2\sigma^2}\right) I_0\left[\frac{\sqrt{2\rho S}}{\sigma^2}\right]$$

where

$$\begin{aligned} \sigma^2 &= \sigma_a^2/(P_u + \sigma_g^2 + \sigma_p^2), \\ \rho &= \rho_a/(P_u + \sigma_g^2 + \sigma_p^2), \\ S^2 &= S_a^2/(P_u + \sigma_g^2 + \sigma_p^2). \end{aligned} \tag{4}$$

Where S_a denotes the line-of sight signal, σ_a^2 is the average scattered power, P_u is the total mean cochannel interference power, σ_g^2 is the mean power for Gaussian noise, σ_p^2 is the mean power for impulsive noise and $I_0[]$ is the modified Bessel function of the first kind and zero order.

The absence of a line-of-sight component in the cochannel interference is a realistic assumption because of the relatively large distances, so we assume that the cochannel interferer is Rayleigh faded and its power is exponentially distributed. For the ith interferer, the pdf of the interference power, p_i, can be expressed as

$$f(p_i) = \frac{1}{P_{oi}} \exp\left(-\frac{p_i}{P_{oi}}\right), \qquad 0 \leq p_i < \infty \tag{5}$$

where P_{oi} is the local mean power of the ith interferer.

Another aspect of microcell propagation is the path loss characteristic. In microcells the logarithmic attenuation slope is modeled by two straight lines, one representing the (inverse square) slope between base site and so called turning point, and the other (inverse fourth-power) between turning point and infinity [14], [15]. The turning point has to be experimentally obtained; 100–200 m is a typical value for an urban environment. This path loss characteristic yields [14]–[16]

$$P_o = P_t C \frac{g_b g_m}{d^a\left(1 + \frac{d}{g}\right)^b} \tag{6}$$

where C is a constant, P_o is the average received power, P_t is the average emitter power, g_b and g_m are the antenna gains, d is the distance between base site and receiver, a is basic attenuation rate for short distances (approximately 2), b is the additional attenuation rate for distances larger than 100 to 200 m and is approximately 2 and g is the turning point. It is expected that b and g depend on the environment (kind of buildings, amount of street traffic, etc.).

IV. DPSK Modulation Performance

In this section a DPSK system is described. The advantage of DPSK over many other modulation types is that in DPSK no absolute phase reference is needed, only nearby constant carrier phase during two bit successive intervals is required. That property of DPSK makes it appropriate for a fading environment.

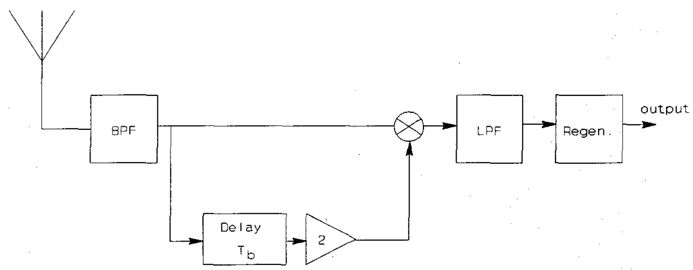

Fig. 1. The suboptimum DPSK receiver.

Given a differentially encoded PSK signal, there are several demodulation methods [17]. In practice this demodulation is often done by delaying the received signal by the bit interval T and using the delayed signal to multiply the received signal. This is visualized in Fig. 1. Then the decision variable is the projection of the received signal on the phase shifted delayed signal. When using binary DPSK with 0 and π as possible phase shifts and the carrier frequency is an integer multiple of the signaling rate, this decision variable is [17]

$$D_v = 2\mathrm{Re}\left[V_n V_{n-1}^*\right] = V_n V_{n-1}^* + V_n^* V_{n-1} \tag{7}$$

where

$$\begin{aligned} V_n &= A_{0,n} e^{j\phi} + N_n \\ N_n &= [n_i \cos(\theta_n) + i n_q \sin(\theta_n)] e^{j\phi}. \end{aligned} \tag{8}$$

Here V_n is the received signal in the nth interval, N_n the noise component, Θ_n noise phase and ϕ is the carrier phase, and the carrier phase is presumed to be constant over at least $2T$. The inphase and quadrature noise components are denoted as n_i and n_q. Without loss of generality, we assume that $A_{0,n} = A_{0,n-1}$ which is allowed under the condition that $A_{0,n}$ is likely to be $+1$. Then D_v yields [16], [18]

$$D_v = A_0^2 + A_0 n_{i,n} + A_0 n_{i,n-1} + n_{i,n} n_{i,n-1} + n_{q,n} n_{q,n-1}. \tag{9}$$

Because the modulation phases are 0 and π, the decision level is zero. This is an advantage in Rayleigh fading environments, since the level itself is not sensitive to that fading. Now the bit error probability can be written as

$$P_{\text{biterror}} = \Pr\{D_v < 0 | A_0^2 = 1\}. \tag{10}$$

For Gaussian noise the pdf of D_v contains a quadratic Gaussian random variable and is derived in [17]. The final expression yields

$$P_{\text{biterror, Gaussian noise}} = \frac{1}{2}\exp(-\rho_b) \tag{11}$$

where ρ_b is the SNR, and the noise is assumed to be independent from bit interval to interval.

However, we are interested in the bit error probability in Middleton's Class A noise. Deriving that probability from the decision variable as defined in (9) is extremely complicated because of the noise product terms. Therefore, an approximation is presented here. This estimated performance measure is proposed by Marras [19]. It is based on the conditional error probability for the noncoherent correlation (i.e., suboptimum) FSK receiver, given by Middleton [20, (51b)]. Using the relation of DPSK to FSK in white Gaussian noise, given by Viterbi [21], Marras derives an approximation that yields [19]

$$P_e(\rho) = \frac{1}{2}\exp(-A)\sum_{m=0}^{\infty}\frac{A^m}{m!}\exp\left(-\frac{\rho}{\sigma_m^2}\right) \tag{12}$$

where ρ is the normalized power (so the SNR) and A and σ_m are the parameters defined in the previous section. Equation (12) is the main tool for the performance analysis.

Another issue in our model description is the inclusion of the Raleigh faded interferers. We assume an additive model as shown in Fig. 2.

All cochannel signals are considered to be independent and uncorrelated. The phases are assumed to be uniform randomly distributed from bit to bit, but constant during one bit period. Since we are considering bandpass signals here, we can write the inphase and quadrature components of the sum $z(t)$, u_i and u_q, respectively

$$\begin{aligned} u_i &= \sum_{j=1}^{n} \alpha_j a_j \cos\phi_j \\ u_q &= \sum_{j=1}^{n} \beta_j a_j \sin\phi_j, \qquad a_j = \pm 1. \end{aligned} \tag{13}$$

In (13), α_j and β_j are Rayleigh distributed. Multiplying these variables with the randomly phased cosine or sine terms gives a Gaussian variable as a product. In this event the joint interference power may be added to the Gaussian noise. This assumption has been applied in several papers [22]–[24]. With mean power p_{oj} of the jth interferer and mean power σ_g^2 for the Gaussian noise, we may write for the new long term average Gaussian power $\sigma_g'^2$

$$\sigma_g'^2 = \sigma_g^2 + \sum_{j=1}^{n} P_{oj} = \sigma_g^2 + P_u. \tag{14}$$

Now the cochannel interference can be included in the Gaussian noise portion, it has been implemented as an addition to the Middleton noise parameter Γ just by substituting σ_g^2 by $\sigma_g'^2$.

The described model is used to analyze the microcellular network. The reciprocal of the mean received signal power to noise plus interference power at a receiving mobile station

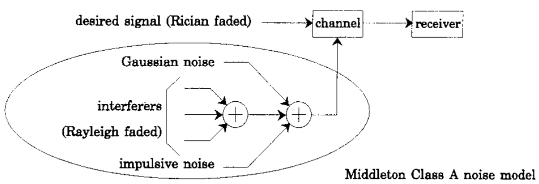

Fig. 2. Diagram of the additive channel model.

located at the microcell border, γ, can be written as

$$\gamma \triangleq \frac{\sigma_g^2 + \sigma_p^2 + P_u}{(1+k)\sigma^2} = \frac{n}{(R_u - 1)^2}\left(\frac{G+1}{G + (R_u - 1)}\right)^2 + \frac{1}{S_n} \tag{15}$$

where $R_u(\triangleq D/R)$ is the (normalized) reuse distance, defined as the ratio of the distance between the centers of the nearest neighboring cochannel cells (D) to the cell radius R, as visualized in Fig. 3, n is the number of cochannel interferers being active, k is the Rician factor $(= S^2/2\sigma^2)$, S_n is the total received signal to mean Gaussian noise power ratio and $G(= g/R)$ is defined as the normalized 'turning point' being the ratio of the path loss turning point g and the cell radius R.

The conditional bit error probability $P(e|n)$ is obtained by integrating the product of $f(\rho)$ and $P_{e(\rho)}$ from zero to infinity.

$$P(e|n) = \frac{1}{2\sigma^2}\exp\left(-A - \frac{S^2}{2\sigma^2}\right)\sum_{m=0}^{\infty}\frac{A^m}{m!} \cdot \int_0^\infty \exp\left\{-\rho\left(\frac{1}{\sigma^2} + \frac{1}{\sigma_m^2}\right)\right\} I_0\left[\frac{\sqrt{2\rho S}}{\sigma^2}\right] d\rho. \tag{16}$$

Equation (16) simplifies to

$$P_n(e|n) = \frac{1}{2}\exp(-A - k)\sum_{m=0}^{\infty}\frac{A^m}{m!}\alpha\exp(k\alpha)$$

where

$$\alpha \triangleq \frac{\sigma_m^2}{\sigma_m^2 + \frac{1}{(1+k)\gamma}}. \tag{17}$$

Finally the bit error probability is evaluated by using

$$P_e \triangleq \sum_n P_n(e|n)F_n(n) \tag{18}$$

where $F_n(n)$ is the probability of n cochannel cells being active. If we adopt the commonly assumed hexagonal cell structure, $F_n(n)$ can be expressed in terms of carried traffic a_c per channel

$$F_n(n) = \binom{6}{n} a_c^n (1 - a_c)^{6-n}. \tag{19}$$

Additionally to the bit error probability, another performance measure, the spectrum efficiency E_s is defined as

$$E_s \triangleq \frac{a_c}{WCS_c} \quad [\text{erlang/MHz/km}^2] \tag{20}$$

where a_c is again the carried traffic per channel in erlang, W the channel bandwidth in megahertz, C the cluster size and S_c the cell area in square kilometers. Spectrum efficiency can be evaluated using (20) and the relation between reuse distance and the number of cells per cluster, i.e.,

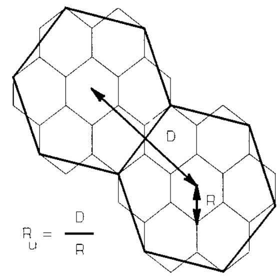

Fig. 3. Configuration of a cellular system and definition of reuse distance.

$$R_u = (3C)^{\frac{1}{2}}. \tag{21}$$

In the above described model, setting the Rice factor k and normalized turning point G to zero corresponds to a macrocellular system, where the desired signal is assumed to be Rayleigh faded and the path loss function is of second order. In this case (17) reduces to

$$P_{k=0}(e|n) = \frac{1}{2}\exp(-A)\sum_{m=0}^{\infty}\frac{A^m}{m!}\frac{\sigma_m^2}{\sigma^2 + \sigma_m^2}. \tag{22}$$

Another special condition is the hypothetical case where all base station transmitter powers increase to infinity. Then the bit error probability is completely determined by the interfering cochannel cells and (17) reduces to

$$P(e|n) = \frac{1}{2}\frac{1+k}{\gamma + 1 + k}\exp\left(\frac{-k\gamma}{\gamma + 1 + k}\right). \tag{23}$$

Another boundary is the situation where the cochannel interference is absent and the bit error probability is completely determined by the Gaussian and impulsive noise. If we again consider (17), it is seen that in this case

$$P(e|n) = \frac{1}{2}\exp(-A - k)\sum_{m=0}^{\infty}\frac{A^m}{m!} \cdot \frac{\sigma_m^2}{S_n + \sigma_m^2}\exp\left(\frac{\sigma_m^2}{S_n + \sigma_m^2}k\right). \tag{24}$$

Notice that if also the transmitter power decreases to zero, the conditional bit error probability increases to 1/2, giving an overall error P_e of 1/2 at the receiver.

Bit error probability is numerically evaluated using (12)–(19) and the results can be found in Fig. 4 to Fig. 6. Fig. 4 shows the bit error probability P_e against reuse distance R_u for a carrier-to-noise ratio of 10 dB, carried traffic per channel $a_c = 0.5$ erlang, Rice factor $k = 7$ dB, normalized turning point $G = 0.2$ and $\Gamma = 0.001$ for several values of A. Parameter A denotes the character of the noise; the individual Gaussian and impulsive noise powers are invariant for A. It can be seen from Fig. 4 that decreasing A improves the performance. Therefore it can be concluded that the more

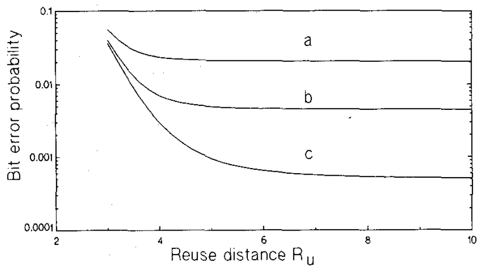

Fig. 4. Bit error probability P_e against reuse distance R_u for a carrier-to-noise ratio of 10 dB, $a_c = 0.5$ erlang, Rice factor $k = 7$ dB, $G = 0.2$, $\Gamma = 0.001$ for several parameter A values (a) $A = 0.1$, (b) $A = 0.01$, (c) $A = 0.001$.

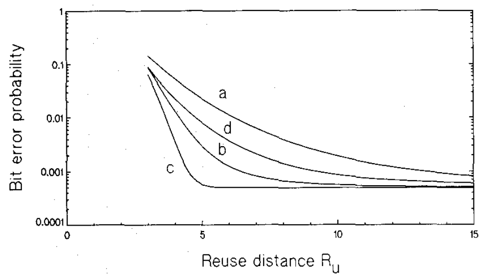

Fig. 5. Bit error probability P_e against reuse distance R_u for a carrier-to-noise ratio of 10 dB, $a_c = 0.5$ erlang, $A = 0.001$, $\Gamma = 0.001$ with Rice factor k and normalized turning point G as parameters (a) $k = 0, G = 2$, (b) $k = 7$ dB, $G = 2$, (c) $k = 12$ dB, $G = 2$, (d) $k = 0$, $G = 0.2$.

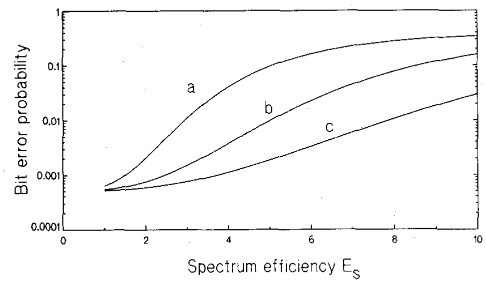

Fig. 6. Bit error probability P_e against spectrum efficiency E_s for a carrier-to-noise ratio of 10 dB, $A = 0.001$, $G = 0.2$, $\Gamma = 0.001$, $W = 0.025$ MHz, $k = 7$ dB with carried channel traffic a_c as a parameter (a) $a_c = 0.23$ erlang, (b) $a_c = 0.5$ erlang, (c) $a_c = 0.99$ erlang.

structured (in time) the noise is, the less impact it has on the bit error probability.

Fig. 5 depicts the bit error probability P_e against reuse distance R_u for a carrier-to-noise ratio of 10 dB, $a_c = 0.5$ erlang, $A = 0.001$, $\Gamma = 0.001$ with Rice factor k and normalized turning point G as parameters. Here bit error decreases with increasing k and decreasing G values. Since for $k = 0$, (4) denotes the PDF of a Rayleigh faded signal, this case can represent the macrocellular case, if additionally also G is taken to be zero.

If Fig. 6 the bit error probability P_e against spectrum efficiency E_s for a carrier-to-noise ratio of 10 dB, $A = 0.001$, $G = 0.2$, $\Gamma = 0.001$, bandwidth $W = 0.025$ MHz, Rice factor $k = 7$ dB is given, with carried channel traffic a_c as a

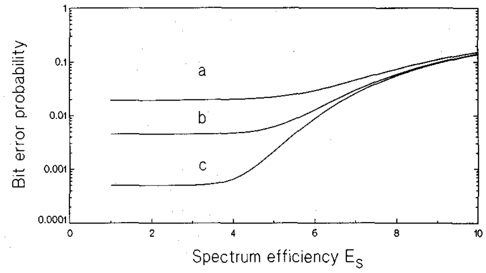

Fig. 7. Bit error probability P_e against spectrum efficiency E_s Here $G = 0.2$, $k = 12$ dB, carrier-to-noise ratio is 10 dB, $a_c = 0.5$ erlang, $\Gamma = 0.001$ $W = 0.025$ MHz. Parameter A values are (a) $A = 0.1$, (b) $A = 0.01$, (c) $A = 0.001$.

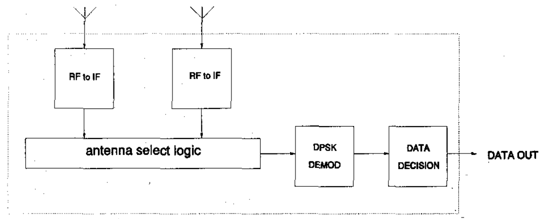

Fig. 8. A possible selection diversity receiver diagram.

parameter. Here the spectrum efficiency increases with carried traffic a_c for a given bit error probability. Bit error probability P_e against spectrum efficiency E_s for several A values is also shown in Fig. 7 but for a Rician factor k of 12 dB and $G = 0.2$, a carrier-to-noise ratio of 10 dB, $a_c = 0.5$ erlang, $\Gamma = 0.001$ and channel bandwidth $W = 0.025$ MHz.

V. Selection Diversity Performance

Selection diversity is used to improve the system performance in the case of fast multipath fading. It is based on selecting the largest one from a group of signals carrying the same information. Largest means in this perspective the signal with the highest signal to noise ratio. In practice, the signal with the largest signal plus noise is usually used, since it is difficult to measure SNR. Fig. 8 shows a possible receiver scheme for two branch selection diversity.

Assume there are L branches, giving L identically distributed signal variables $\{\beta_1, \cdots, \beta_L\}$. Each variable is distributed according to (4). For the largest out of the L signals, $\beta_{\max}$, holds

$$\beta_{\max} > \beta_i, \qquad i \in \{i \cdots L\}. \tag{25}$$

Now the CDF of $\beta_{\max}$, $F_{\beta\max}$ is

$$F_{\beta_{\max}}(\rho) = \prod_{i=1}^{L} F_{\beta_i}(\rho) = \left[1 - Q\left(\frac{S}{\sigma}, \frac{\sqrt{2\rho}}{\sigma}\right)\right]^L \tag{26}$$

where $Q(.,.)$ is Marcum's Q-function. From (26) the PDF can be obtained by differentiating. After transforming this result to the binomial form, this PDF is shown to be

$$f_{\beta_{\max}}(\rho) = L \sum_{i=0}^{L-1} \binom{L-1}{i} (-1)^i Q^i\left(\frac{S}{\sigma}, \frac{\sqrt{2\rho}}{\sigma}\right) \cdot f(\rho). \tag{27}$$

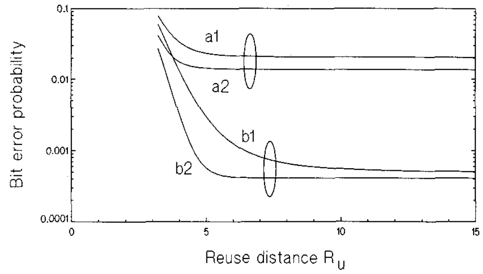

Fig. 9. Bit error probability P_e against reuse distance R_u using selection diversity. Carrier-to-noise ratio of 10 dB, $a_c = 0.5$ erlang, Rice factor $k = 7$ dB, $G = 2$, $\Gamma = 0.001$ for several parameter A values (a1) $A = 0.1$ $L = 1$, (a2) $A = 0.1$ $L = 2$, (b1) $A = 0.001$ $L = 1$, (b2) $A = 0.001$ $L = 2$.

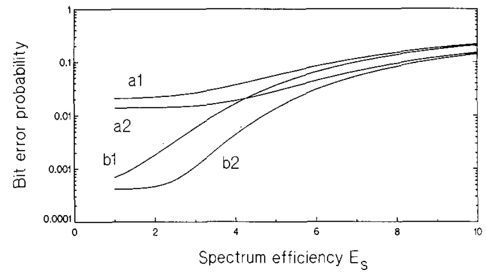

Fig. 10. Bit error probability P_e against spectrum efficiency E_s for a carrier-to-noise ratio of 10 dB, $A = 0.001$, $G = 2$, $\Gamma = 0.001$, $W = 0.025$ MHz, $k = 7$ dB with A as parameter. (a1) $A = 0.1$ $L = 1$, (a2) $A = 0.1$ $L = 2$, (b1) $A = 0.001$ $L = 1$, (b2) $A = 0.001$ $L = 2$.

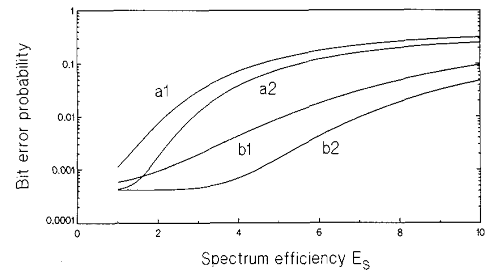

Fig. 11. Bit error probability P_e against spectrum efficiency E_s for a carrier-to-noise ratio of 10 dB, $A = 0.001$, $G = 2$, $\Gamma = 0.001$, $W = 0.025$ MHz, $k = 7$ dB with carried channel traffic a_c as a parameter $a_c = 0.23$: (a1) $L = 1$ (a2) $L = 2$; $a_c = 0.99$: (b1) $L = 1$ (b2) $L = 2$.

With this PDF the performance can be evaluated analogue to the previous section. The resulting conditional bit error probability yields

$$P(e|n,L) = \frac{L\cdot\exp(-A-k)}{2\sigma^2}\sum_{i=0}^{L-1}\sum_{m=0}^{\infty}\binom{L-1}{i}(-1)^i\frac{A^m}{m!} \cdot\int_0^{\infty} Q^i\left(\frac{S}{\sigma},\frac{\sqrt{2\rho}}{\sigma}\right)\cdot I_0\left[\frac{\sqrt{2\rho}S}{\sigma^2}\right] \cdot\exp\left(-\rho\left(\frac{1}{\sigma_m^2}+\frac{1}{\sigma^2}\right)\right)d\rho. \tag{28}$$

For $L = 2$, we write (28) with the aid from [25, pp. 451]

$$P(e|n,2) = \exp(-A-k)\sum_{m=0}^{\infty}\frac{A^m}{m!}\alpha\exp(\alpha k) \cdot\left\{1-Q\left(\sqrt{\frac{2k}{1+\alpha}},\alpha\sqrt{\frac{2k}{1+\alpha}}\right) +\frac{\alpha}{1+\alpha}\cdot\exp\left(-k\frac{1+\alpha^2}{1+\alpha}\right)\cdot I_0\left[\frac{2k\alpha}{\alpha+1}\right]\right\}. \tag{29}$$

With α as defined in (17).

With the obtained results, we analyze the effectiveness of selection diversity using Figs. 9–11. Unless otherwise stated the carrier-to-noise ratio is 10 dB, $a_c = 0.5$ erlang, Rice factor $k = 7$ dB, $G = 2$, $\Gamma = 0.001$ and bandwidth W is 25 kHz. In Fig. 9 the bit error probability P_e against reuse distance R_u is plotted using selection diversity for several parameter A values. The plot shows an improvement in performance for both A values, but the progression is small for large reuse distance R_u.

Spectrum efficiency is shown in Fig. 10. Fig. 10 shows us again that a smaller impulsive index A, results in a better performance. But it shows also that despite the lower index, selection combining can be effective, provided that there is enough cochannel interference to give the total noise a less impulsive character.

Finally in Fig. 11 the curves for two carried traffic values are considered. Here parameter A is 0.001. From this plot, it can be seen that for higher carried traffic, the improvement on the performance when applying selection diversity is considerably larger.

VI. Conclusions

In this paper a model has been developed that can be used as an aid for system design of digital microcellular land mobile radio systems. The model includes Rician faded desired signal, taking into account a potential line-of-sight situation, and Rayleigh faded cochannel interference. Also included are Gaussian noise, narrowband impulsive noise, and a realistic path loss model, which enables the designer to investigate small base to mobile distance. The considered modulation is DPSK, where an approximation of the suboptimum receiver has been used.

For a cellular mobile system where the noise is partially narrowband impulsive, it is found that the performance is better compared to a system with purely Gaussian noise, given equal noise power in both cases. Another conclusion that can be drawn is that given an amount of impulsive noise, a smaller impulsive index A yields an improvement in performance. So in other words, large amplitude short duration pulses degrade the system less than small amplitude large duration pulses. Once the impulsive index is above a certain threshold, the impulsive noise power is irrelevant; only the frequency of impulsive events times their mean duration determines the error performance.

The next conclusion is that the selection diversity scheme is not so effective in an impulsive noise environment as with

Gaussian noise only. When the noise pulses are large and of short duration (i.e., small impulsive index A), the benefit of selection diversity holds only for that reuse distances where cochannel interference has considerable impact on the desired signal.

Finally, it can be stated that in this paper a tool for microcellular radio system planning has been presented. The system performance can be evaluated from two viewpoints: bit error probability and spectrum efficiency. From these perspectives the model allows to study the effects of the varying parameters (e.g., cell size or transmitter power) on the performance.

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