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PAPER *Special Section on Spread Spectrum Techniques and Applications*

Code Acquisition of a DS/SS Signal with Transmit and Receive Antenna Diversity

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SUMMARY In this paper, we propose the introduction of space diversity techniques to the code acquisition of a direct-sequence spread-spectrum signal. In this scheme, both a transmitter and a receiver have multiple antennas and the signals corresponding to all the combinations of the transmitter and receiver antennas are combined at the acquisition circuit of the receiver. The performance is evaluated for an indoor packet radio communication system from the viewpoints of the average time for acquisition, the probability of success of acquisition, and the necessary preamble length. As the result, we show great performance improvements by the proposed scheme under slow and flat Rayleigh fading environment.

key words: spread-spectrum, acquisition, indoor radio transmission, diversity, fading

1. Introduction

The focus of this paper is the introduction of space diversity techniques to the code acquisition of a direct-sequence spread-spectrum (DS/SS) signal. Though the the space diversity techniques have attractive features in the fading channel, these have not been considered well for the code acquisition of a DS/SS signal. This paper aims to investigate the acquisition scheme utilizing the space diversity techniques and reveal its advantages for indoor packet radio communication systems.

In indoor environment, the radio waves reflected from walls and furniture are relatively strong and they may cause multi-path fading, which is one of the most important factors of performance degradation. The delay spread of indoor environment is often in the range of 25–50 ns [1]–[3]. In this condition, if we transmit a signal with a moderate chip rate, 1 MHz for example, the delay spread is much smaller than the chip duration, and each wave of the multi-path fading cannot be resolved. In this situation, we may have flat fading which causes the drop of the almost all received power. Furthermore, the speed of indoor fading is often very slow, so the received power might be degraded for long period.

In this slow and flat fading environment, receive antenna diversity and transmit antenna diversity [4]–[6] are known to be effective to improve the communication quality such as bit error rate. But, these studies discuss the performance provided that the code acquisition is already performed, and the space diversity techniques for the establishment of the code acquisition itself is not considered.

Recently, there appear the proposals of the employment of the space diversity technique for the code acquisition of a DS/SS signal [7], [8]. They demonstrate the improvement of code acquisition performance by the receive antenna diversity, but still not with the transmit antenna diversity.

In this paper, we introduce the transmit antenna diversity to the code acquisition of a DS/SS signal together with the receive antenna diversity. We evaluate the code acquisition performance from the viewpoints of average time of acquisition, probability of success or failure of acquisition, and necessary preamble length. The latter two measures have not been used in the former works, but are practical measures for packet communication systems. By the numerical results, we show the great performance improvements by the transmit and receive antenna diversity techniques under slow and flat Rayleigh fading environment.

2. System Model

The transmitter and the receiver proposed in this paper are shown in Figs. 1 and 2. The transmitter and the receiver have M and K antennas, respectively. It is assumed that each transmitter antenna is spatially separated from others by several wavelengths of the carrier, and this assumption is also applied to the receiver antennas. Since the correlation of the fading between each pair of transmitter and receiver antennas can be made small enough with reasonable antenna separation at the transmitter or the receiver [9], the statistics of the fading between each pair is assumed to be independent.

The transmitter uses a set of M different PN codes of a period L to identify each antenna. The transmitted signal from the m-th transmitter antenna is given by

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

where S is the total transmit power, $c_m(\cdot)$ is the PN code assigned for the m-th transmitter antenna, and ω_0 is the angular carrier frequency. The transmit power

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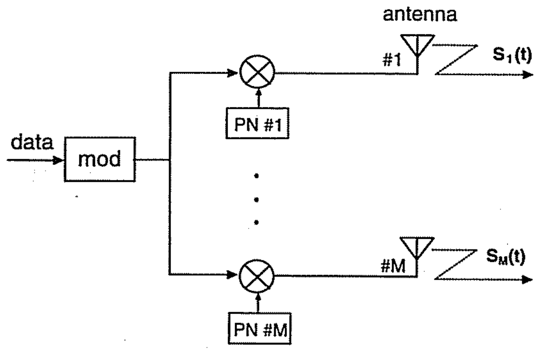

Fig. 1 Transmitter model.

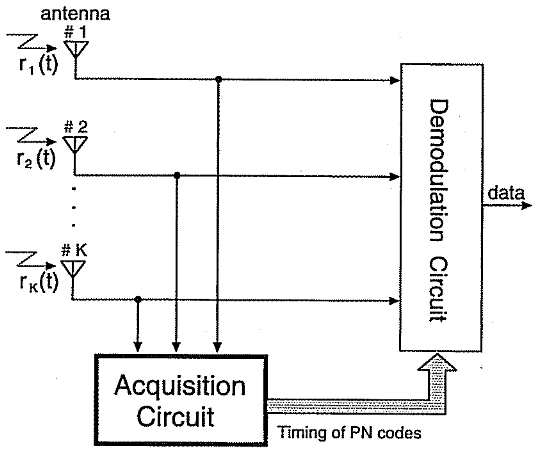

Fig. 2 Receiver model.

at each transmitter antenna is normalized by $1/M$ to keep the total transmit power unity. Since the code acquisition is performed within the period prior to data modulation (preamble), (1) does not have the term for data modulation.

The received signal at the k-th receiver antenna is given by

$$r_k(t) = \sum_{m=1}^{M} \sqrt{2S/M}\beta_{mk}c_m(t-\zeta T_c) \cdot \cos(\omega_0 t - \theta_{mk}) + n_k(t), \quad (2)$$

where β_{mk} is fading attenuation between the m-th transmitter antenna and the k-th receiver antenna. The initial phase offsets of the PN code and the carrier at the receiver are denoted as ζT_c and θ_{mk}, respectively. Additive white Gaussian noise (AWGN) is denoted as $n_k(t)$, which has zero mean and one-sided spectral density of N_0. Since every pair of transmitter and receiver antennas have approximately the same distances, it can be assumed that all the signals arrive at K receiver antennas simultaneously. Hence, the phase offset of PN code ζ is common for all over the M and K antennas. In this paper, we assume Rayleigh fading. This is often used as the channel model without line of sight (LOS) signal. But in indoor systems, this model can be used with and without the LOS signal, because the reflected signals may have a power as large as the LOS signal. The fading attenuation of each pair of antennas is assumed to be mutually independent, and then β_{mk} is an independently and identically distributed (i.i.d.) Rayleigh random variable. Then its probability density function (pdf) becomes [10]

$$P_{\beta_{mk}}(x) = 2x\exp\left(-x^2\right) \quad x \geq 0 \quad \text{for all} \quad m, k \quad (3)$$

if the mean square value of β_{mk} is normalized to be unity. In this paper, we assume that the fading attenuation is constant during the code acquisition.

All the signals from K receiver antennas are fed to the demodulator to reproduce transmitted data with the help of the acquisition circuit described in the next section.

3. Acquisition Circuit

The proposed acquisition circuit is shown in Fig. 3, and the IQ-MF in the figure is shown in Fig. 4. Since we consider a packet radio communication scheme, the acquisition process is expected to be completed during the preamble of a packet which has no data modulation.

3.1 Signals at Each Stage of Acquisition Circuit

The output of the IQ-MF corresponding to the m-th transmitter and the k-th receiver antenna at (a) of Fig. 3 is represented as

$$\begin{aligned} R_{mk}(t) &= \left(\sqrt{S/M}\beta_{mk}\Lambda_m(t-\zeta T_c)\cos\theta_{mk} + n_{mk}^I(t)\right)^2 \\ &\quad + \left(\sqrt{S/M}\beta_{mk}\Lambda_m(t-\zeta T_c)\sin\theta_{mk} + n_{mk}^Q(t)\right)^2, \end{aligned} \quad (4)$$

where $n_{mk}^I(t)$ and $n_{mk}^Q(t)$ is the noise components at I and Q branches of IQ-MF respectively, and $\Lambda_m(\cdot)$ is the auto-correlation function of m-th PN code. For simplicity, we approximate that the auto-correlation function is a simple triangle function given by

$$\begin{aligned} \Lambda_m(\tau) &= \int_0^{LT_c} c_m(\delta)c_m(\delta-\tau)d\delta \\ &= \begin{cases} L(T_c - |\tau|) & |\tau| \leq T_c \\ 0 & T_c < |\tau| \end{cases} \quad \text{for all } m, \end{aligned} \quad (5)$$

and that the cross correlation between the different PN codes are zero. In this paper, we use the term "sync-timing" to refer the timing that $t = \zeta T_c$, which is common for all m, k.

The output of each IQ-MF is then summed up and we have

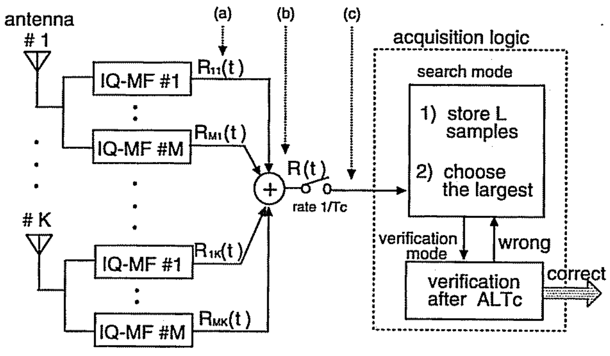

Fig. 3 Acquisition circuit.

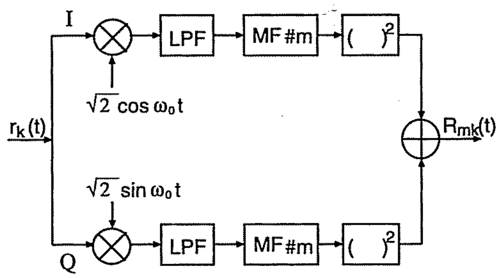

Fig. 4 The m-th IQ-MF.

$$R(t) = \sum_{m=1}^{M} \sum_{k=1}^{K} R_{mk}(t) \tag{6}$$

at Fig. 3 (b). This signal is sampled to be fed to the code acquisition logic (Fig.3 (c)). For simplicity, we assume that the sampling is made at the center of each chip.

3.2 Acquisition Process

The acquisition process in the acquisition logic of Fig. 3 has two modes, i.e., the search mode and the verification mode. Acquisition is performed based on the following algorithm.

1) The search mode employs the parallel search strategy. The L samples for LT_c seconds are stored in a memory.
2) The largest one among the L samples is selected, and the timing of this is considered, tentatively, to be sync-timing, then the acquisition system is turned to the verification mode to test this hypothesis.
3) In the verification mode, the above hypothesis is examined in A-bit duration (ALT_c second). If the sample is verified to be of sync-timing, acquisition is declared and despreading is performed based on the timing of this sample, otherwise, the system goes back to 1). In this paper, we assume that the verification mode works ideally [11], [12].

3.3 Probability Distribution of the Samples

The probability density function (pdf) of the IQ-MF output at the sampling timings $R_{mk}(nT_c)$ is given by (7) and (8) [13], where H_1 represents the case that the sampling timing corresponds to the sync-timing, and H_0 is the case that the sampling timing does not correspond to the sync-timing.

$$P_{R_{mk}}(x|\mathrm{H}_1) = f_{NC_{\chi^2}}(x, {\sigma_n}^2, a_{mk}^2, 1) \tag{7}$$

$$P_{R_{mk}}(x|\mathrm{H}_0) = f_{C_{\chi^2}}(x, {\sigma_n}^2, 1), \tag{8}$$

where

$$a_{mk}^2 = \frac{{\beta_{mk}}^2 L^2 {T_c}^2 S}{M} \qquad {\sigma_n}^2 = \frac{N_0 L T_c}{2}. \tag{9}$$

In the above equations, $f_{NC_{\chi^2}}(x, \cdot, \cdot, n)$ and $f_{C_{\chi^2}}(x, \cdot, n)$ are the pdf of noncentral and central chi-square distribution with $2n$ degrees of freedom respectively. They are expressed as follows [14].

$$\begin{aligned} &f_{NC_{\chi^2}}(x, \sigma^2, s^2, n) \\ &= \frac{1}{2\sigma^2}\left(\frac{x}{s^2}\right)^{\frac{n-1}{2}} \exp\left(-\frac{x+s^2}{2\sigma^2}\right) I_{n-1}\left(\sqrt{x}\frac{s}{\sigma^2}\right) \end{aligned} \tag{10}$$

$$\begin{aligned} &f_{C_{\chi^2}}(x, \sigma^2, n) \\ &= \frac{1}{(2\sigma^2)^n (n-1)!} x^{n-1} \exp\left(-\frac{x}{2\sigma^2}\right). \end{aligned} \tag{11}$$

In the above equations, $I_\alpha(x)$ is the αth-order modified Bessel function of the first kind, represented by the infinite series

$$I_\alpha(x) = \sum_{k=0}^{\infty} \frac{(x/2)^{\alpha+2k}}{k!\Gamma(\alpha+k+1)}, \tag{12}$$

where $\Gamma(\cdot)$ is the gamma function, defined as

$$\begin{aligned} \Gamma(x) &= \int_0^\infty t^{x-1} e^{-t} dt \qquad && x > 0 \\ \Gamma(x) &= (x-1)! && x: integer, x > 0. \end{aligned} \tag{13}$$

Next, let us consider the pdf of the samples used in the acquisition logic, i.e. $R(nT_c)$. Since the samples at Fig. 3 (c) is the sum of MK i.i.d. chi-square random variables, each with 2 degrees of freedom, the samples $R(nT_c)$ follow chi-square distribution with $2MK$ degrees of freedom as follows.

$$P_R(x|\mathrm{H}_1) = f_{NC_{\chi^2}}(x, {\sigma_n}^2, a^2, MK) \tag{14}$$

$$P_R(x|\mathrm{H}_0) = f_{C_{\chi^2}}(x, {\sigma_n}^2, MK). \tag{15}$$

The parameter a^2 is

$$a^2 = \sum_{m=1}^{M} \sum_{k=1}^{K} \frac{{\beta_{mk}}^2 L^2 {T_c}^2 S}{M}$$

$$= \frac{\alpha L^2 T_c^2 S}{M}, \tag{16}$$

where the value α is the sum of β_{mk}^2 all over the m and k, that is, $\alpha = \sum_{m=1}^{M} \sum_{k=1}^{K} \beta_{mk}^2$. Since each β_{mk} is Rayleigh distributed random variable, the value α becomes central χ^2 random variables with $2MK$ degrees of freedom, and has probability density function

$$P_\alpha(x) = f_{C_{\chi^2}}(x, 1/2, MK). \tag{17}$$

4. Performance Analysis

4.1 Measures of Performance

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

1) *Mean Acquisition Time*
Mean acquisition time has been most widely used as the performance measure 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, or the packet will be lost. Thus, as a new performance measure, we proposed the misacquisition probability that the acquisition circuit cannot acquire the timing of PN code within a given preamble length, NLT_c.

3) *Required Preamble Length: N_{req}*
From the viewpoint of the efficiency of the channel capacity, the preamble length should be as short as possible. Thus, as another new performance measure we introduce the preamble length required to acquire the timing of PN code with the probability more than P_{acq}.

In packet communication systems, the interest is whether the code acquisition is performed within the preamble or not, and the necessary time for acquisition itself is not so important. Thus, in packet communication systems, the latter two measures are more practical than the mean acquisition time.

4.2 Mean Acquisition Time

In this subsection, we derive the mean acquisition time of the proposed scheme in a similar way to [13]. The state transition diagram is shown in Fig. 5. The state "S" represents the condition that the acquisition circuit is detecting 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 that the candidate of the search mode is rejected in the verification mode. $G_1(z)$ and $G_2(z)$ are given by

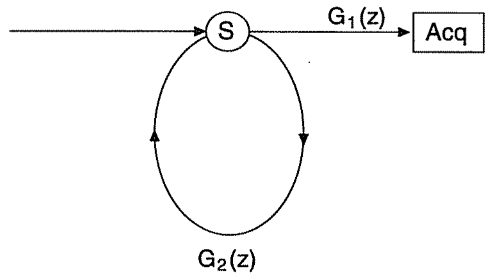

Fig. 5 Transition state diagram.

$$G_1(z) = P_D z^{(A+1)LT_c} \tag{18}$$
$$G_2(z) = (1 - P_D) z^{(A+1)LT_c}, \tag{19}$$

where P_D denotes the probability that the selected sample in the search mode corresponds to the sync-timing.

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

$$G(z) = G_1(z) + G_2(z)G_1(z) + G_2{}^2(z)G_1(z) + \cdots = \frac{G_1(z)}{1 - G_2(z)}. \tag{20}$$

We can see that the acquisition probability $G(1) = 1$ when the length of the preamble is infinite. The acquisition time is a random variable due to noise, and the average of this is given by

$$\overline{T_{acq}} = \frac{d}{dz} \ln G(z)|_{z=1}, \tag{21}$$

where $\overline{x}$ means ensemble average over the noise. With (18)–(20), (21) becomes

$$\overline{T_{acq}} = \frac{1+A}{P_D} LT_c. \tag{22}$$

The detection probability P_D is the probability that the sample corresponding to sync-timing is larger than other $L-1$ samples, and is represented as

$$P_D = \int_0^\infty P_R(y|\mathrm{H}_1) \left[\int_0^y P_R(x|\mathrm{H}_0) dx \right]^{L-1} dy. \tag{23}$$

From (14)–(16), (23) becomes

$$P_D(\alpha) = \int_0^\infty f_{NC_{\chi^2}}(x, \sigma_n{}^2, \frac{\alpha L^2 T_c^2 S}{M}, MK) \cdot \left[\int_0^y f_{C_{\chi^2}}(x, \sigma_n{}^2, MK) dx \right]^{L-1} dy. \tag{24}$$

In this equation, we notice that the detection probability P_D is the function of α in fading channel. In fading channel, the mean acquisition time is represented as

$$E[\overline{T_{acq}}] = \int_0^\infty \frac{1+A}{P_D(x)} LT_c \cdot P_\alpha(x)dx, \qquad (25)$$

where E[x] denotes the average over the attenuation due to fading.

4.3 Misacquisition Probability and Required Preamble Length

The misacquisition probability is represented as

$$P_{macq} = \int_0^\infty P_\alpha(x)\left(1 - P_D(x)\right)^{\frac{N}{1+A}} dx, \qquad (26)$$

and we can find the required preamble length N_{req} by increasing N until $(1 - P_{macq}) > P_{acq}$ is fulfilled.

5. Numerical Examples

We calculate the mean acquisition time, misacquisition probability, and required preamble length of the proposed scheme under the following conditions.

- The length of PN code: $L = 63$.
- Time required in the verification mode: $ALT_c = 4LT_c$.

The mean acquisition time of the proposed scheme is shown in Fig. 6. In this and the following figures, the chip energy E_c is defined as ST_c. From this figure, it can be seen that the receive antenna diversity offers a substantial performance improvement. Figure 6 also shows that the transmit antenna diversity improves the performance when the number of the receiver antenna is one and E_c/N_0 is larger than about -17 dB. This is the result of the tradeoff between the prevention of the drop of the total received power by antenna diversity and the noncoherent combining loss due to the dispersion of the transmitted power. When E_c/N_0 is larger than about -17 dB, the former factor influences larger, and the latter dominates if E_c/N_0 is smaller. If the receiver uses plural antennas, however, the transmit antenna diversity less improves the performance even when E_c/N_0 is large. The reason of this is that when the receive antenna diversity is in use, we can mitigate fading to some extent even when $M = 1$, and performance improvement by transmit antenna diversity does not influences much.

Figure 7 shows the misacquisition probability of the proposed scheme. We assume that the length of the preamble is 40bits. This figure presents again that the receive antenna diversity improves the performance substantially. Not as in Fig. 6, however, the transmit antenna diversity also improves the performance in large E_c/N_0, even when the number of the receiver antenna is more than one. This fact implies that the transmit antenna diversity mitigates the effect of fading especially when the received power drops largely and the acquisition circuit tends to fail the synchronization of PN code. This situation dominates misacquisition probability, and thus the performance improvement is considerably large, while the mean acquisition time is less improved.

In Fig. 8, the required preamble length of the proposed scheme is shown. We assume that the required acquisition probability P_{acq} is 99.9%. In this figure, it can be found again that the receive antenna diversity improves the performance. The interesting result from this figure is that the transmit antenna diversity improves the performance in almost all E_c/N_0, and that the improvement itself is remarkably large, especially when the number of the transmitter antenna increases from one to two. The reason of this performance improvement is the same as for Fig. 7, i.e. transmit antenna diversity improves performance of acquisition especially for the case when the fading is deep, which dominates the required preamble length.

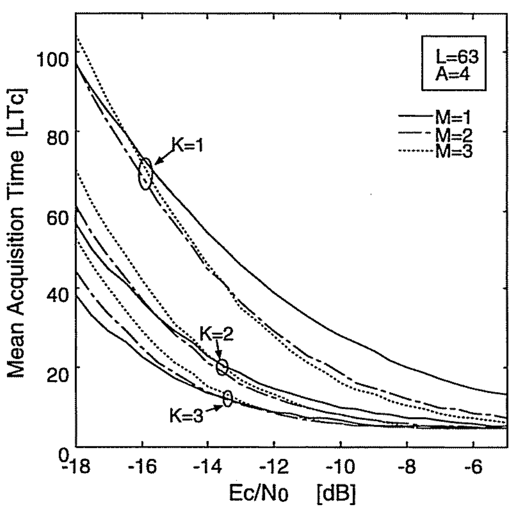

Fig. 6 Mean acquisition time.

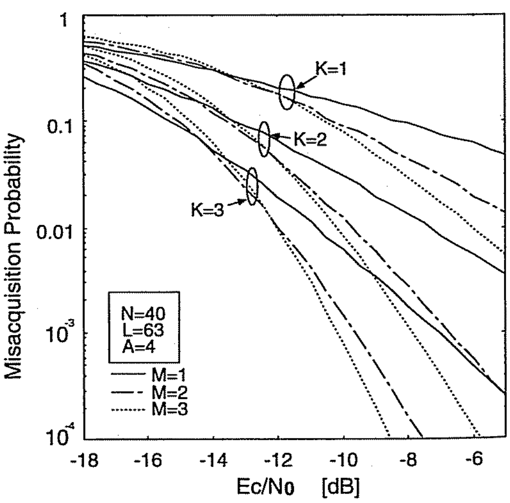

Fig. 7 Misacquisition probability.

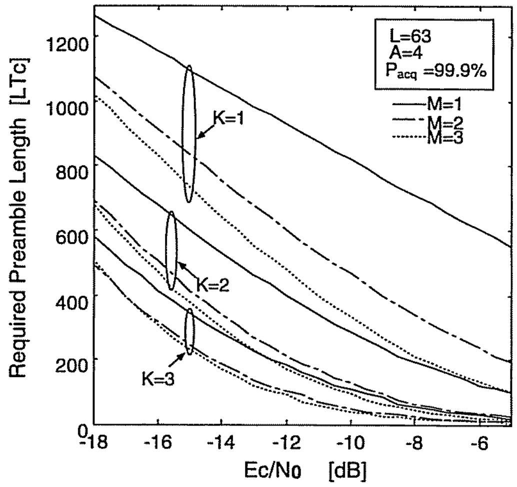

Fig. 8 Required preamble length.

6. Conclusion

In this paper, we considered the introduction of space diversity techniques for the acquisition of PN code of a direct-sequence spread-spectrum signal and proposed a new acquisition scheme with the transmit and the receive antenna diversity.

From the numerical examples, we have shown that the receive antenna diversity is very effective in acquisition of PN code of a DS/SS signal in flat and slow Rayleigh fading channels. Though the performance improvement by transmit antenna diversity is not as large as that by the receive antenna diversity, this technique is still effective in many cases. For example, when the receive antenna diversity cannot be used in downlink as the mobile receiver is too small, the acquisition performance can still be improved using the transmit antenna diversity at the base-station. Also we have found that the transmit antenna diversity is effective in packet communication systems, and the acquisition performance can be much improved using the transmit antenna diversity together with the receive antenna diversity.

Acknowledment

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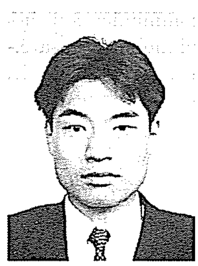

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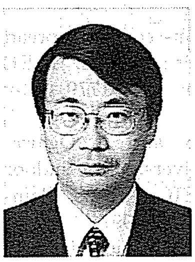

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