안테나 배열을 사용한 DS-SS 시스템을 위한 직렬 포착 방식과 페이딩 채널에서의 성능

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A Serial Acquisition Scheme for DS-SS Systems Using Antenna Arrays and Its Performance in a Fading Channel

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요 약

안테나 배열을 이용하는 직접 대역 확산 시스템에서 포착 가능한 SNR의 범위를 크게 낮출 수 있는 안테나 배열을 사용한 직렬 포착 방식을 제안한다. 제안된 포착 방식에서는 SNR 향상을 위하여, 안테나 요소마다 동일한 PN 위상을 갖도록 조정된 비동기 I-Q 정합 여파기들에서의 독립된 판정 변수들의 합을 초기 포착을 위한 판정 변수로 사용한다. 제안된 방식의 성능 분석을 위하여 가우시안 전송로와 Rayleigh 페이딩 전송로에서 검출확률과 오인확률이 유도되며, 이를 이용하여 평균 포착시간이 분석된다. 수치적 분석 결과를 통하여, 안테나 개수가 늘어남에 따라 제안된 방식의 포착 성능이 계속적으로 향상됨을 확인할 수 있다.

ABSTRACT

We propose a serial acquisition scheme using antenna arrays for initial acquisition of direct sequence spread spectrum (DS-SS) signals, which can lower substantially the range of detectable signal-to-noise ratio (SNR). The proposed scheme uses the sum of the independent decision samples from pseudo-noise (PN) co-phased noncoherent I-Q matched filters (MFs) associated with antenna arrays as a decision variable in order to enhance SNR of the resulting signal. We analyze its mean acquisition time performance under an additive white Gaussian noise (AWGN) channel and a flat Rayleigh fading channel by deriving the expressions for the probabilities of detection and false alarm. From numerical results, we see that the acquisition performance of the proposed scheme becomes improved continually as the number of antennas increases.

I. Introduction

DS-SS techniques have been used widely for a variety of reasons including anti-interference and multiple access capabilities [1]-[4]. However, these advantages can be fully exploited only if precise receiver synchronization is acquired. Synchronization is done in two steps of acquisition and tracking. Acquisition aligns the phases of the local PN code and the incoming one within one chip interval or less [5]-[11]. Tracking brings the phase difference of two sequences to zero and maintains the alignment [11], [12]. Especially, acquisition is the first process to be performed in the receiver and has a great influence on the overall system performance. Among various acquisition schemes, serial schemes [5]-[9] have a major importance

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since these are simple and provide a good performance in a low SNR environment. In this paper, the focus is mainly on serial acquisition of DS-SS signals.

In wireless communications, the use of smart antennas (possibly antenna arrays) is widely recognized as a promising means to increase the system capacity, to improve the signal quality and to extend the coverage [13]-[21]. Several types of smart antenna systems have been proposed and their performance has been analyzed [15]-[18]. Also, smart antenna techniques for DS-CDMA systems have been studied, and the performance of antenna arrays in the DS-CDMA systems has been analyzed [18]-[22]. However, all such systems based on the DS-CDMA techniques have assumed perfect code synchronization in the receiver. The improvement of initial synchronization using antenna arrays has rarely been considered.

In this paper, we propose a serial acquisition scheme using antenna arrays for initial synchronization of DS-SS signals, which can lower substantially the range of detectable SNR. The proposed scheme uses the sum of the independent decision samples from PN co-phased noncoherent I-Q MFs associated with antenna arrays as a decision variable in order to enhance SNR of the resulting signal. The remaining acquisition processes are the same as those of the conventional serial scheme with a single antenna of [6] (hereafter, called as the conventional scheme or the single antenna scheme). Under a frequency-selective fading channel, the acquisition performance may be degraded significantly [23],[24]. To overcome this performance degradation, different system structures, such as a RAKE receiver and an adaptive equalizer [25] can be used. The objective of this paper is to analyze the performance potential of the serial acquisition scheme using antenna arrays according to the number of antenna elements. Hence, the analyses under the AWGN channel and the flat Rayleigh fading channel are enough to evaluate how well the proposed system performs.

Ⅱ. Serial Acquisition Scheme Using Antenna Arrays

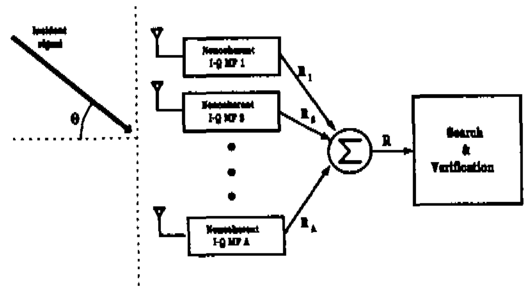


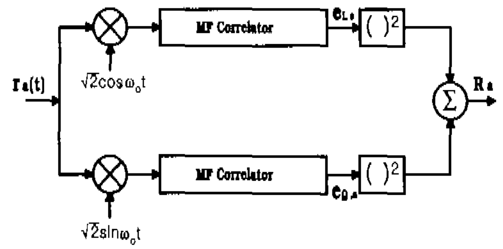
그림 1. 제안된 방식의 시스템 블록도

Fig.1 shows the basic structure of the proposed serial scheme using antenna arrays. Assuming A antennas, the proposed scheme uses the sum of A simultaneous independent decision samples from PN co-phased noncoherent I-Q MFs associated with antenna arrays as a decision variable. The PN code weight registers with length M over all the noncoherent I-Q MF blocks store an identical segment of a given PN code. Fig. 2 shows the basic structure of an individual noncoherent I-Q MF block. In the noncoherent I-Q MF block at the a-th antenna, the outputs of I-branch and Q-branch MF correlators are squared and summed to make the decision sample for noncoherent detection as follow:

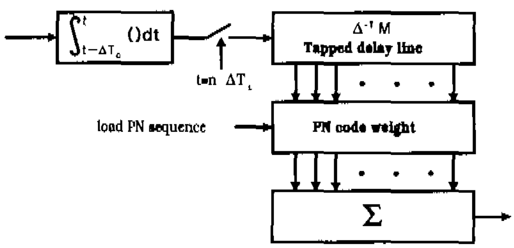
$$R_a = e_{I,a}^2 + e_{Q,a}^2, \tag{1}$$

where $e_{I,a}$ and $e_{Q,a}$ denote the outputs of I-branch and Q-branch MF correlators, respectively, at the a-th antenna. The decision samples based on the PN correlation over MT_c seconds are generated at every ΔT_c seconds ($0<\Delta<1$), that is, at a multiple of the code rate $1/T_c$, with Δ being a phase adjusting factor. Then, the simultaneous A independent decision samples from antenna arrays are summed to make the summed decision sample as follow (See Fig. 1.):

$$R = R_1 + R_2 + \cdots + R_A = e_{I,1}^2 + e_{Q,1}^2 + e_{I,2}^2 + e_{Q,2}^2 + \cdots + e_{I,A}^2 + e_{Q,A}^2 \tag{2}$$



(a) 비동기 I-Q 정합 여과기



(b) 정합 여과 상관기.

그림 2. a번째 안테나에서의 비동기 I-Q 정합 여과기 구조.

Hence, the proposed scheme generates a summed decision sample at every ΔT_c seconds.

The proposed scheme using antenna arrays has two acquisition modes: the search mode and the consecutive verification mode. The search mode makes a tentative decision on the correct phase based on the summed decision sample through the successive search processes over the uncertainty region. Then the verification mode makes a final decision on the selected phase through the coincident detection (CD) to provide the good protection against a very costly false-lock-loop activation. Note that the acquisition processes in the proposed scheme are the same as those of the *single antenna scheme* [6], except for using the summed decision sample. In the search mode, if the summed decision sample exceeds a search mode threshold γ_S, the system decides a tentative correct phase and then goes to the verification mode. In the verification mode, C summed decision samples corresponding to the tentative correct phase over the C consecutive non-overlapping MT_c intervals are collected. If D or more of C summed decision samples exceed a verification mode threshold γ_V, the correct acquisition is decided. Otherwise, the acquisition process is continued at the next phase. In the false alarm case, the system goes back to the search mode after a penalty time, JMT_c.

Ⅲ. Mean Acquisition Time

In this section, we analyze the mean acquisition time performance of the proposed scheme under the following assumptions [5]-[6], [10]:

(A1) There is only one cell corresponding to the correct phase (H_1 cell).

(A2) All summed decision samples are independent.

(A3) M is large enough such that the correlation of the received sequence and the local code yields zero when they are not in phase (H_0 cells).

(A4) The uncertainty region is the full code length LT_c.

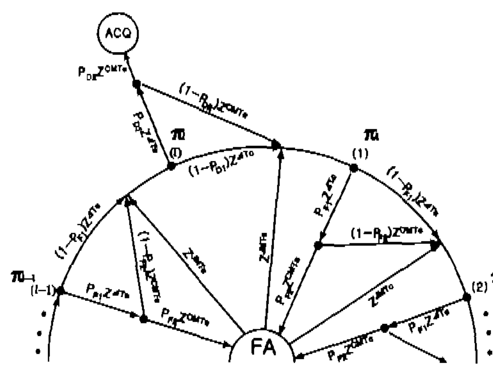


그림 3. 제안된 방식의 상태 천이도.

Due to the Markovian nature of the acquisition process, we derive the mean acquisition time expression by using the state transition diagram. Fig. 3 shows the state transition diagram used to analyze the acquisition performance of the proposed serial scheme. (The similar state transition diagram was originally used in the single antenna scheme [6]). The state diagram consists of the correct acquisition (ACQ) state, the false alarm (FA) states, the detection state (H_1 cell), and (l-1) offset states (H_0 cells) with l=L

Δ^{-1}. An initial state is decided with *a priori distribution*, π_i, $i=1,2,\cdots,l$ Branch gains are given by [5]-[6], [10]

$$\begin{aligned} H_P(Z) &= Z^{JMT_c}, \\ H_D(Z) &= P_{D1}Z^{\Delta T_c}P_{D2}Z^{CMT_c}, \\ H_M(Z) &= (1-P_{D1})Z^{\Delta T_c}+P_{D1}Z^{\Delta T_c}(1-P_{D2})Z^{CMT_c}, \quad (3) \\ H_0(Z) &= (1-P_{F1})Z^{\Delta T_c}+P_{F1}Z^{\Delta T_c}(1-P_{F2})Z^{CMT_c} \\ &\quad +P_{F1}Z^{\Delta T_c}P_{F2}Z^{CMT_c}Z^{JMT_c}, \end{aligned}$$

where $H_P(z)$ is the branch gain passing from the FA states to the next state (i.e., state i+1); $H_D(z)$ is the overall branch gain leading from H_1 cell (the detection state) to the ACQ state; $H_M(z)$ is the overall branch gain connecting H_1 cell to the first offset state (i.e., state 1); $H_0(z)$ is the overall gain connecting any other two successive states (i, i+1), $i=1,2,\cdots,l-1$; P_{D1} and P_{F1} denote the detection and false alarm probabilities in the search mode, respectively; and P_{D2} and P_{F2} denote the detection and false alarm probabilities in the verification mode, respectively.

Then, the generating function [5] is

$$H(Z)=\frac{H_D(Z)}{1-H_M(Z)H_0^{l-1}(Z)}\left\{\sum_{i=1}^{l}\pi_i H_0^{l-i}(Z)\right\} \quad (4)$$

Assuming uniform priori distribution, $\pi_i=1/l$, $i=1,2,\cdots,l$, the mean acquisition time, $E\{T_{acq}\}$ is obtained as [5]

$$\begin{aligned} E\{T_{acq}\} &= \left.\frac{dH(Z)}{dZ}\right|_{Z=1}, \\ &= \frac{T_c}{P_D}\left\{\Delta+CM\,P_{D1}+(\Delta+CM\,P_{F1}+JM\,P_F)(l-1)(1-\frac{P_D}{2})\right\}, \end{aligned} \quad (5)$$

where $P_D=P_{D1}P_{D2}$ and $P_F=P_{F1}P_{F2}$ denote the overall detection probability and the overall false alarm probability, respectively.

Ⅳ. Performance Analysis in an AWGN Channel

In this section, we derive the detection and false alarm probabilities of the proposed serial scheme using antenna arrays in an AWGN channel. From now on, we assume a narrowband signal incident on a uniform linear array with the element spacing of d from only a single direction θ Hence, all the antenna arrays suffer the identical fading, except for the phase rotations due to their array responses. Then, the A1 complex received signal vector can be written as

$$\mathbf{r}(t)=\sum_{i=-\infty}^{\infty}\sqrt{2S}\,c_i(t-iT_c)\,\mathbf{a}(\theta)\,e^{j(\omega_o t+\phi)}+\mathbf{n}(t), \quad (6)$$

where S is the received signal power; $c_i(t)$ is the rectangular pulse over 0 to T_c seconds, which takes the values of 1 or -1 according to the i-th code of the PN sequence; $\mathbf{a}(\theta)$ is the $A\times1$ complex array response vector to the direction θ, defined as

$$\mathbf{a}(\theta)=\begin{bmatrix}1 & e^{-j\omega_o\frac{d}{c}\sin\theta} & \cdots & e^{-j\omega_o\frac{(A-1)d}{c}\sin\theta}\end{bmatrix}^T, \quad (7)$$

with the speed of light c; ω_o and ϕ are the carrier frequency in rad/s and its phase, respectively; and $\mathbf{n}(t)$ is the $A\times1$ zero-mean complex Gaussian noise vector with the one-sided power spectral density matrix of $N_0\mathbf{I}$.

The outputs of the I-branch and Q-branch MF correlators at the a-th antenna are given by, respectively,

$$e_{I,a}=S_{I,a}+N_{I,a}, \quad (8)$$

$$e_{Q,a}=S_{Q,a}+N_{Q,a}, \quad (9)$$

where $S_{I,a}$ and $S_{Q,a}$ are deterministic signal components caused by the transmitted signal at the I- and Q-branches, respectively. Under hypothesis H_1, referring to the assumption (A1), Eq. (6) and Fig. 2(a), we obtain

$$S_{I,a} \mid H_1 = \sqrt{S} MT_c \cos\left(\phi - \omega_o \frac{(a-1)d}{c} \sin\theta\right), \quad (10)$$

$$S_{Q,a} \mid H_1 = \sqrt{S} MT_c \sin\left(\phi - \omega_o \frac{(a-1)d}{c} \sin\theta\right), \quad (11)$$

while under hypotheses H_0, referring to the assumption (A3), we obtain

$$S_{I,a} \,\&\, S_{Q,a} \mid H_0 = 0 \quad (12)$$

$N_{I,a}$ and $N_{Q,a}$ are independent and identically distributed (i.i.d.) zero-mean Gaussian random variables (r.v.'s) at the I- and Q-branches, respectively, which are obtained by the PN correlation of the complex AWGN at the a-th antenna, and their variance σ_f^2 is given by

$$\sigma_f^2 = \frac{N_0 MT_c}{2} \quad (13)$$

Therefore, $e_{I,a}$ and $e_{Q,a}$ follow the Gaussian distribution

$$e_{I,a} \mid H_1 = G\left(S_{I,a}, \sigma_f^2\right), \quad (14)$$

$$e_{Q,a} \mid H_1 = G(S_{Q,a}, \sigma_f^2), \quad (15)$$

$$e_{I,a} \,\&\, e_{Q,a} \mid H_0 = G(0, \sigma_f^2) \quad (16)$$

From the above results, the decision sample R_a follows the noncentral chi-square distribution with 2 degrees of freedom under H_1 cell, and the central chi-square distribution with 2 degrees of freedom under H_0 cells. Under H_1 cell, the noncentrality parameter is given by

$$m_a^2 = S_{I,a}^2 + S_{Q,a}^2 = S(MT_c)^2 \quad (17)$$

In addition, from (1) and (8) to (12), we can note that the decision samples from all the antenna arrays have independent and identical distributions without respect of the incident directions and their phases.

As described above (See Fig. 1 and Eq. (2).), the proposed serial scheme uses the sum of A simultaneous independent decision samples from the PN co-phased noncoherent I-Q MFs associated with antenna arrays. In addition, note that $\{N_{I,a}\}$ and $\{N_{Q,a}\}$, a=1, 2, , A, are i.i.d. zero-mean Gaussian r.v.'s with the variance σ_f^2. Then, the summed decision sample R follows the noncentral chi-square distribution with $2A$ degrees of freedom under H_1 cell, and the central chi-square distribution with $2A$ degrees of freedom under H_0 cells [26]. These probability density functions (pdfs) are given by

$$P_R(h|H_1) = \frac{1}{2\sigma_f^2}\left(\frac{h}{K^2}\right)^{\frac{A-1}{2}} \exp\left(-\frac{K^2+h}{2\sigma_f^2}\right) I_{A-1}\left(\frac{\sqrt{hK^2}}{\sigma_f^2}\right), \quad (18)$$

$$P_R(g|H_0) = \frac{1}{\sigma_f^{2A} 2^A \Gamma(A)} g^{A-1} \exp\left(-\frac{g}{2\sigma_f^2}\right), \quad (19)$$

where $I_i(\cdot)$ denotes the modified Bessel function of the first kind and i-th order, $\Gamma(\cdot)$ denotes the Gamma function, and K^2 is the noncentrality parameter for the summed decision sample under H_1 cell, defined as

$$K^2 = \sum_{a=1}^{A} m_a{}^2 = AS(MT_c)^2 \quad (20)$$

Now, we derive the detection and false alarm probabilities for the search mode. The detection and false alarm probabilities in the search mode, P_{D1} and P_{F1} are defined as the probabilities that the summed decision sample R exceeds the search mode threshold γ_s, under H_1 and H_0 cells, respectively. These are

$$P_{D1} = \int_{\gamma_s}^{\infty} P_R(h|H_1)\,dh, \quad (21)$$

$$P_{F1} = \int_{\gamma_s}^{\infty} P_R(g|H_0)\,dg \quad (22)$$

Substituting (18) into (21), and (19) into (22), after some manipulations, these are simplified as

$$P_{D1} = Q(a,b) + \exp\left(-\frac{a^2+b^2}{2}\right)\sum_{k=1}^{A-1}\left(\frac{b}{a}\right)^k I_k(ab), \tag{23}$$

$$P_{F1} = \frac{1}{\Gamma(A)}\exp\left(-\frac{b^2}{2}\right)\sum_{k=0}^{A-1}\left\{\frac{(A-1)!}{2^k \cdot k!}b^k\right\}, \tag{24}$$

where, $a^2 = 2S(MT_c)^2 A/\sigma_f^2$, $b^2 = \gamma_S/\sigma_f^2$ and $Q(\cdot,\cdot)$ denotes the Marcum's Q-function [26].

In the verification mode, if D or more of C summed decision samples corresponding to the tentative correct phase over consecutive CMT_c seconds exceed the verification mode threshold γ_V, the correct acquisition is decided. Hence, the detection and false alarm probabilities for the verification mode are given by, respectively,

$$P_{D2} = \sum_{n=D}^{C}\binom{C}{n}P_1^n(1-P_1)^{C-n}, \tag{25}$$

$$P_{F2} = \sum_{n=D}^{C}\binom{C}{n}P_2^n(1-P_2)^{C-n}, \tag{26}$$

where P_1 and P_2 are defined as follows:

$$P_1 = \int_{\gamma_V}^{\infty} P_R(h|H_1)dh, \tag{27}$$

$$P_2 = \int_{\gamma_V}^{\infty} P_R(g|H_0)dg \tag{28}$$

The probabilities P_1 and P_2 are simply obtained by replacing γ_S by γ_V from (23) and (24), respectively.

Finally, to obtain the mean acquisition time of the proposed scheme in terms of SNR/chip in the AWGN channel, the above probabilities, P_{D1}, P_{F1}, P_{D2} and P_{F2} must be substituted into (5).

V. Performance Analysis in a Rayleigh Fading Channel

In this section, we analyze the mean acquisition time performance of the proposed scheme under the slowly time-varying flat Rayleigh fading channel. For the fading channel considered here, the fading process at each antenna is regarded as a constant over k successive chips, $k \ll M$, and these successive groups of k chips are correlated. Clearly, the smaller the value of k and the correlation coefficients among groups of k chips, the faster the fade rate. The fading processes from one antenna to another are identical except for the relative phase rotations associated with the distances from a reference antenna, assuming the narrowband signal from a single direction as before.

We can write the $A \times 1$ complex received signal vector as

$$\mathbf{r}(t) = \sum_{l=-\infty}^{\infty}\sqrt{2}\left(x_{\lceil l/k \rceil}(t) + jy_{\lceil l/k \rceil}(t)\right)c_l(t-lT_c)\mathbf{a}(\theta)e^{j(\omega_0 t+\phi)} + \mathbf{n}(t), \tag{29}$$

where $\lceil m/n \rceil$ is the largest integer which is below m/n. $x_l(t)$ and $y_l(t)$ are zero-mean Gaussian fading processes at the I- and Q-branches of the reference antenna, respectively, with variances,

$$E[x_l^2(t)] = E[y_l^2(t)] = \sigma_f^2, \tag{30}$$

and autocorrelations

$$E[x_i(t)x_j(t)] = E[y_i(t)y_j(t)] = \rho_{|i-j|}\sigma_f^2, \quad i \neq j, \tag{31}$$

where $1 \geq \rho_1 \geq \rho_2 \geq \cdots \geq 0$ are the autocorrelation coefficients among groups of k chips.

The outputs of the I-branch and Q-branch MF correlators at the a-th antenna are given by, respectively,

$$e_{I,a} = \cos\left\{\phi - \omega_0\overline{(a-1)d}\sin\theta\right\}xT_c + \sin\left\{\phi - \omega_0\overline{(a-1)d}\sin\theta\right\}yT_c + N_{I,a}, \\ = S_{I,a} + N_{I,a}, \tag{32}$$

$$e_{Q,a} = \sin\left\{\phi - \omega_0\overline{(a-1)d}\sin\theta\right\}xT_c - \cos\left\{\phi - \omega_0\overline{(a-1)d}\sin\theta\right\}yT_c + N_{Q,a}, \\ = S_{Q,a} + N_{Q,a}, \tag{33}$$

where x and y are independent zero-mean Gaussian r.v.'s, which are obtained by the PN correlation of the fading processes $x_l(t)$ and $y_l(t)$, respectively. Since x and y are identically distributed, we consider only x.

Under hypothesis H_1, referring to Fig. 2(b), x is the summation of the $l+2$ zero-mean Gaussian r.v.'s given by $(x_1\ x_2\ x_3 \cdots x_l\ x_{l+1}\ x_{l+2})$ with $l = \lceil M/k \rceil$ The r.v.'s x_2, x_3, $\cdots$, x_{l+1} are due to the complete groups of k successive chips with constant fading envelope in the interior of the tapped delay line, and the two remaining r.v.'s, x_1 and x_{l+2} are due to the incomplete groups of k successive chips with constant envelope at the two edges of the tapped delay line. Under hypotheses H_0, since the phases of the received PN sequence and the locally generated PN code are not matched, the MF coefficients are considered approximately independent, ± 1-valued random variables [10]. Hence, x is the summation of M i.i.d. zero-mean Gaussian r.v.'s with the identical variance σ_s^2. After some manipulations, we obtain the variances of x and y under hypotheses H_1 and H_0, respectively, as follows:

$$E\left[x^2 | H_1\right] = E\left[y^2 | H_1\right] = W\sigma_s^2, \tag{34}$$

$$E\left[x^2 | H_0\right] = E\left[y^2 | H_0\right] = M\sigma_s^2, \tag{35}$$

where W is given by

$$W = k^2\left[l + \frac{1}{3}\left(\frac{k'}{k}\right)^2 (2+\rho_{l+1}) + 2\sum_{j=1}^{l}\left(\frac{M}{k} - j\right)\rho_j + \frac{1}{3}\frac{k'}{k^2}(1-\rho_{l+1})\right], \tag{36}$$

with $k' = k\left[M/k - \lceil M/k \rceil \right]$ [10].

From (32) and (33), we note that all the antenna arrays under the assumption of the narrowband signal impinging from a single direction suffer the identical fading except for the phase rotations. Hence, $e_{I,a}$ and $e_{Q,a}$ follow the conditional Gaussian distribution with the conditional means $S_{I,a}$ and $S_{Q,a}$, and the equal variance σ_f^2 as follows:

$$e_{I,a} | H_i = G(S_{I,a}, \sigma_f^2 | x, y), \tag{37}$$

$$e_{Q,a} | H_i = G(S_{Q,a}, \sigma_f^2 | x, y), \quad i = 0, 1. \tag{38}$$

Then, R_a follows the conditional noncentral chi-square distribution with 2 degrees of freedom. The noncentrality parameter is given by

$$m_a^2 = S_{I,a}^2 + S_{Q,a}^2 = (xT_c)^2 + (yT_c)^2 \tag{39}$$

Letting $s = (xT_c)^2 + (yT_c)^2$, the noncentrality parameter follows the following distribution:

$$P_S(s|H_i) = \frac{1}{2\sigma_i^2}\exp\left(-\frac{s}{2\sigma_i^2}\right), \quad i = 0, 1, \tag{40}$$

where $\sigma_0^2 = MT_c^2\sigma_s^2$ and $\sigma_1^2 = WT_c^2\sigma_s^2$.

Again, noting that $\{N_{I,a}\}$ and $\{N_{Q,a}\}$ are i.i.d. zero-mean Gaussian r.v.'s with the variance σ_f^2, the summed decision sample R follows the conditional noncentral chi-square distribution with $2A$ degrees of freedom, and its pdf is given by

$$P_R(h | s) = \frac{1}{2\sigma_f^2}\left(\frac{h}{K^2}\right)^{A-1} \exp\left(-\frac{h+K^2}{2\sigma_f^2}\right) I_{A-1}\left(\frac{\sqrt{hK^2}}{\sigma_f^2}\right), \tag{41}$$

where K^2 is the noncentrality parameter for the summed decision sample given by

$$K^2 = \sum_{a=1}^{A} m_a^2 = A\,s \tag{42}$$

From the probability theory [27], the pdf of R can be written as

$$P_R(h | H_i) = \int_0^\infty P_R(h | s) P_T(s|H_i) ds, \quad i = 0, 1. \tag{43}$$

Substituting (40) and (41) into (43), we get

$$P_R(h | H_i) = \frac{1}{4\sigma_f^2\sigma_i^2}\left(\frac{h}{A}\right)^{\frac{A-1}{2}} \exp\left(-\frac{h}{2\sigma_f^2}\right)$$

$$\times\int_0^{\infty}\left(\frac{1}{s}\right)^{\frac{A-1}{2}} I_{A-1}\left(\frac{\sqrt{hAs}}{\sigma_f^2}\right)\exp\left(-\frac{As}{2\sigma_f^2}-\frac{s}{2\sigma_i^2}\right)ds,$$

$$i=0,1 \qquad (44)$$

After some manipulations, we get a simplified form as

$$P_R(h\mid H_i)=\frac{1}{4\alpha_i\sigma_f^2\sigma_i^2\Gamma(A)}\left(\frac{h}{2\sigma_f^2}\right)^{A-1}\exp\left(-\frac{h}{2\sigma_f^2}\right){}_1F_1\left(1;A;\frac{\beta^2}{\alpha_i}\right),$$

$$i=0,1, \qquad (45)$$

where $\alpha_i=\frac{A}{2\sigma_f^2}+\frac{1}{2A\sigma_i^2}$, $\beta=\frac{\sqrt{Ah}}{2\sigma_f^2}$, and ${}_1F_1(\cdot;\cdot;\cdot)$ is the Kummer confluent hypergeometric function [28].

Now, we derive the expressions for the detection and false alarm probabilities in the search and verification modes. In the search mode, the detection and false alarm probabilities, P_{D1} and P_{F1} are defined as, respectively,

$$P_{D1}=\int_{\gamma_s}^{\infty}P_R(h|H_1)dh, \qquad (46)$$

$$P_{F1}=\int_{\gamma_s}^{\infty}P_R(g|H_0)dg \qquad (47)$$

Substituting (45) into (46) and (47), P_{D1} and P_{F1} are obtained as

$$P_{D1}=\frac{1}{4\alpha_1\sigma_f^2\sigma_1^2\Gamma(A)}\int_{\gamma_s}^{\infty}\left(\frac{h}{2\sigma_f^2}\right)^{A-1}\exp\left(-\frac{h}{2\sigma_f^2}\right){}_1F_1\left(1;A;\frac{\beta^2}{\alpha_1}\right)dh, \qquad (48)$$

$$P_{F1}=\frac{1}{4\alpha_0\sigma_f^2\sigma_0^2\Gamma(A)}\int_{\gamma_s}^{\infty}\left(\frac{g}{2\sigma_f^2}\right)^{A-1}\exp\left(-\frac{g}{2\sigma_f^2}\right){}_1F_1\left(1;A;\frac{\beta^2}{\alpha_0}\right)dg. \qquad (49)$$

In the verification mode, if D or more of C summed decision samples corresponding to the tentative correct phase over CMT_c seconds exceed the verification mode threshold γ_V, the correct acquisition is decided. Hence, the detection and false alarm probabilities are given by, respectively,

$$P_{D2}=\sum_{n=D}^{C}\binom{C}{n}P_1^n(1-P_1)^{C-n}, \qquad (50)$$

$$P_{F2}=\sum_{n=D}^{C}\binom{C}{n}P_2^n(1-P_2)^{C-n}, \qquad (51)$$

where P_1 and P_2 can be simply obtained by replacing γ_s by γ_V from (48) and (49), respectively.

To obtain the mean acquisition time of the proposed scheme in terms of SNR/chip under the flat Rayleigh fading channel, the above probabilities, P_{D1}, P_{F1}, P_{D2} and P_{F2} must be substituted into (5).

Ⅵ. Numerical Results

In this paper, we have analyzed the mean acquisition time performance of the proposed serial scheme using antenna arrays under the AWGN channel and the Rayleigh fading channel. We assumed a PN code with length $L=1023$ and the chip rate $1/T_c=1$ Mchip/s, the MF length $M=64$, the phase adjusting factor $\Delta=1/2$ and the false alarm penalty factor $J=1000$. We also considered 1, 2, 4 and 8 for the number of antennas and used $C=4$ and $D=2$ as the CD parameters for the verification mode [6]. In order to demonstrate the acquisition performance of the proposed scheme under the Rayleigh fading channel, the correlation coefficients ρ_i were taken as ρ^i [10]. We use the normalized mean acquisition time in terms of SNR/chip, which is derived by dividing the mean acquisition time by the uncertainty region LT_c. The thresholds were optimized to minimize the mean acquisition time for each value of SNR/chip.

Fig. 4 shows the mean acquisition time performance for the proposed scheme with A as a parameter under the AWGN channel. We can see that the proposed scheme using antenna arrays can lower substantially the range of detectable SNR than the single antenna scheme, and the

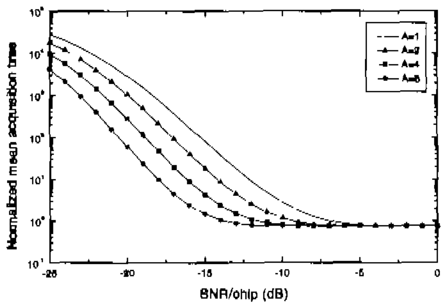


그림 4. 가우시안 전송로에서 안테나 수 (A)를 파라미터로 한 칩당 신호대잡음비에 따른 제안된 방식의 정규화된 평균 포착 시간.

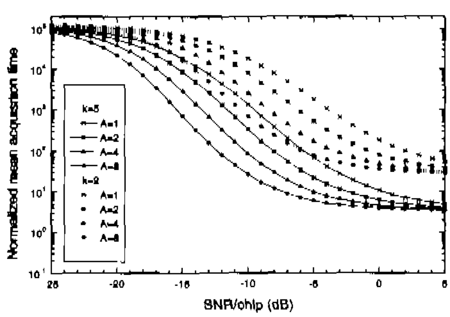


그림 5. 레일리 페이딩 전송로에서 A와 k를 파라미터로 한 칩당 신호대잡음비에 따른 제안된 방식의 정규화된 평균 포착 시간 (ρ =0.3).

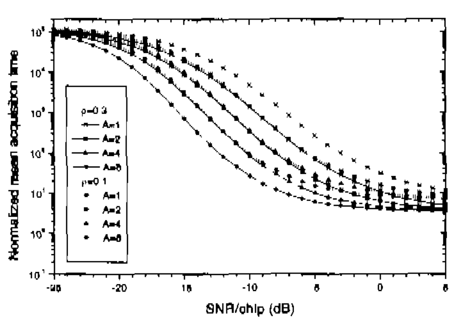


그림 6. 레일리 페이딩 전송로에서 A와 ρ를 파라미터로 한 칩당 신호대잡음비에 따른 제안된 방식의 정규화된 평균 포착 시간 (k=5).

acquisition performance of the proposed scheme becomes improved continually as the number of antennas increases. It is also noted that the range of detectable SNR is lowered by about 2dB every twice number of antennas. The acquisition performance of the proposed scheme under the Rayleigh fading channel and the influence of fade rapidity are shown in Figs. 5 and 6, where the normalized mean acquisition time is depicted as a function of SNR/chip with A and k, and A and ρ, respectively, as parameters. In Fig. 5, the effect of the change in k was examined, in which k = 2 and 5 were considered with $\rho = 0.3$ fixed. In Fig. 6, the effect of the change in ρ was examined, in which $\rho = 0.1$ and 0.3 were considered with $k = 5$ fixed. From Figs. 5 and 6, we see that even under the flat Rayleigh fading channel, detectable SNR can be saved by about 2dB every twice number of antennas. As expected, for faster fading rates (that is, as k and/or ρ become smaller), the mean acquisition time performance becomes worse. In addition, it is noted that the change in k affects the mean acquisition time more sensitively than the change in ρ.

Ⅶ. Conclusions

Antenna arrays are widely being applied in wireless communication systems in order to enhance the system capacity and the quality of service. In this paper, we proposed the serial scheme using antenna arrays for initial acquisition of DS-SS signals. The proposed scheme uses the sum of the simultaneous independent decision samples from PN co-phased noncoherent I-Q MFs associated with antenna arrays as a decision variable in order to enhance SNR of the resulting signal. We analyzed the mean acquisition time performance by deriving the detection and false alarm probabilities for the search and verification modes under the AWGN channel and the Rayleigh fading channel.

From numerical results, we saw that the proposed scheme using antenna arrays can lower substantially the range of detectable SNR than the conventional serial acquisition scheme with a single antenna under the Rayleigh fading channel as well as the AWGN channel. In addition, we saw that detectable SNR can be lowered continually by about 2 dB every twice number of antennas, and this advantage of antenna arrays is maintained even under severe fading channel (k and/or ρ become smaller).

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