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An Initial Code Acquisition Scheme for Indoor Packet DS/SS Systems with Macro/Micro Antenna Diversity

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SUMMARY In this paper, we study macro/micro diversity techniques for code acquisition of a direct-sequence spread-spectrum signal in an indoor packet communication system. In the system discussed, the base station has several radio ports each with a cluster of antennas, and the terminal also has multiple antennas. The performance in the uplink of this system is analyzed under Lognormal shadowing and flat Rayleigh fading. The numerical results show great performance improvements by proposed diversity techniques. In addition, it is clarified that the mean acquisition time, which is often used as the measure of performance, is not suitable for packet radio systems as it underestimates the necessary preamble length for initial code acquisition.

key words: spread spectrum, code acquisition, indoor packet radio, spatial diversity, fading and shadowing

1. Introduction

Not only for outdoor mobile systems, wireless communications are advantageous also for indoor systems. Wireless local area network (Wireless-LAN) is an example of indoor radio systems. The substitution of short cables between devices of a personal computer and consumer electronics devices is recently attracting attention. Another interesting application of wireless data communication systems is in factory production lines, where remote-controlling machines in motion is becoming more and more important.

For these indoor wireless communication systems, especially, the case of remote controlling in factories, and the case of Wireless-LAN, direct sequence spread spectrum (DS/SS) system using packet transmission is one of the prevailing schemes, for its robustness against man-made noise and interference from other systems.

The delay spread of indoor multi-path environment is usually very small (of the order of 25–50 ns) [1]–[3]; therefore, the incoming paths of the signals transmitted are hard to be resolved. In this situation, the signal may encounter flat fading, which causes the drop of the almost all received power. Furthermore, the speed of indoor fading is usually slow, and thus degradation of the received power continues for a long period. In addition to the fading, if furniture or factory machines interrupt the paths between the transmitter and the receiver, the signal may encounter strong attenuation called shadowing. These slow flat fading and shadowing phenomena are known to be the major factors responsible for performance degradation of indoor radio systems [4], [5].

In fading and shadowing environments, the initial code acquisition of a DS/SS signal is difficult to achieve. Especially in packet DS/SS systems, the failure to establish the initial code acquisition during the preamble of a packet often dominates overall performance. Thus the improvement of code acquisition performance of indoor DS/SS packet systems may represent an important technical challenge.

As an effective countermeasure against fading and shadowing, the introduction of space diversity techniques is a good candidate. These techniques can be categorized into two groups: micro diversity and macro diversity. The micro diversity is based on the independence of fading at different antennas, while the macro diversity is used to mitigate the effect of shadowing using several radio-ports with large separations [6].

In our former studies [7]–[10], we introduced a micro diversity technique to the initial code acquisition; while many studies on diversity techniques have discussed only the improvement of error performance. In [7], we proposed the introduction of the micro diversity at the receiver, and in [8]–[10], we also discussed micro diversity for the transmitting end, where several antennas at a transmitter are used to obtain different paths in addition to the antenna diversity for reception. In this paper, following these results, we introduce macro diversity at the receiver of a base station of indoor wireless communication systems, and confirm the performance improvement in the initial code acquisition of a DS/SS signal under both shadowing and fading.

2. System Model

Figure 1 illustrates the system model of the indoor wire-

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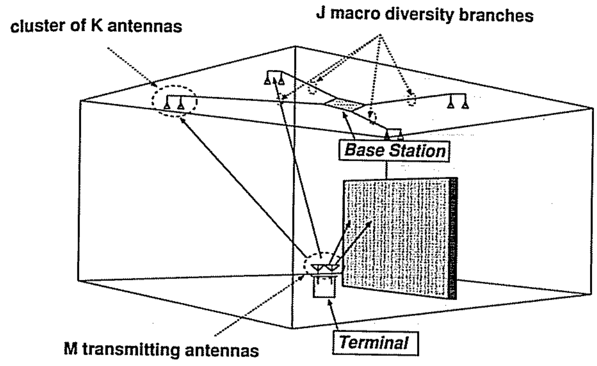

Fig. 1 System model.

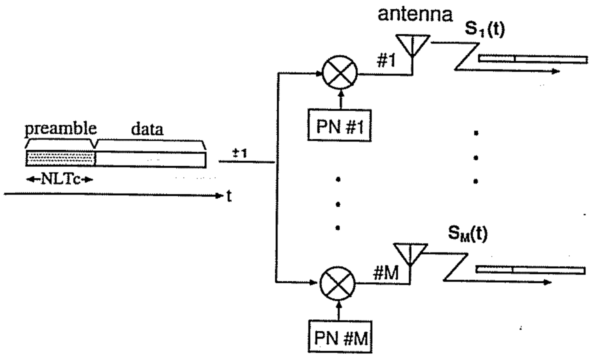

Fig. 2 Transmitter model.

less packet communication system discussed in this paper. We consider the uplink; i.e., from a terminal to a base station. The terminal has M antennas and the base station has J macro diversity branches each with a cluster of K antennas. These M antennas are spaced by several wavelengths of the carrier, and so are K antennas in a cluster. The channel between a pair of transmitting and receiving antennas is modeled by Rayleigh fading whose average power is fluctuated by Lognormal shadowing. When the antenna separation is more than several wavelengths of the carrier, correlations of fading at each antenna is negligibly small [2], [11], thus we assume that each path between transmitting and receiving antennas is influenced by statistically independent fading. We also assume that all the receiving antennas in the same cluster suffer from the same level of shadowing and that the distance between each cluster is large enough to make the statistic of shadowing of different branches independent.

The transmitter proposed in this paper is shown in Fig. 2. At the transmitter, incoming data stream is divided into blocks to form packets. And each packet is transmitted from a set of M different antennas simultaneously. In order to distinguish these signals, different transmitting antennas use different PN codes. As shown in Fig. 2, each packet has preamble prior to data sequence and initial code acquisition is performed within the preamble, which has no data modulation.

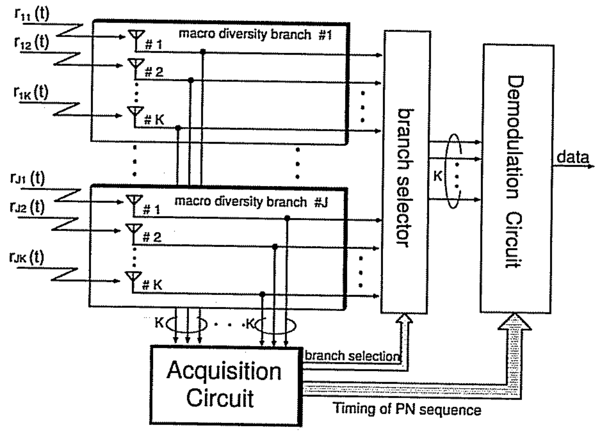

Fig. 3 Receiver model.

Within the preamble, PN code of length L chips is repeated N times.

The transmitted signal from the m-th transmitting antenna during the preamble is given by

$$s_m(t) = \sqrt{2S/M}c_m(t)\cos\omega_0 t, \tag{1}$$

where S is the total transmitting power, $c_m(\cdot)$ is the PN code assigned for the m-th transmitting antenna, and ω_0 is the angular carrier frequency. The power at each transmitting antenna is normalized by $1/M$ to keep the total power the same.

For the reception of this uplink signal, the receiver at the base station has structure as shown in Fig. 3. The received signal (during preamble) at the k-th receiving antenna in the j-th macro diversity branch of the receiver is given by

$$\begin{aligned} r_{jk}(t) = &\sum_{m=1}^{M} \sqrt{2S/M}\beta_{mjk}c_m(t-(\nu_j+\varepsilon_j)T_c) \\ &\cdot\cos(\omega_0 t-\theta_{mjk})+n_{jk}(t), \end{aligned} \tag{2}$$

where β_{mjk} is the channel factor including shadowing loss and fading attenuation between the m-th transmitting antenna and the k-th antenna of the j-th macro diversity branch. The initial phase offset of the PN codes at the j-th branch is $(\nu_j+\varepsilon_j)T_c$; where T_c is a chip duration, ν_j is a integer in the range of $0 \le \nu_j < L$, and ε_j is a constant of $0 \le \varepsilon_j < 1$. The initial phase of the received carrier of the m-th transmit antenna received at the k-th antenna of the j-th branch is expressed as θ_{jmk}. Additive white Gaussian noise (AWGN) at the k-th antenna of the j-th branch is $n_{jk}(t)$, which has zero mean and one-sided spectral density of N_0.

Since the separations between M transmitting antennas, and also between K receiving antennas in a cluster, are several wavelengths of the carrier, the differences of delay offset of each path are much smaller than a chip duration. (For example, if chip rate is 10 MHz, the distance that radio wave of 2.4 GHz travels within a chip duration is 30 m, while the wavelength of the

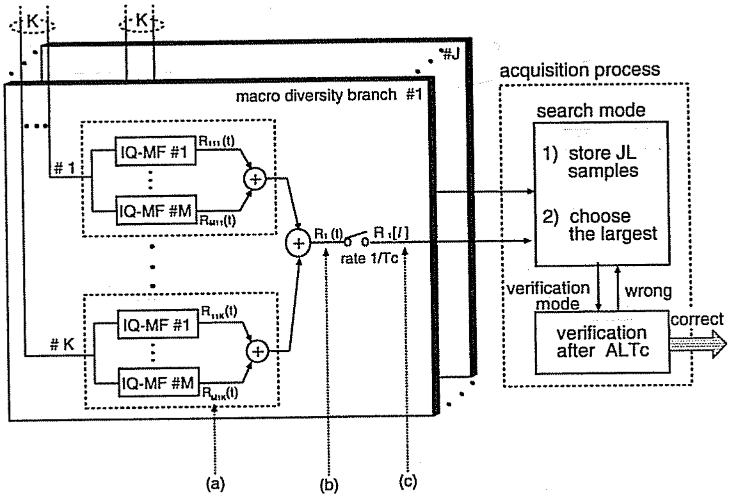

Fig. 4 Acquisition circuit.

carrier is only 0.125 m.) Thus, all the signals from M transmitting antennas can be assumed to arrive simultaneously at K receiving antennas of a macro diversity branch. On the contrary, the delay times between the transmitter and different radio-ports may not be the same, especially in big factories. Hence, in this manuscript, the phase offset of PN code is assumed to be common for all m and k but different for j, and expressed as $(\nu_j + \varepsilon_j)T_c$.

Since the fading characteristics of different paths between transmitting and receiving antennas are modeled to be mutually independent, the channel factor β_{mjk} is independently and identically distributed (i.i.d.) Rayleigh random variable with the probability density function (p.d.f.),

$$P_\beta(x|Z_j) = \frac{2x}{Z_j}\exp\left(\frac{-x^2}{Z_j}\right) \quad x \geq 0 \quad \text{for all} \quad m, j, k, \tag{3}$$

where $Z_j = \sum_{m=1}^{M}\sum_{k=1}^{K}\beta_{mjk}^2/MK$. The value Z_j fluctuates due to shadowing, and it is assumed to have i.i.d. Lognormal distribution given by

$$P_{Z_j}(x) = \overline{Z_j}10^{\frac{g(x;\sigma_s)}{10}}, \tag{4}$$

where $\overline{Z_j}$ is the ensemble average of Z_j over shadowing statistics which include propagation loss due to distance between the transmitter and each cluster at the receiver. The notation $g(x;\sigma_s)$ represents the p.d.f. of Gaussian distributed random variables with zero mean and the variance σ_s^2. In this paper, we assume that average propagation loss $\overline{Z_j}$ is the same for all j: it means that the transmitter is placed at the center of the clusters. Since the fading speed is often very slow in indoor environment, the attenuation by fading and shadowing is assumed to be constant during the code acquisition procedure.

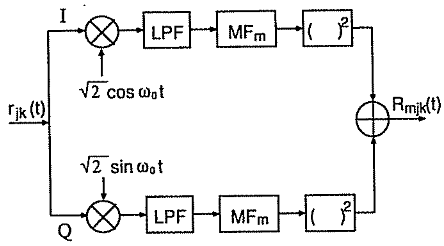

Fig. 5 The m-th IQ-MF.

3. Acquisition Circuit

3.1 Decision Variables for Code Acquisition

As described in Sect. 2, the delay offset of each path of the same macro-diversity branch is assumed to be the same. Therefore, the signals in each macro diversity branch can be combined to find the code timing. On the contrary, the difference of the delay offsets for different macro-diversity branches may not be small compared to the chip duration. Therefore, the estimation of relative delay of each branch is necessary for the combining macro-diversity. But this requirement contradicts the fact that code timing is not known at the code acquisition phase. For this reason, we use selective macro-diversity.

The proposed acquisition circuit is shown in Fig. 4, and the IQ-MF in this figure is shown in Fig. 5. In Fig. 5, MF_m is the filter matched to the m-th PN code.

In the acquisition circuit (Fig. 4), the output of each antenna is fed to a bank of IQ-matched filters corresponding to M PN codes used at the transmitter. The output of the m-th IQ-MF for the k-th antenna in the j-th macro diversity branch ((a) of Fig. 4) is represented as

$R_{mjk}(t)$

$$= \left(\sqrt{S/M}\beta_{mjk}\Lambda_m(t-(\nu_j+\varepsilon_j)T_c)\cos\theta_{mjk} + n^I_{mjk}(t)\right)^2 + \left(\sqrt{S/M}\beta_{mjk}\Lambda_m(t-(\nu_j+\varepsilon_j)T_c) \cdot \sin\theta_{mjk} + n^Q_{mjk}(t)\right)^2, \quad (5)$$

where $n^I_{mjk}(t)$ and $n^Q_{mjk}(t)$ are the inphase and quadrature noise components at the m-th IQ-MF for the k-th antenna in the j-th macro diversity branch, and $\Lambda_m(\cdot)$ is the auto-correlation function of m-th PN code. Note that the system considered is not CDMA, and every user may share the same PN code set. (Colliding packets of different users are lost.) In addition, signals from M different antennas of a user are transmitted synchronously. Therefore, it is not difficult to find a set of PN codes having good auto-correlation and cross-correlation properties. Thus we assume that the auto-correlation function is a simple triangle given by

$$\Lambda_m(\tau) = \int_0^{LT_c} c_m(\delta)c_m(\delta-\tau)d\delta = \begin{cases} L(T_c-|\tau|) & |\tau| \le T_c \\ 0 & T_c < |\tau| \end{cases} \text{ for all } m, \quad (6)$$

and that the cross correlation between the different PN codes of the same user is approximated to be zero.

The output of each IQ-MF at the same timing is summed up to yield the branch output,

$$R_j(t) = \sum_{k=1}^{K}\sum_{m=1}^{M} R_{mjk}(t) \quad (7)$$

at (b) of Fig. 4. These signals are then sampled and fed to the code acquisition logic ((c) of Fig. 4). For simplicity, we assume that the samples are taken at the center of each chip. Thus the ℓ-th sample of the j-th branch becomes $R_j[\ell] = R((\nu_j+\varepsilon_j)T_c)$.

3.2 Acquisition Process

The code acquisition for the j-th branch can be interpreted as the procedure to find ℓ which satisfies "$\ell \equiv \nu_j$, (mod L)." In this paper, this timing is named "sync-timing." The algorithm of code acquisition, which also finds the best branch, consists of search and verification modes as follows.

1) The search mode employs the parallel search strategy. The samples of J branches for LT_c seconds, thus JL samples in total, are stored in a memory.
2) The largest one among the JL samples is selected and tentatively considered to be corresponding to the sync-timing of its branch, then the acquisition system is turned to the verification mode to test this hypothesis.
3) In the verification mode, the above hypothesis is examined with the received signals at the branch for A-bit duration (ALT_c second). If the sample is verified to be of the sync-timing, the branch is selected and used for demodulation using the code timing, otherwise the system goes back to the search mode, 1). In this paper, we assume that the verification mode is ideal [12], [13].

As the result of the above procedure, the acquisition circuit selects one of J branches to be used for data demodulation and provides the code timing of the branch.

3.3 Probability Distribution of the Samples

The performance of code acquisition depends on the statistical behavior of each samples, $R_j[\ell]$. Thus we derive its p.d.f. in this section.

Let IQ-MF output at the sampling timings be expressed as $R_{mjk}[\ell] = R_{mjk}((\ell+\varepsilon_j)T_c)$, then its p.d.f. is given by [14],

$$P'(x|m,j,k,\ell) = \begin{cases} f_{NC\chi^2}(x,\sigma_n^2,a^2_{mjk},1) & \text{if } \ell \equiv \nu_j \pmod{L}, \\ f_{C\chi^2}(x,{\sigma_n}^2,1) & \text{otherwise} \end{cases} \quad (8)$$

where

$$a^2_{mjk} = \frac{{\beta_{mjk}}^2L^2{T_c}^2S}{M}, \quad {\sigma_n}^2 = \frac{N_0LT_c}{2}. \quad (9)$$

The functions $f_{NC\chi^2}(x,\cdot,\cdot,n)$ and $f_{C\chi^2}(x,\cdot,n)$ represent noncentral and central chi-square distribution with $2n$ degrees of freedom, and are expressed as follows [15].

$$f_{NC\chi^2}(x,\sigma^2,s^2,n) = \frac{1}{2\sigma^2}\left(\frac{x}{s^2}\right)^{\frac{n-1}{2}} \cdot \exp\left(-\frac{x+s^2}{2\sigma^2}\right)I_{n-1}\left(\sqrt{x}\frac{s}{\sigma^2}\right) \quad (10)$$

$$f_{C\chi^2}(x,\sigma^2,n) = \frac{1}{(2\sigma^2)^n(n-1)!}x^{n-1}\exp\left(-\frac{x}{2\sigma^2}\right). \quad (11)$$

In the above equations, $I_{n-1}(x)$ is the $(n-1)$th-order modified Bessel function of the first kind.

From (7), $R_j[\ell]$ is the sum of MK i.i.d. chi-square random variables of 2 degrees of freedom, i.e. $R_{mjk}[\ell]$. Thus the p.d.f. of the sample $R_j[\ell]$ can be expressed by chi-square distribution with $2MK$ degrees of freedom as follows.

$$P_R(x|j,\ell) = \begin{cases} f_{NC\chi^2}(x,{\sigma_n}^2,a_j^2,MK) & \text{if } \ell \equiv \nu_j \pmod{L} \\ f_{C\chi^2}(x,{\sigma_n}^2,MK) & \text{otherwise} \end{cases} \quad (12)$$

where

$$a_j^2 = \sum_{m=1}^{M}\sum_{k=1}^{K} \frac{{\beta_{mjk}}^2 L^2 {T_c}^2 S}{M} = \frac{\alpha_j L^2 T_c^2 S}{M}, \qquad (13)$$

and

$$\alpha_j = \sum_{m=1}^{M}\sum_{k=1}^{K} \beta_{mjk}^2. \qquad (14)$$

Since each β_{mjk} is i.i.d. Rayleigh distributed random variable with average Z_j, the value α_j becomes central chi-square random variable with $2MK$ degrees of freedom, and its p.d.f. conditioned on Z_j is represented as

$$P_\alpha(x|Z_j) = f_{C\chi^2}(x, Z_j/2, MK). \qquad (15)$$

4. Performance Analysis

4.1 Measures of Performance

In this paper, we evaluate the performance of the proposed system by the following three measures.

1) *Mean Acquisition Time*
Mean acquisition time has been most widely used as the measure of performance of acquisition schemes. This is the expectation of the time needed to acquire the timing of PN code when the preamble length is enough large.
2) *Misacquisition Probability: P_{macq}*
In packet radio communications, the acquisition must be completed within a preamble of a packet, otherwise the packet will be lost. Thus, as a performance measure, we use the misacquisition probability of the occurrence of the event that the acquisition circuit cannot acquire the timing of PN code within a given preamble length, NLT_c.
3) *Required Preamble Length: N_{req}*
After the establishment of the code acquisition, (residual) preamble does not contribute to the communication. Thus, from the viewpoint of the efficient use of communication channel, the preamble length should be as short as possible, while it has to be long enough to maintain good acquisition performance. For this reason, we employ another performance measure, the preamble length required to guaranteed the establishment of code acquisition with the misacquisition probability lower than a given threshold value "$P_{macq} < \Theta$."

4.2 Mean Acquisition Time

In this subsection, we derive the mean acquisition time of the proposed system in a similar way as [14]. The state transition diagram of code acquisition process is shown in Fig. 6. The state "S" represents the condition that the acquisition circuit is searching/verifying the samples $R_j[\ell]$ to find the sync-timing, and the state "Acq", which is the sole absorbing state, represents the condition that the acquisition is completed. In this figure, $G_1(z)$ denotes the generating function of the correct decision through the search mode and the verification mode, and $G_2(z)$ denotes the generating function of the event that the candidate of the search mode is rejected in the verification mode. They are given by

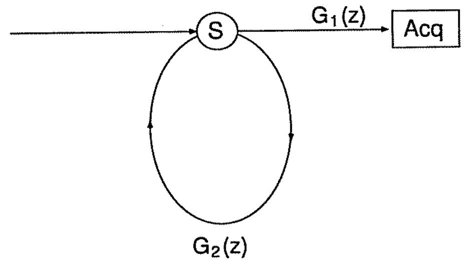

Fig. 6 Transition state diagram.

$$G_1(z) = P_D z^{(A+1)LT_c} \qquad (16)$$
$$G_2(z) = (1-P_D) z^{(A+1)LT_c}, \qquad (17)$$

where P_D is the detection probability which denotes the probability of the event that the selected sample in the search mode corresponds to a correct sync-timing of its macro diversity branch.

Using these functions, the generating function of the acquisition is obtained as

$$G(z) = G_1(z) + G_2(z)G_1(z) + {G_2}^2(z)G_1(z) + \cdots = \frac{G_1(z)}{1-G_2(z)}. \qquad (18)$$

The acquisition time T_{acq} is a random variable due to noise, and its ensemble average over the noise is given by

$$\overline{T_{acq}} = \frac{d}{dz}\ln G(z)|_{z=1}, \qquad (19)$$

and with (16)-(18),

$$\overline{T_{acq}} = \frac{1+A}{P_D} LT_c. \qquad (20)$$

1) $J=1$
When the base station has only one branch, the detection probability P_D in (20) is represented as follows. (Since P_D is conditioned by α_1, let us express it as $P_D(\alpha_1)$.)

$$P_D(\alpha_1) = \int_0^\infty P_R(y|1, \ell \equiv \nu_1) \cdot \left[\int_0^y P_R(x|1, \ell \not\equiv \nu_1)dx\right]^{L-1} dy = \int_0^\infty f_{NC\chi^2}\left(y, {\sigma_n}^2, \frac{\alpha_1 L^2 T_c^2 S}{M}, MK\right)$$

$$\cdot \left[\int_0^y f_{C\chi^2}(x, \sigma_n{}^2, MK) dx \right]^{L-1} dy. \quad (21)$$

With this probability, the mean acquisition time is represented as

$$\mathrm{E}[\overline{T_{acq}}] = \int_0^\infty \int_{-\infty}^\infty \frac{(1+A)LT_c}{P_D(\alpha_1)} \cdot P_\alpha(\alpha_1|Z_1) P_Z(Z_1) d\alpha_1 dZ_1, \quad (22)$$

where $\mathrm{E}[x]$ denotes the average over the attenuation due to fading and shadowing.

2) $J \geq 2$

When the macro diversity is in use, the detection probability P_D of (20) is represented as

$$P_D(\alpha_1, \ldots, \alpha_J) = \sum_{j=1}^J P_j(\alpha_1, \ldots, \alpha_J), \quad (23)$$

where $P_j(\alpha_1, \ldots, \alpha_J)$ is the probability of event that the sample corresponding to sync-timing at the j-th macro diversity branch is larger than other $JL-1$ samples.

This event can be expressed as

$$\left[\bigwedge_{i=1}^J \bigwedge_{\substack{\ell=1 \\ \ell \neq \nu_i}}^L R_i[\ell] < R_j[\nu_j] \right] \bigwedge \left[\bigwedge_{\substack{i=1 \\ i \neq j}}^J R_i[\nu_j] < R_j[\nu_j] \right], \quad (24)$$

and the probability of the occurrence of this event is represented as

$$\begin{aligned} &P_j(\alpha_1, \ldots, \alpha_J) \\ &= \int_0^\infty P_R(y|j, \ell \equiv \nu_j) \left[\int_0^y P_R(x|i, \ell \not\equiv \nu_i) dx \right]^{J(L-1)} \\ &\quad \cdot \prod_{\substack{i=1 \\ i \neq j}}^J \left[\int_0^y P_R(x|i, \ell \equiv \nu_i) dx \right] dy \\ &= \int_0^\infty f_{NC\chi^2}(y, \sigma_n{}^2, \frac{\alpha_j L^2 T_c^2 S}{M}, MK) \\ &\quad \cdot \left[\int_0^y f_{C\chi^2}(x, \sigma_n{}^2, MK) dx \right]^{J(L-1)} \\ &\quad \cdot \prod_{\substack{i=1 \\ i \neq j}}^J \left[\int_0^y f_{NC\chi^2}(x, \sigma_n{}^2, \frac{\alpha_i L^2 T_c^2 S}{M}, MK) dx \right] dy. \end{aligned} \quad (25)$$

Note that every P_j and thus P_D are conditioned by $(\alpha_1, \ldots, \alpha_J)$. Then, the mean acquisition time can be denoted as

$$\mathrm{E}[\overline{T_{acq}}] = \int \cdots \int \frac{(1+A)LT_c}{P_D(\alpha_1, \ldots, \alpha_J)} \prod_{j=1}^J P_\alpha(\alpha_j|Z_j) \cdot P_{Z_j}(Z_j) d\alpha_1 \cdots d\alpha_J dZ_1 \cdots dZ_J. \quad (26)$$

4.3 Misacquisition Probability and Required Preamble Length

The misacquisition probability P_{macq} is derived as

$$\begin{aligned} &P_{macq} \\ &= \int \cdots \int (1 - P_D(\alpha_1, \ldots, \alpha_J))^{\frac{N}{1+A}} \\ &\quad \cdot \prod_{j=1}^J P_\alpha(\alpha_j|Z_j) P_{Z_j}(\alpha_j) d\alpha_1 \cdots d\alpha_J dZ_1 \cdots dZ_J. \end{aligned} \quad (27)$$

The required preamble length N_{req} is obtained by increasing N until "$P_{macq} < \Theta$" is fulfilled.

5. Numerical Examples

In this section, numerical examples derived under the following conditions are shown.

- Length of PN code: $L = 63$.
- Time required in the verification mode: $ALT_c = 4LT_c$.
- Standard deviation of shadowing attenuation in dB: $\sigma_s = 8$. [5]

The average chip energy E_c is defined as the mean value of the received power at all the macro diversity branches within one chip duration, and the value is $E_c = \mathrm{E}[ST_c Z_j] = ST_c \overline{Z_j}$.

The mean acquisition time of the proposed system is shown in Fig. 7. The misacquisition probability of the proposed system for 40 bits preamble is shown in Fig. 8, and the required preamble length of the proposed system for the required acquisition probability $1-\Theta$ of 99.9% is shown in Fig. 9.

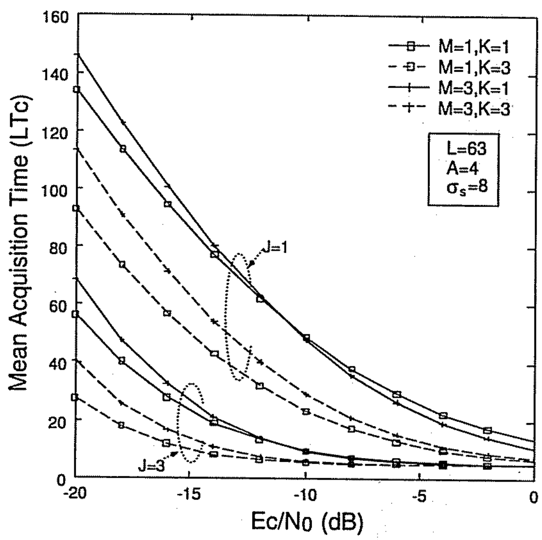

Fig. 7 Mean acquisition time.

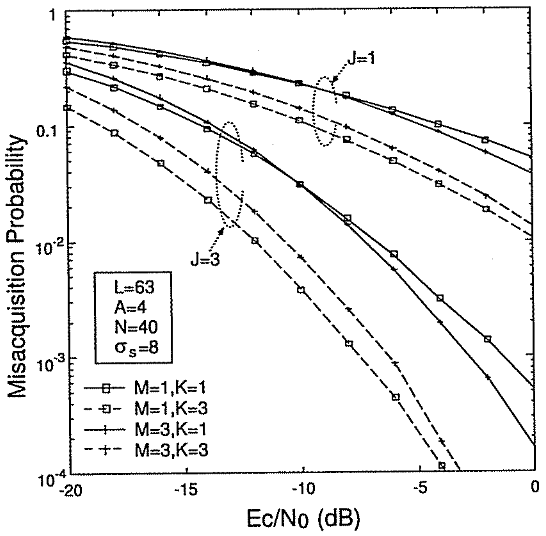

Fig. 8 Misacquisition probability.

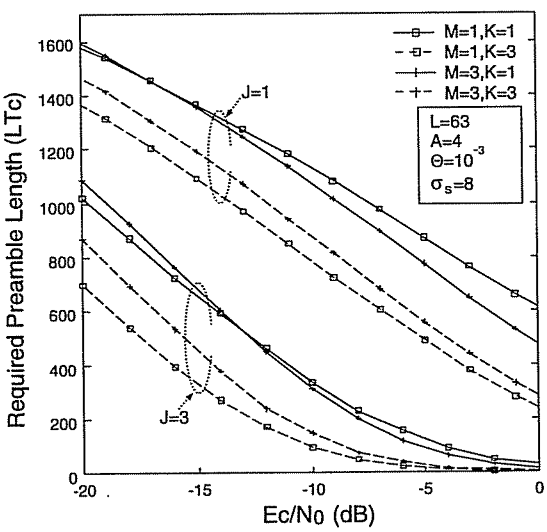

Fig. 9 Required preamble length.

From these figures, it can be seen that both macro and micro diversity schemes at the receiving end offer substantial performance improvement.

On the contrary, the transmit antenna (micro) diversity improves the performance slightly only under large E_c/N_0 in the absence of macro and micro diversity at the receiver. This is the result of the trade-off between the diversity gain for Rayleigh fading and the noncoherent combining loss due to the dispersion of the transmitted power. When the receive antenna diversity is in use or when E_c/N_0 is small, the combining loss dominates the performance. Moreover, if the cross-correlation property between the codes of different transmitting antennas is not perfect, the transmit antenna (micro) diversity faces additional performance degradation, i.e., interference between the signals of different transmit antennas. Considering these facts, we can conclude that the transmit diversity at the terminal is not effective countermeasure to improve code acquisition performance.

Comparing both receive antenna diversity schemes, we find that the macro diversity surpasses micro diversity in diversity gain. This is because both fading and shadowing components at the antennas of different macro diversity branches are independent, while all antennas of the same branch suffer from the same shadowing attenuation. Note that the combination of the two receive antenna diversity schemes improves performance, unlike the combination of transmit and receive antenna diversity. Thus, if the number of receiver antenna is given, macro diversity can be the solution. And if it is difficult to arrange a large number of macro diversity branches, the combination of macro and micro receive antenna diversity is the alternative.

In this manuscript, three different performance measures are employed. Among them, the mean acquisition time of Fig. 7 is the most popular for the evaluation of code acquisition performance. However, in packet radio systems, shorter mean acquisition time does not always results in better performance, since the requirement is to establish code acquisition during the provided fix preamble length. For this reason, the misacquisition probability shown in Fig. 8 is the more suitable measure of performance, once system parameters are given. On the other hand, if we need to design the system parameters such as the structure of a packet, the required preamble length is a useful measure. It is interesting that the required preamble length shown in Fig. 9 is much larger than the mean acquisition time, Fig. 7 because of a long tail in the p.d.f. of T_{acq}. This fact suggests that the mean acquisition time may underestimate the preamble length necessary to establish the code acquisition.

6. Conclusion

In this paper, we have introduced macro antenna diversity for the base station receiver of an indoor DS/SS packet system together with micro transmit and receive antenna diversity. From numerical examples, we have shown that the both receive antenna diversity techniques achieve large performance improvements on code acquisition under slow/flat Rayleigh fading and Lognormal shadowing, while transmit micro diversity is not effective. In addition, it is shown that the required preamble length for initial code acquisition is much longer than the mean acquisition time, which is often used as the measure of code acquisition performance of DS/SS communication systems.

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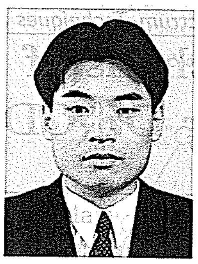

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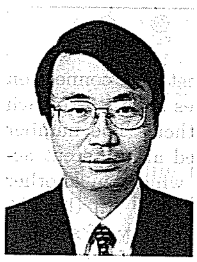

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