PAPER

Effect of Noise-Only-Paths on the Performance Improvement of Post-Demodulation Selection Diversity in DS/SS Mobile Radio

Akihiro HIGASHI†, Tadashi MATSUMOTO† *and* **Mohsen KAVEHRAD††**, *Members*

SUMMARY The path diversity improvement inherent in direct sequence spread spectrum (DS/SS) signalling under multi-path propagation environments is investigated for mobile/personal radio communications systems that employ DPSK modulation. The bit error rate (BER) performance of post-demodulation selection diversity reception is theoretically analyzed in the presence of noise-only-paths in the time window for diversity combining. Results of laboratory experiments conducted to evaluate the BER performance are also presented. It is shown that the experimental results agree well with the theoretical BER.

key words: *spread spectrum, direct sequence, RAKE reception, selection diversity, noise-only-path*

1. Introduction

The path diversity improvement inherent in direct sequence spread spectrum (DS/SS) signaling under multipath propagation environments is very attractive for mobile/personal communications systems. The desired signal components extracted from the received composite signal suffer from independent fading, and these can be used as diversity branches. This is the great advantage of DS/SS signaling and, therefore, receiver systems featuring this path diversity technique have been widely researched previously under "RAKE" combining system.(1)-(3)

Fading environments lead to fast variations in the received signal phase and envelope, and this makes coherent detection very difficult to use. Differential detection requires no carrier recovery function, and is, thus from the practical point of view, promising for mobile radio applications. Even for differential PSK (DPSK), it is still difficult to precisely track the time at which the desired signal components corresponding to a propagation path are received in order to demodulate the desired signals. This is because the propagation profile may vary rapidly, and furthermore, the received signal may suffer from Doppler shifting due to vehicle motion.

The post-demodulation integration RAKE (PDI-RAKE) system can solve this problem by setting a time window so that some of the desired demodulated signals are included. The time window position is determined from the detected correlation peak in the despreading process. The PDI receiver requires no accurate path tracking function such as a delay lock loop or tau-dither loop; a simple digital phase locked loop (DPLL) is applicable to this time window positioning technique.

The PDI receiver combines demodulated signals in the time window. If at any of the chips in the time window the demodulator output has no desired signal component, and if these "noise-only-paths (NOP's)" are combined with equal weight, the combined received signal quality degrades. This NOP effect on the bit error rate (BER) performance of the PDI-RAKE system was analyzed in Refs. (4)-(6). Reference (4) evaluated the BER performance of the PDI-RAKE system with orthogonal signaling in the presence of NOP's; square-law combining was assumed. Reference (5) also investigated the NOP effect on the BER performance of orthogonal signaling with square-law combining, and proposed an adaptive threshold control scheme that mitigates the NOP effect. Reference (6) analyzed the path diversity improvement of the PDI-RAKE system with DPSK modulation, in which an exponential delay profile and a fixed time window size was assumed.

Another simple type of post-demodulation (PD) diversity reception is to select the demodulated signal with the largest signal envelope from among the demodulated signals in each time window. This paper analyzes the effects of NOP's in the time window on the diversity improvement of PD selection diversity reception, assuming binary DPSK modulation with differential detection. Section 2 describes the system model used in the analysis. Section 3 theoretically evaluates the bit error rate (BER) of the PD diversity reception in the presence of NOP's in the time window. Section 4 describes the construction of an equivalent baseband laboratory experimental system, and presents experimental results of BER performance evaluations.

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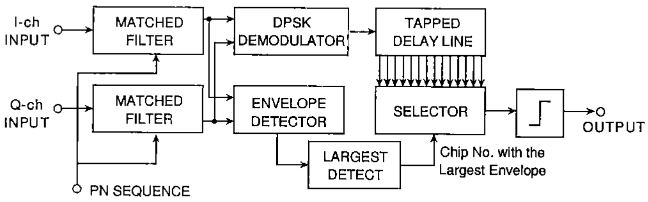

Fig. 1 Model of PD selection diversity reception.

2. System Model

Figure 1 shows a model of the PD selection diversity scheme being analyzed. Binary DPSK (BDPSK) is assumed. The PD selection diversity receiver has a transversal type correlator matched to the reference user's spreading sequence. The in-phase and quadrature-phase channel (I-ch and Q-ch) components of the correlator output are brought to the BDPSK demodulator where the phase difference between the two (I, Q) pairs of the correlator output samples separated in time by the information bit duration is detected each chip time.

The BDPSK demodulator output samples are then fed to a tapped-delay line whose unit delay time is chosen to be equal to the spreading chip duration. The correlation peak position is determined from the correlator output signal power (I^2+Q^2). The time window for PD selection diversity is positioned on the tapped-delay line so that the desired signal components of the demodulated output signals are included, and shifted gradually to track the desired signals.

The chip duration is assumed to be small enough to resolve each propagation path. If the time window size is sufficiently large compared to the channel delay spread, the desired signal components corresponding to each propagation path can be combined. However, too wide a time window allows NOP's.

3. Performance Analysis

3.1 BER Expression

If the k-th chip in the time window has a desired signal component corresponding to a propagation path, the complex representation $z_k(t)$ of the despread received signal, with a unity transmission power, is expressed as

$$z_k(t)=\sqrt{2}z_s(t)e^{j\varphi(t)}+z_n(t), \tag{1}$$

where $z_s(t)$ is the fading complex envelope of the desired signal component, $z_n(t)$ is the additive white Gaussian noise (AWGN), $\varphi(t)$ is the transmitted signal phase, and $j=\sqrt{-1}$. The BDPSK demodulator outputs $\mathrm{Im}\{z_k(t_s)z_k^*(t_s-T)\}$ where Im{ } denotes the imaginary part of the complex variable and t_s is the sampling instant. The phase difference $\varphi(t_s)-\varphi(t_s-T)=0$ or π if the received desired signal suffers from neither noise nor fading. It is apparent that with $z_k(t_s)$ given, $z_k^*(t_s-T)$ becomes a complex Gaussian variable with a mean $\mu^*z_k(t_s)$ and a variance $\sigma^2(1-|\mu|^2)$, where $\sigma^2=\langle|z_k(t_s)|^2\rangle/2$ and $\mu=\mu_{re}+j\mu_{im}=\langle z_k(t_s)z_k^*(t_s-T)\rangle/2\sigma^2$.[7]

Assuming that there are m propagation paths having equal path gains, each of the m equivalent envelopes $|z_k(t_s)|$ distributes over the Rayleigh probability density function $(pdf)\,p(|z_k(t_s)|)$ of

$$p(|z_k(t_s)|)=\frac{1}{\sigma^2}|z_k(t_s)|\cdot\exp\left\{-\frac{|z_k(t_s)|^2}{2\sigma^2}\right\}, \tag{2}$$

where

$$\sigma^2=\sigma_s^2+\sigma_n^2, \tag{3}$$

and σ_s^2 and σ_n^2 are the variances of the complex fading envelope for the desired signal and the AWGN, respectively. If the k-th chip is the NOP,

$$z_k(t)=z_n(t), \tag{4}$$

and also each of the $M-m$ equivalent envelopes $|z_k(t_s)|$ has the Rayleigh pdf of Eq. (2), where σ^2 is replaced by σ_n^2.

The PD selection diversity receiver selects the demodulator output signal having the largest signal envelope R from among the M chips in the time window as

$$R=\operatorname*{Max}_k[|z_k(t_s)|]. \tag{5}$$

Hence, the *pdf* $p(R_s)$ of the largest equivalent envelope of the $m|z_k(t_s)|$'s having the desired signal components, R_s, is expressed as

$$p(R_s)=\frac{m}{\sigma^2}R_s\exp\left(-\frac{R_s^2}{2\sigma^2}\right)\left\{1-\exp\left(-\frac{R_s^2}{2\sigma^2}\right)\right\}^{m-1}. \tag{6}$$

On the other hand, the *pdf* $p(R_n)$ of the largest equivalent envelope of the $M-m|z_k(t_s)|$'s having the NOP components, R_n, becomes

$$p(R_n)=\frac{M-m}{\sigma_n^2}R_n\exp\left(-\frac{R_n^2}{2\sigma_n^2}\right)\cdot\left\{1-\exp\left(-\frac{R_n^2}{2\sigma_n^2}\right)\right\}^{M-m-1}. \tag{7}$$

Therefore, if $R=R_s=\mathrm{Max}(R_s, R_n)$, the selected demodulator output has the desired signal component. In this case, for the transmitted binary data of 1, the bit error probability $p_e(R_s)$ conditioned on the signal envelope R_s is given by

$$p_e(R_s)=\frac{1}{2}\mathrm{erfc}\left(\frac{\mu_{im}}{\sqrt{1-|\mu|^2}}\cdot\frac{R_s}{\sqrt{2}\sigma}\right). \tag{8}$$

If $R=R_n=\mathrm{Max}(R_s, R_n)$, the selected path is the NOP, and a bit error results with a probability of 0.5, Therefore, the average BER P_b can be calculated using Eqs. (6), (7) and (8) as

$$P_b=\int_0^{\infty}\int_0^{R_s} p_e(R_s)\,p(R_s)\,p(R_n)\,dR_n dR_s + \int_0^{\infty}\int_0^{R_n} \frac{1}{2} p(R_n)\,p(R_s)\,dR_s dR_n \qquad (9)$$

for $M>m$.

After some manipulations, we have

$$P_b=\frac{m(M-m)}{2}\sum_{i=0}^{M-m-1}\sum_{k=0}^{m-1}(-1)^{i+k} \cdot\binom{M-m-1}{i}\binom{m-1}{k} \cdot\frac{1}{i+1}\left[\frac{1}{k+1}\left\{2-\sqrt{\frac{\mu_{im}^2}{\mu_{im}^2+(k+1)(1-|\mu|^2)}} -\frac{(i+1)(1+\Gamma)}{i+k+2+i\Gamma+\Gamma}\right\} -\frac{1}{i+k+2+i\Gamma+\Gamma} \cdot\left\{1-\sqrt{\frac{\mu_{im}^2}{\mu_{im}^2+(1-|\mu|^2)(i+k+2+i\Gamma+\Gamma)}}\right\}\right],$$

$$M>m \qquad (10)$$

where $\Gamma=\sigma_s^2/\sigma_n^2$ is the average received signal energy per bit-to-noise power spectral density ratio (E_b/N_0) on each path, and $\binom{A}{B}$ is the binomial coefficient (see Appendix).

If $M=m$, the average BER is equivalent to that of the m-branch selection diversity, and is, for BDPSK, given by[7]

$$P_b=\int_0^{\infty} p_e(R_s)\,p(R_s)\,dR_s =\frac{1}{2}\left\{1-\sum_{k=1}^{m}(-1)^k\binom{m}{k}\sqrt{\frac{\mu_{im}^2}{\mu_{im}^2+k(1-|\mu|^2)}}\right\}. \qquad (11)$$

3.2 Numerical Calculations

The average BER was numerically calculated using Eq. (10). Assuming that fading is very slow compared to the bit rate, i.e., $f_D T\to 0$ where f_D is the maximum Doppler frequency, $\mu=j\Gamma/(\Gamma+1)$. Figure 2 shows the calculated BER versus the average received $E_b/N_0(=m\Gamma)$ with the number of the propagation paths m and the window size M as parameters, where for any value of m, the $m\Gamma$ value has been kept constant since the transmitted signal power is divided into m paths. It is found from Fig. 2 that for $m=2$, the BER performance with $M=6$ is about 2 dB worse than that with $M=2$. For $m=1$, the BER with $M=6$ is about 4 dB worse than that with $M=1$. Figure 3 shows the average BER for $m\Gamma=17$ dB versus the number $M-m$ of the NOP's with m as a parameter. It is found that for small values of $M-m$, the BER increases rapidly as $M-m$ increases. However, for large values of $M-m$, the BER becomes relatively less sensitive to increases in the number $M-m$ of the NOP's.

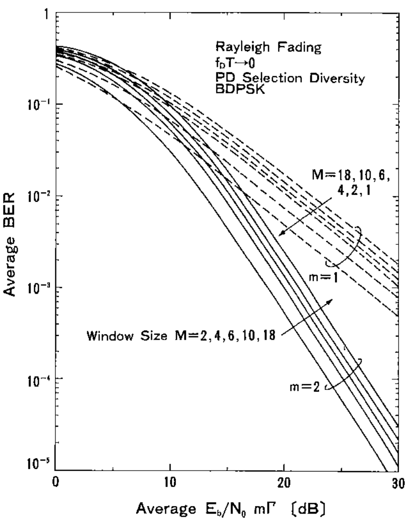

Fig. 2 Average BER vs. average received E_d/N_0.

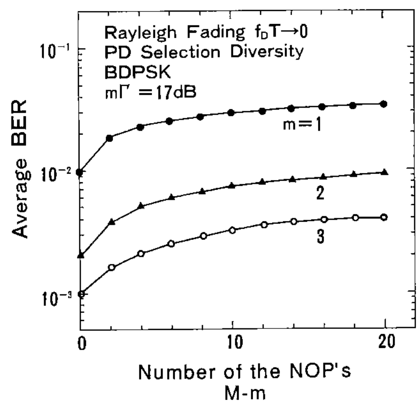

Fig. 3 Average BER vs. the number of the NOP's.

4. Experiments

A major factor that dominates signal transmission performance in DS/SS multi-user communications systems is interference because the same frequency band is shared by many transmitter-receiver pairs. This makes it very difficult to estimate the signal transmission performance of a DS/SS system in a multi-user environment because unless a lot of real transmitter-receiver pairs are field or laboratory tested, the actual performance can not be evaluated.

In order to reduce this difficulty, and to improve the accuracy of performance evaluations through laboratory experiments, an equivalent baseband experiment system was built. The system simulates, in real time, broadband fading signal transmission environments. Also, a prototype of the PD selection receiver was constructed. This section introduces the experimental system, and presents measured BER performance of the PD selection diversity reception system.

4. 1 Channel Simulator

Figure 4 shows the block diagram of the equivalent baseband fading simulator. The double-spike model was employed as the channel delay profile, where the two propagation paths were assumed to have equal gain. The input I-ch and Q-ch components were sampled at a sampling interval of 1/4.096 MHz asynchronous to the input chip timing. The input sample sequence was replicated to simulate the double-spike propagation environment.

Two I-ch and Q-ch pairs of the complex fading envelope were generated based on the model described in Ref. (8). The input sample sequence was multiplied by one of these two pairs in the complex domain. The delayed version of the input sample sequence was multiplied by the other pair in the complex domain. These were then summed, and the resulting I-ch and Q-ch samples are the output of the fading simulator. The maximum Doppler frequency could be set independently of the input chip rate within the range 1 Hz to 400 Hz. The adjustable range of the delay spread was from 0 to 31 μs.

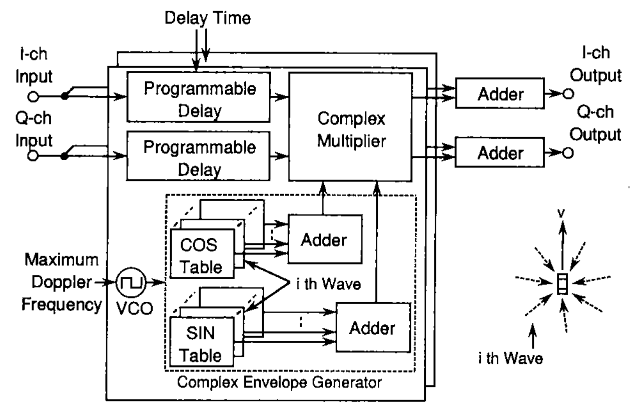

Fig. 4 Block diagram of equivalent baseband frequency selective fading simulator.

Figure 5 shows the block diagram of the laboratory experimental system. One desired and two interference complex signals, each of which suffers from independent frequency selective Rayleigh fading generated by the baseband fading simulator, and two dimensional (I-ch and Q-ch) AWGN are combined in the complex domain by two 4-input full adders. The resulting signals are then brought to the prototype PD selection diversity receiver.

4. 2 Receiver Prototype

In the prototype system, the chip rate was 1.024 Mb/s, and the symbol rate before spreading was 8.063 (=1024/127) kb/s. Gold codes of length 127 chips were used for the spreading sequence (process gain is 21 dB). Two matched filter IC's, Plessy PDSP 16256, were used as I-ch and Q-ch despreaders. Another Plessy IC, PDSP16116, was used to calculate I^2+Q^2. The peak position detector was constructed with random logic circuitry. The I-ch and Q-ch outputs and detected peak position were then input to a Motorola DSP96000, which performed PD selection diversity reception with a window size of 6. Moreover, time window positioning was performed on the DSP using a conventional DPLL algorithm.

4. 3 BER Performance Evaluations

Laboratory experiments were conducted to evaluate the BER performance of PD selection diversity reception using the equivalent baseband laboratory experiment system. The spreading sequences were randomly selected from among the Gold codes and assigned to the interference signals so that the effect of cross correlation between the desired and interference

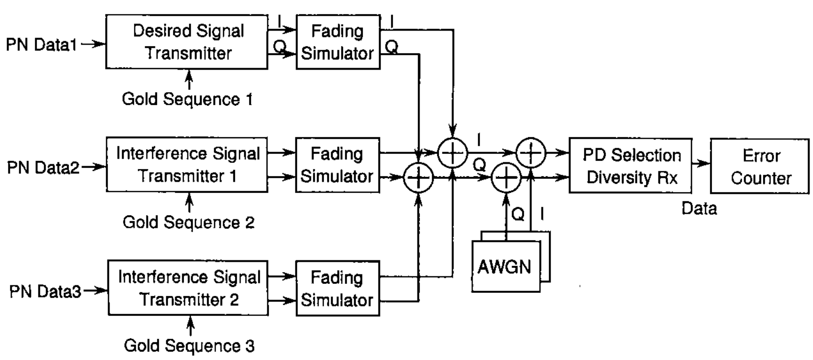

Fig. 5 Block diagram of laboratory experimental system.

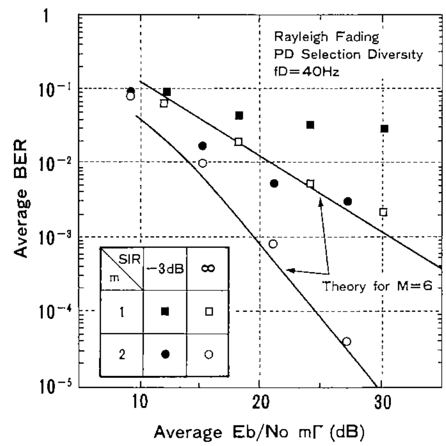

Fig. 6 Experimental average BER vs. average received E_b/N_0.

spreading sequences could be averaged over the sequences of the Gold codes.

Figure 6 shows for $f_D=40$ Hz the average BER of the PD selection diversity reception for infinite and -3 dB signal-to-interference power ratio (SIR) values versus the average received $E_b/N_0(=m\Gamma)$ after despreading with the number of the propagation paths m as a parameter. The path gain for $m=2$ was set at one half of that for $m=1$. The theoretical BER's calculated from Eq. (10) for the infinite SIR are also plotted for $m=1$ and $m=2$ with the time window size M of 6. It is found that the experiment results agree well with the theoretical BER for $M=6$. When the average SIR is -3 dB, the average BER decreases as the average E_b/N_0 increases; however, the BER values reach a floor determined by the -3 dB average SIR.

5. Conclusions

The path diversity improvement inherent in direct sequence spread spectrum (DS/SS) signaling under multipath propagation environments has been investigated for mobile/personal radio communications systems that employ DPSK modulation. The BER performance of post-demodulation (PD) selection diversity reception was analyzed in the presence of noise-only-paths (NOP's) in the time window for diversity combining. It has been shown that for a double-spike delay profile ($m=1$), the BER performance with a window size $M=6$ is about 2 dB worse than that with $M=2$. For a single path propagation environment ($m=1$), the BER with $M=6$ is about 4 dB worse than that with $M=1$.

A prototype of the PD selection diversity receiver was constructed together with an equivalent baseband simulation system. This allows real time simulations of broadband fading signal transmission environments. Results of the laboratory experiments conducted to evaluate the BER performance were also presented. It has been shown that the experimental results agree well with the theoretical BER.

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References

(1) Turin, G. L., "Introduction to Spread-Spectrum Antimultipath Techniques and Their Application to Urban Digital Radio," *Proc. of IEEE*, vol. 68, no. 3, pp. 328-353, Mar. 1980.

(2) Kavehrad, M. and Ramamurthi, B., "Direct-Sequence Spread Spectrum with DPSK Modulation and Diversity for Indoor Wireless Communications," *IEEE Trans. Commun.*, vol. COM-35, no. 2, pp. 224-236, Feb. 1987.

(3) Grob, U., Welti, A. L., Zollinger, E., Kung R. and Kaufmann, H., "Microcellular Direct-Sequence Spread-Spectrum Radio System Using N-Path RAKE Receiver," *IEEE JSAC*, vol. SAC-8, no. 5, pp. 772-780, Jun. 1990.

(4) Proakis, J. G., *Digital Communications*, 2nd Ed., McGraw-Hill, pp. 728-739, 1989.

(5) Chyi, G. T., Proakis J. G. and Keller, C. M., "Diversity Selection/Combining Schemes with Excess Noise-Only Diversity Receptions over A Rayleigh-Fading Multipath Channel," *Proc. Conf. Inform. Sciences Syst.*, Princeton University, Princeton, N. J. pp. 420-425, Mar. 1988.

(6) Chang, I. K., Stuber, G. L. and Bush, A. M., "Performance of Diversity Combining Techniques for DS/DPSK Signaling Over a Pulse Jammed Multipath-Fading Channel," *IEEE Trans. Commun.*, vol. COM-38, no. 10, pp. 1823-1834, Oct. 1990.

(7) Adachi, F. and Parsons, J. D., "Error Rate Performance of Digital FM Mobile Radio with Postdetection Diversity," *IEEE Trans. Commun.*, vol. COM-37, no. 3, pp. 200-210, Mar. 1989.

(8) Jakes, Jr, W. C., *Microwave Mobile Communications*, John Wiley & Sons, pp. 70-76, 1974.

Appendix: Derivation of Eq. (10)

Let I_1 and I_2 be

$$I_1=\int_0^\infty\int_0^{R_s} p_e(R_s)\,p(R_s)\,p(R_n)\,dR_n dR_s, \qquad (\mathrm{A}\cdot 1)$$

and

$$I_2=\int_0^\infty\int_0^{R_n} \frac{1}{2}p(R_n)\,p(R_s)\,dR_s dR_n, \qquad (\mathrm{A}\cdot 2)$$

respectively, with $P_b=I_1+I_2$. We first calculate I_1. Substituting the binomial expansion forms of Eqs. (6) and (7) into Eq. (A·1) yields

$$\begin{aligned}I_1=&\frac{m(M-m)}{2}\sum_{i=0}^{M-m-1}\sum_{k=0}^{m-1}\frac{(-1)^{i+k}}{i+1}\\ &\cdot\binom{M-m-1}{i}\binom{m-1}{k}\\ &\cdot\Bigg[\int_0^\infty dR_s\ \mathrm{erfc}\left\{\frac{\mu_{im}}{\sqrt{1-|\mu|^2}}\cdot\frac{R_s}{\sqrt{2}\sigma}\right\}\cdot\frac{R_s^2}{\sigma^2}\\ &\cdot\exp\left\{\frac{-(k+1)R_s^2}{2\sigma^2}\right\}\\ &-\int_0^\infty dR_s\ \mathrm{erfc}\left\{\frac{\mu_{im}}{\sqrt{1-|\mu|^2}}\cdot\frac{R_s}{\sqrt{2}\sigma}\right\}\cdot\frac{R_s^2}{\sigma^2}\\ &\cdot\exp\left\{-R_s^2\left(\frac{i+1}{2\sigma_n^2}+\frac{k+1}{2\sigma^2}\right)\right\}\Bigg]. \qquad (\mathrm{A}\cdot 3)\end{aligned}$$

Using the formula of

$$\int_0^\infty \mathrm{erfc}(ax)\cdot x\cdot\exp(-bx)\,dx=\frac{1}{2b}\left(1-\frac{a}{\sqrt{a^2+b}}\right), \qquad (\mathrm{A}\cdot 4)$$

we have

$$I_1=\frac{m(M-m)}{2}\sum_{i=0}^{M-m-1}\sum_{k=0}^{m-1}\frac{(-1)^{i+k}}{i+1}\cdot\binom{M-m-1}{i}\binom{m-1}{k}\cdot\left[\frac{1}{k+1}\left\{1-\sqrt{\frac{\mu_{im}^2}{\mu_{im}^2+(k+1)(1-|\mu|^2)}}\right\}-\frac{1}{i+k+2+i\Gamma+\Gamma}\cdot\left\{1-\sqrt{\frac{\mu_{im}^2}{\mu_{im}^2+(1-|\mu|^2)(i+k+2+i\Gamma+\Gamma)}}\right\}\right], \tag{A·5}$$

where $\Gamma=\sigma_s^2/\sigma_n^2$.

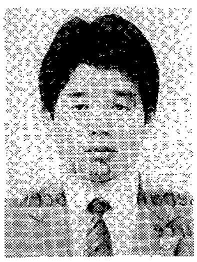
Calculation of I_2 is quite simple. Substituting the binomial expansion form of Eq. (6) into Eq. (A·2) yields

$$I_2=\frac{m(M-m)}{2\sigma^2}\sum_{k=0}^{m-1}\sum_{i=0}^{M-m-1}\frac{(-1)^{i+k}}{k+1}\cdot\binom{M-m-1}{i}\binom{m-1}{k}\cdot\int_0^\infty dR_n\left[R_n\cdot\exp\left\{-\frac{(i+1)R_n^2}{2\sigma_n^2}\right\}\cdot\left(1-\exp\left\{-\frac{(k+1)R_n^2}{2\sigma^2}\right\}\right)\right]. \tag{A·6}$$

After some manipulations, I_2 becomes

$$I_2=\frac{m(M-m)}{2}\sum_{i=0}^{M-m-1}\sum_{k=0}^{m-1}\frac{(-1)^{i+k}}{k+1}\cdot\binom{M-m-1}{i}\binom{m-1}{k}\left(\frac{1}{i+1}-\frac{1+\Gamma}{i+k+2+i\Gamma+\Gamma}\right). \tag{A·7}$$

Using Eqs. (A·5) and (A·7), we have the expression for P_b of Eq. (10).



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