A PN Acquisition Scheme using RAKE Receiver and Double Dwell Techniques for DS/SS Communications in Frequency-Selective Rayleigh Fading Channel

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Abstract: Performance of conventional PN acquisition schemes deteriorates when it is used in a fading channel. A new modified PN acquisition scheme is proposed to improve that performance in a frequency-selective fading channel. Simulation is used for verifying the performance which shows that the new scheme outperforms the convention PN acquisition scheme using similar hardware.

1. Introduction

Direct-sequence spread-spectrum (DS/SS) scheme becomes a popular modulation technique today because it has several advantages, such as interference suppression, energy density reduction, fine time resolution and multiple access [1], as well as "soft handoff" in cellular phone systems.

An important component is pseudo-noise (PN) which is used for spreading the bandwidth of the message at the transmitter and for despreading the received signal at the receiver. Despreading is possible only if the local PN signal is synchronized with the received PN signal. One of the major functions of the receiver is to generate a local PN signal which is synchronous with the PN signal in the incoming signal. Traditionally such a synchronization is performed in two steps: acquisition (coarse phase alignment) and tracking (fine tuning). Conventional acquisition schemes perform correlation between the incoming signal and the PN signal [2],[3] and use the correlation result to make a decision for coarse alignment.

There are a lot of reports on PN acquisition but most of them assumed an additive white Gaussian noise (AWGN) channel. If the channel is actually a fading channel, the performance of these schemes would dramatically decrease [4]. To take care of fading, techniques such a post-detector may be used [5] [6].

In cellular systems, the channel may be modeled as Rician and Rayleigh fading channel. In a Rician fading channel, the received signal consists of a direct path and multiple reflective paths, while the received signal of a Rayleigh fading channel consists of multiple reflective paths without a direct path. The characteristics of fading channel have been explained in [7] which describes the fading channel in both time and frequency domains. A suitable channel model for DS/SS systems is frequency-selective Rayleigh fading channel because the bandwidth of the DS/SS system is larger than the coherence bandwidth. Frequency-selective fading causes intersymbol interference (ISI), which can be reduced by using a RAKE receiver [8]. A RAKE receiver uses the benefit of path diversity to increase the signal strength.

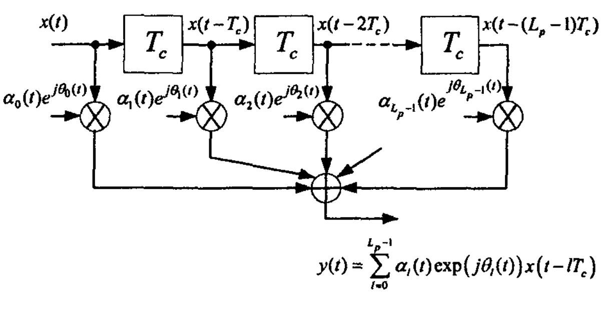

Figure 1. Frequency-selective fading channel model.

The idea of RAKE receiver was applied to the parallel search acquisition scheme in [9]. In this paper we propose a PN acquisition scheme which combines the benefits of RAKE receiver and double dwell technique. The combination improves the performance without additional hardware.

This paper is organized as follows. Section 2 describes the channel model, Section 3 describes the proposed scheme. Section 4 is simulation result and finally Section 5 is conclusion.

2. Channel Model

A frequency-selective fading channel of DS/SS can be modeled as a tapped-delay line with a tap spacing of one chip [10] as shown in Figure 1. The channel impulse response at time t due to an impulse at time ζ is

$$h(t,\zeta)=\sum_{l=0}^{L_p-1}\alpha_l(t)\exp(j\theta_l(t))\delta(t-\zeta-lT_c), \qquad (1)$$

where $\alpha_l(t)$ and $\theta_l(t)$ are tap gain and phase of l-th path at time t, respectively. The tap gains, $\alpha_l(t)$, are assumed to be independent identically distributed (i.i.d.) Rayleigh random variables with a probability density function (pdf) given by

$$f_{\alpha_l}(x)=\frac{2x}{\Omega}\exp\left(-\frac{x^2}{\Omega}\right), \qquad x\geq 0 \qquad (2)$$

where $E[\alpha_l^2]=\Omega$. For convenience, we normalize Ω to 1. The phases $\theta_l(t)$ are assumed to be i.i.d. random variables uniformly distributed in $[0,2\pi)$.

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For the given channel model, the equivalent baseband representation of the received DS/SS signal can be written as

$$r(t) = \sqrt{2P} \sum_{l=0}^{L_p-1} a_l(t)c(t - lT_c - \tau)e^{j\phi} + z(t), \qquad (3)$$

where P is average signal power, L_p is the number of resolvable paths, $c(t)$ is the spreading PN signal, τ is the incoming PN code phase, $\phi_l = -\omega_c(lT_c+\tau)+\theta_l$, ω_c is the carrier frequency, and $z(t)$ represents the equivalent baseband noise which is a complex white Gaussian noise with two-side power spectrum density (PSD) of $2N_0$ watts per Hertz. The PN signal $c(t)$ is given by

$$c(t) = \sum_{k=-\infty}^{\infty} c_k P_{T_c}(t - kT_c), \qquad (4)$$

where c_k is the k-th chip of the PN sequence, P_{T_c} is the unit-amplitude rectangular pulse in the interval $[0, T_c]$, and T_c is the chip duration. We wish to obtain an estimate $\hat{\tau}$ of the phase (delay) τ.

3. Proposed Scheme

The proposed scheme is shown in Figure 2. It consists of M branches of noncoherent correlator, a decision processing block, a tapped-delay line and a PN generator. Inside of a noncoherent detector consists of a multiplier, an integrate and dump circuit, real part and imaginary part operators, two square operators and an adder as shown in Figure 3. There are 2 integration durations which depend on the operation mode, i.e., n_1T_c and n_2T_c for the first and the second steps, respectively. The decision processing block performs 4 functions that are summation, threshold test, integrate and dump control, and phase update.

The received signal is fed to the tapped-delay line. After that the delay signals are noncoherently correlated with the local PN signal and the results are sent to the decision processing block. Inside of the decision processing block, all of the correlated signals are summed together and the summation result is compared with a threshold. If it is lower than the threshold, the local PN signal phase, $\hat{\tau}$, is updated by MT_c, correlation is reset and start new correlations. On the other hand, if it exceeds the first threshold, the process continues in the second step. In the second step, results of longer correlations are sent to the decision processing block. All of the correlation results are summed together and then compared with the second threshold. If it is lower than the threshold, $\hat{\tau}$ is updated by MT_c and the process goes back to first step. On the other hand, if it exceeds the threshold, the delayed version of the received signal which gives the highest correlation value is sent to the tracking circuit to initiate the fine adjustment process.

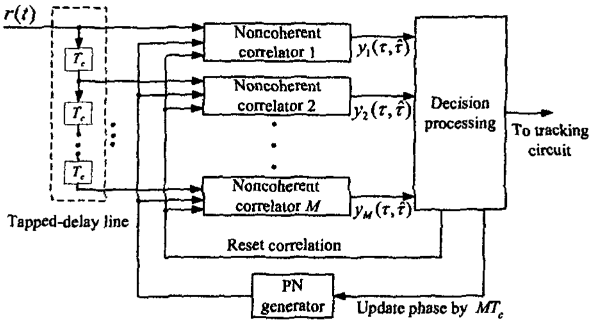

Figure 2. The proposed scheme.

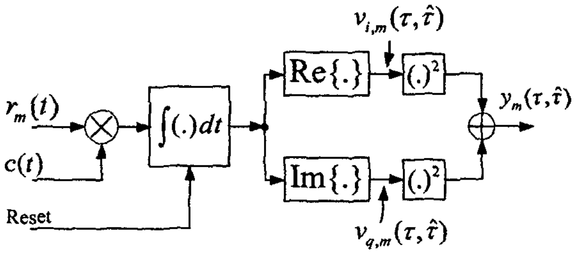

Figure 3. Noncoherent correlator.

3.1 The First Step: The First Dwell

In the first step, the integration length is set to n_1T_c. The output of the m-th noncoherent detector is

$$y_{1,m}(\tau,\hat{\tau}) = v_{i,1,m}^2(\tau,\hat{\tau}) + v_{q,1,m}^2(\tau,\hat{\tau}), \qquad (5)$$

for $m \in \{1, 2, ..., M\}$, where

$$v_{i,1,m}(\tau,\hat{\tau}) = \sum_{l=0}^{L_p-1} \cos(\phi_l)s_{l,1,m}(\tau,\hat{\tau}) + \eta_{i,1,m}, \quad (6)$$

$$v_{q,1,m}(\tau,\hat{\tau}) = \sum_{l=0}^{L_p-1} \sin(\phi_l)s_{l,1,m}(\tau,\hat{\tau}) + \eta_{q,1,m}. \quad (7)$$

The signal $s_{l,1,m}(\tau,\hat{\tau})$ is defined as

$$\begin{aligned} s_{l,1,m}(\tau,\hat{\tau}) &= \alpha_l\sqrt{2P}\int_0^{n_1T_c} c(t-\hat{\tau}) \\ &\quad \cdot c(t - lT_c - (m-1)T_c - \tau)dt. \quad (8) \end{aligned}$$

The terms $\eta_{i,1,m}$ and $\eta_{q,1,m}$ are the noise parts of $v_{i,1,m}$ and $v_{q,1,m}$, respectively. They are given by

$$\eta_{i,1,m} = \int_0^{n_1T_c} z_R(t)c(t-\hat{\tau})dt, \qquad (9)$$

$$\eta_{q,1,m} = \int_0^{n_1T_c} z_I(t)c(t-\hat{\tau})dt, \qquad (10)$$

where $z_R(t)$ and $z_I(t)$ are the real and the imaginary parts of $z(t)$. It can be shown that noises $\eta_{q,1,m}$ and $\eta_{i,1,m}$ are zero mean Gaussian random variables with variances

$$\sigma^2_{\eta_{i,1,m}} = \sigma^2_{\eta_{q,1,m}} = n_1N_0T_c. \qquad (11)$$

Inside of the decision processing block, the noncoherently correlated signals are summed and compared with threshold Γ_1. The summation signal is

$$y_1(\tau,\hat{\tau}) = \sum_{m=1}^{M} y_{1,m}(\tau,\hat{\tau}). \qquad (12)$$

If it exceeds Γ_1, the correlator goes on to the second step. On the other hand if it is lower than Γ_1, the phase is updated by MT_c and the correlation is reset.

3.2 The Second Step: The Second Dwell

In the second step, the output of the m-th noncoherent correlator is

$$\begin{aligned} y_{2,m}(\tau,\hat{\tau}) = \ & [v_{i,1,m}(\tau,\hat{\tau}) + v_{i,2,m}(\tau,\hat{\tau})]^2 \\ & + [v_{q,1,m}(\tau,\hat{\tau}) + v_{q,2,m}(\tau,\hat{\tau})]^2, \end{aligned} \qquad (13)$$

where $v_{i,1,m}$ and $v_{q,1,m}$ are the correlation results in the first step as shown in Subsection 3.1, while $v_{i,2,m}$ and $v_{q,2,m}$ are additive correlation results in the second dwell, which are given by

$$v_{i,2,m}(\tau,\hat{\tau}) = \sum_{l=0}^{L_p-1} \cos(\phi_l) s_{l,2,m}(\tau,\hat{\tau}) + \eta_{i,2,m} \qquad (14)$$

$$v_{q,2,m}(\tau,\hat{\tau}) = \sum_{l=0}^{L_p-1} \sin(\phi_l) s_{l,2,m}(\tau,\hat{\tau}) + \eta_{q,2,m} \qquad (15)$$

where $s_{l,2,m}(\tau,\hat{\tau})$, $\eta_{i,2,m}$ and $\eta_{q,2,m}$ are similar to (8), (9) and (10), respectively, except that the integration length n_1T_c is changed to n_2T_c. It can be shown that the noises $\eta_{q,2,m}$ and $\eta_{i,2,m}$ are zero mean Gaussian random variables with variances

$$\sigma^2_{\eta_{i,2,m}} = \sigma^2_{\eta_{q,2,m}} = n_2 N_0 T_c. \qquad (16)$$

Similarly to the first step, within the decision processing block, the correlated signals are summed and compared threshold Γ_2. The summation signal is

$$y_2(\tau,\hat{\tau}) = \sum_{m=1}^{M} y_{2,m}(\tau,\hat{\tau}). \qquad (17)$$

If the summation signal is higher than Γ_2, it has a high probability that one or more of the noncoherent correlators are inphase with the dispersion signal. We select the phase of the branch with the highest correlation result for tracking. Because of a noisy envelopment, there is a chance that a wrong phase is selected for tracking. However the tracking circuit has a verifying period, during which it tries to track the probable phase. If the time expires and it still cannot track, the process goes back to the first step. This time duration is called "penalty time." However, if the summation signal is lower than Γ_2, the process goes back to the the first step without any penalty time.

3.3 Probabilities of Detection and False Alarm

Let H_1 be the hypothesis that the local PN phase is within $T_c/2$ of at least one of the PN phases in the received L_p paths. Also let H_0 be the hypothesis that the local PN phase is more than $T_c/2$ away from all the PN phases in the received L_p paths.

Detection probability is the probability of accepting phase alignment when H_1 is true, while false alarm probability is the probability of accepting phase alignment when H_0 is true. Probabilities of detection and false alarm are important for computing the mean acquisition time (MAT). In this paper the acquisition time is evaluated by simulation.

Detection and false alarm probabilities of the proposed scheme are controlled by setting the value of both integration lengths n_1T_c and n_2T_c, and both thresholds, Γ_1 and Γ_2. Detection of the PN phase occurs only when the threshold in both steps are exceeded, and the maximum correlation branch is the correct branch. False alarm occurs only when the threshold in both are exceeded. Therefore, their probabilities are

$$P_d = Pr\{y_1 \geq \Gamma_1, y_2 \geq \Gamma_2, y_{max} = y_c \mid H_1\}, \qquad (18)$$

$$P_{fa} = Pr\{y_1 \geq \Gamma_1, y_2 \geq \Gamma_2 \mid H_0\}, \qquad (19)$$

where y_{max} is the value of the maximum correlation branch and y_c is the value of the branch which the phase difference is in H_1. To find these probabilities, the probability density function (pdf) of the correlated signal are obtained. However the pdf is very complicated to derive because of the fading channel model and the proposed scheme. To simplified this problem we use simulation to find their values.

4. Simulation Result

To show the performance of the proposed scheme, simulation is used. We use the conventional hybrid scheme with the same hardware for performance comparison, i.e., the conventional hybrid scheme with the same number of branches. We set the simulation at $M = 4$, and $L_p = 4$. The PN sequence is an m-sequence with polynomial $1 + x^4 + x^9$, so that the period is 511 chips, SNR = -5 and -10 dBs and penalty time is $5000T_c$. Under these conditions, simulation shows that the minimum mean acquisition time (MAT) occurs at setting values shown in Table 1. We can see that P_d and P_{fa} that give the minimum MAT are around 0.5 and 0.01, respectively, for both schemes.

The values in Table 1 were used for simulation again. Result is shown in Figure 4 which plots MAT versus the penalty time. We can see that the proposed scheme outperforms the conventional hybrid scheme at any value of SNR and penalty time. Figure 5 shows the ratio of the MAT of the conventional scheme over the MAT of the proposed scheme. We can see that the proposed scheme acquires the PN phase faster than the conventional scheme by about 1.4-2.3 times without any additional hardware.

Table 1. Simulation of best setting value of n_1, n_2, Γ_1 and Γ_2.

Scheme	SNR(dB)	n_1	Γ_1	n_2	Γ_2	P_d	P_{fa}
Con	-5	272	8.228×10^4	-	-	0.547	0.016
	-10	859	8.228×10^5	-	-	0.484	0.009
Pro	-5	72	2.419×10^4	56	6.784×10^4	0.525	0.008
	-10	226	2.419×10^5	176	6.784×10^5	0.417	0.009

Remark: Con = conventional hybrid scheme, and Pro = the proposed scheme.

5. Conclusion

We propose an improved PN acquisition scheme using RAKE receiver and double dwell technique for frequency-selective Rayleigh fading channel. The RAKE structure provides robustness against to the effect of multipath fading channel while double dwell technique provides the improved speed of acquisition. Performance is verified by simulation which shows that proposed scheme acquires the phase faster than the conventional hybrid scheme about by 1.4-2.3 times without any additional hardware.

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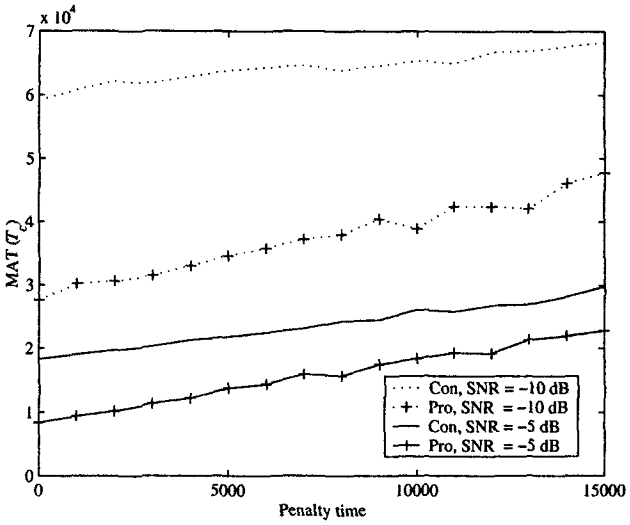

Figure 4. Mean acquisition time of the conventional hybrid scheme and the proposed scheme.

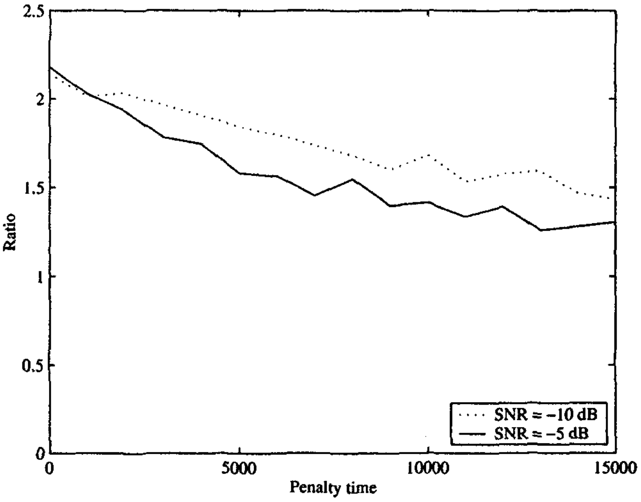

Figure 5. Ratio of mean acquisition time of the conventional scheme over the mean acquisition time of the proposed scheme at the same condition.