ered prediction orders is almost coincident, in agreement with [3], where ν = 3, or at most ν = 4, is claimed to be sufficient.

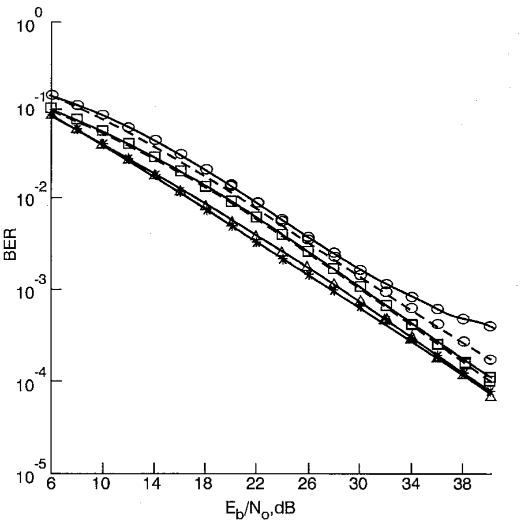

Fig. 1 *Bit error rate comparison between time-discrete observation model and time-continuous model for QPSK and* ν = 4

– – – – time-discrete model
——— time-continuous model
○——○ B_D = 0.100
□——□ B_D = 0.050
△——△ B_D = 0.010
✳——✳ B_D = 0.005
○– – –○ B_D = 0.100
□– – –□ B_D = 0.050
△– – –△ B_D = 0.010
✳– – –✳ B_D = 0.005

Fig. 1 shows, for QPSK and ν = 4, a comparison between the time-discrete observation model adopted in this Letter, which holds for slow fading, and a time continuous observation, which correctly models fast fading channels as well. In fact, in a flat fading channel, a fast time variation causes inter symbol interference (ISI) in the observation. This comparison shows that this ISI has almost no influence on the receiver performance for slowly fading channels (B_D = 0.005, 0.01, 0.05). For fast fading (B_D = 0.1), the difference in performance between the two models is still moderate, even for E_b/N_0 = 40dB.

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References

1 CLARKE, R.: 'A statistical theory of mobile radio reception', *Bell Syst. Tech. J.*, 1968, **47**, pp. 957–1000

2 YU, X., and PASUPATHY, S.: 'Innovations-based MLSE for Rayleigh fading channels', *IEEE Trans. Commun.*, 1995, **43**, part III, pp. 1534–1544

3 MAKRAKIS, D., MATHIOPOULOS, P., and BOURAS, D.: 'Optimal decoding of coded PSK and QAM signals in correlated fast fading channels and AWGN: a combined envelope, multiple differential and coherent detection approach', *IEEE Trans. Commun.*, 1994, **42**, (1), pp. 63–75

4 KONG, H., and SHWEDYK, E.: 'Detection of amplitude-phase modulated signals over frequency nonselective Rayleigh fading channels with adaptive symbol-aided channel estimation'. Veh. Tech. Conf. (VTC '96), 1996, pp. 983–987

Noncoherent SPRT-based acquisition scheme for DSSS

Jia-Chin Lin

A noncoherent sequential PN code acquisition scheme is proposed. The out-of-phase and on-phase sequences are properly modelled to avoid significantly high error probabilities occurring with the conventional SPRT-based acquisition. In addition, data modulation and frequency offset can be effectively overcome using this technique.

Introduction: By modelling the acquisition problem as testing between two hypotheses, a sequential test technique under coherent demodulation environments has been proposed and analysed [1]. However, it is, in practice, almost impossible to achieve coherent demodulation, because the signal-to-noise ratio (SNR) before despreading is very low. Several sequential probability ratio tests (SPRTs) designed for PN code acquisition under noncoherent demodulation environments have also been discussed [2 – 4], but under the assumption that the out-of-phase sequence could be modelled as a zero sequence. However, such an assumption may not be very practical, for reasons explained in the following. The modified technique proposed here is then described and simulated.

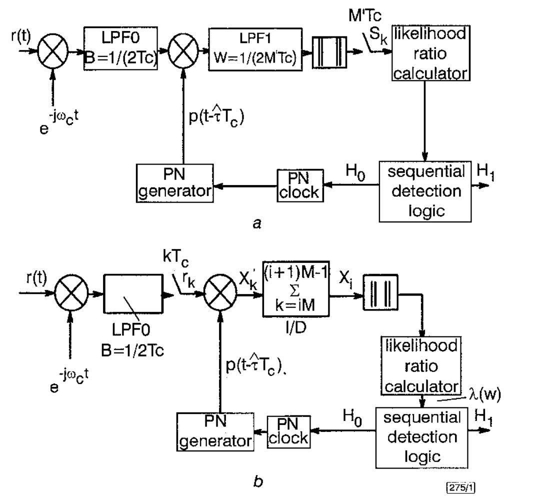

Fig. 1 *Conventional and proposed noncoherent sequential acquisition techniques*

a Conventional
b Proposed

Conventional sequential acquisition techniques: The conventional noncoherent acquisition technique based on SPRT is shown in Fig. 1*a*. The received signal is first down-converted by means of a noncoherent local carrier, $e^{j\omega_c t}$, and then passed through a lowpass filter LPF0. The resulting signal is then cross-correlated with the local code sequence. The cross-correlation signal is passed through a second lowpass filter LPF1. The output of LPF1 is then fed into an envelope detector. The envelope samples S_k are obtained by sampling the output of the envelope detector at a rate low enough (i.e. $1/M'T_c$) that the samples can be considered to be independent. Finally, the envelope samples enter the likelihood ratio calculator and the sequential detection logic.

However, in such a structure, the lowpass filter LPF1 may be very difficult to design. If its bandwidth is too wide, say $W \simeq 1/2T_c$, the cross-correlation of the incoming and the local sequences under the out-of-phase condition cannot be reduced at all. This leads to a high false alarm probability. The bandwidth W of this filter must be narrow enough, e.g. $W \ll 1/2T_c$, to reject residual cross-correlation under the out-of-phase conditions, because the likelihood ratio calculator and sequential detection logic are designed based on the assumption that the out-of-phase sequence can be modelled as a zero sequence here. Conversely, the SPRT algorithm is derived with the assumption that the samples entering the likelihood ratio calculator are sufficiently spaced, e.g. at $M'T_c$ where $M' \gg 1$, and can be considered independent, but with that assumption the carrier frequency offset and data modulation effects could possibly change the statistics of the on-phase sequence significantly and lead to a high miss probability. Therefore, it may be very difficult in practice to design and implement the conventional noncoherent sequential acquisition scheme in Fig. 1*a*.

Proposed noncoherent sequential acquisition technique: The noncoherent sequential acquisition technique proposed in this Letter is shown in Fig. 1*b*. The received signal is down-converted noncoherently, passed through a lowpass filter (LPFO) and then sampled at the chip rate to generate the baseband received sample stream r_k

$$r_k = \sqrt{\frac{E_c}{T_c}} d((k-\tau)T_c) p((k-\tau)T_c) e^{j\theta_k} + n_k \qquad (1)$$

where $\theta_k = 2\pi\Delta f \cdot kT_c + \theta_0$, and n_k is the noise sample. The resulting sequence, $\{r_k\}$, is cross-correlated with the local code sequence $p((k-\hat{\tau})T_c)$, where $\hat{\tau}$ is the local PN code phase offset and is an integer; the samples $\{X_k'\}$ are then generated as

$$X_k' = Y_k' d((k-\tau)T_c) e^{j\theta_k} + N_k' \qquad (2)$$

where $Y_k' = \sqrt{(Ec/Tc)} p((k-\tau)T_c) p((k-\hat{\tau})T_c)$ and $N_k' = n_k \cdot p((k-\hat{\tau})T_c)$. In eqn. 2, the sequence $\{Y_k'\}$ provides the information desired to detect whether the received and the local PN sequences are synchronised. It is corrupted by the data modulation effect, $d((k-\tau)T_c)$, and the carrier frequency offset effect, $e^{j\theta_k}$. It can be shown that the noise term N_k' is a complex Gaussian random variable with zero mean and the same variance as n_k. The cross-correlation samples, X_k', are integrated/dumped (I/D) for each M (non-overlapped) samples to generate the I/D cross-correlation samples X_i:

$$X_i = \sum_{k=iM}^{(i+1)M-1} X_k' = Y_i e^{j\theta_0} + N_i \qquad (3)$$

where

$$Y_i = \sum_{k=iM}^{(i+1)M-1} Y_k' e^{j(2\pi\Delta f k T_c)} \quad \text{and} \quad N_i = \sum_{k=iM}^{(i+1)M-1} N_k'$$

Based on the model of the partial correlation of PN sequences [1], the acquisition problem can be considered as the testing of a composite hypothesis H_0 against a simple alternative H_1 by means of the envelopes of the I/D cross-correlation samples, X_i. Hence, the desired information sequence, $\{Y_i\}$, in eqn. 3 can be modelled either by the constant $\hat{Y}^1 = f_1 \cdot M$ for H_1 (i.e. $\tau = \hat{\tau}$) or $\hat{Y}^0 \leq f_0 \cdot M$ for H_0 (i.e. $\tau \neq \hat{\tau}$), where $f_1 = \|\Sigma_{k=1}^{M} e^{j(2\pi\Delta f k T_c)}\|$ is a factor introduced by the carrier frequency offset which approaches 1 as M decreases, and $f_0 = 0.5$ is used here because it corresponds to the worst-case correlation under synchronisation for the case whenever the chip timing is unknown and the local PN generator is updated by T_c for each H_0 claimed. Based on the model of Y_i, and after some manipulation, the envelope samples, $\|X_i\|$, of the I/D correlation samples can be described by iid Rician random variables with parameter either $\hat{Y}^1$ for H_1 or $\hat{Y}^0$ for H_0. The log-likelihood ratio for the ith envelope sample, r_i, is then

$$Z_i = \frac{(\hat{Y}^0)^2 - (\hat{Y}^1)^2}{2M\sigma_0^2} + \log\left(I_0\left(\frac{\hat{Y}^1 r_i}{M\sigma_0^2}\right)\right) - \log\left(I_0\left(\frac{\hat{Y}^0 r_i}{M\sigma_0^2}\right)\right) \qquad (4)$$

and we will consider the sequential test using the statistic $\lambda(w) = \Sigma_{i=1}^{w} Z_i$. The SPRT is described by

$$\lambda(w) \begin{cases} \geq B & \Rightarrow H_1 \\ \leq A & \Rightarrow H_0 \\ \in (A, B) & \Rightarrow \text{take another envelope sample} \end{cases} \qquad (5)$$

where the threshold levels, A and B, are given [3], so that the test can achieve the false alarm probability α and the miss probability $1-\beta$.

Simulation results: The false alarm and miss probabilities of the proposed and conventional sequential acquisition techniques are shown in Fig. 2. The false alarm and miss probabilities in the case of a carrier frequency offset $\Delta f/(1/T_c) = 0.025$ (i.e. $\Delta f/(1/T) = 3.2$ for processing gain = 128), $M' = 4$ for the conventional technique and $M = 4$ for the proposed technique are plotted in Fig. 2*a* and *b*, while those under the same condition except for $M' = 8/M = 8$ are shown in Fig. 2*c* and *d*. From Fig. 2*a*, the false alarm probability of the conventional SPRT is significantly higher than the given false alarm probability when M' is small, e.g. $M' = 4$, because the bandwidth, $W = 1/2M'T_c$, of LPF1 is too wide to reject the residual cross-correlation under the out-of-phase conditions, but the out-of-phase sequences were incorrectly modelled as a zero sequence. For a larger M' value, e.g. $M' = 8$, the false alarm probability of the conventional technique is reduced (Fig. 2*c*), because the narrower bandwidth of LPF1 can reject more residual cross-correlation; however, such a narrow bandwidth may accumulate the effects of the carrier frequency offset and then change the statistics of the on-phase sequence. This may lead to a much higher miss probability (Fig. 2*d*). From Fig. 2, no matter how the bandwidth of LPF1 is chosen, the error probabilities (false alarm/miss probabilities) of the conventional technique cannot be accepted; no such problem occurs with the proposed technique.

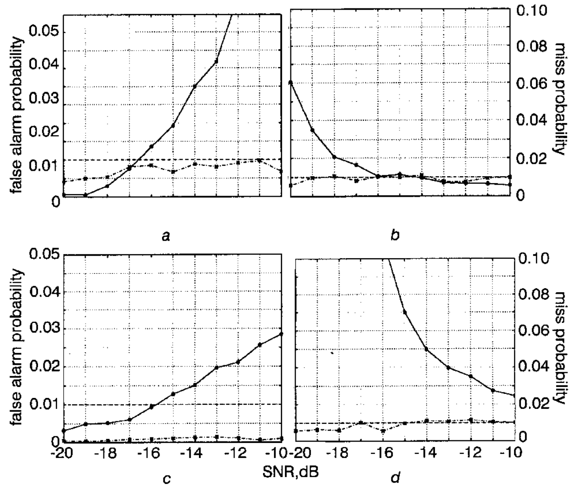

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Fig. 2 *False alarm and miss probabilities for conventional and proposed SPRT when $\Delta f/(1/T_c) = 0.025$ and desired error probability is 10^{-2}*

—O— conventional
– – – – desired
– · * · – proposed
a False alarm; M', $M = 4$
b Miss; M', $M = 4$
c False alarm; M', $M = 8$
d Miss; M', $M = 8$

Conclusion: In this Letter, a noncoherent sequential PN code acquisition technique is proposed. The out-of-phase sequence has been modelled as the upper bound of the cross-correlation between any two non-synchronised sequences, and the on-phase sequence has been modified by a factor caused by the carrier frequency offset effect to avoid significantly high error probabilities occurring with the conventional SPRT-based acquisition technique.

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References

1 LEE, Y.-H., and TANTARATANA, S.: 'Sequential acquisition of PN sequences for DS/SS communications: Design and performance', *IEEE J. Sel. Areas Commun.*, 1992, pp. 750–759

2 SIMON, M., OMURA, J., SCHOLTZ, R., and LIVITT, K.: 'Spread spectrum communication' vol. III, (Computer Science Press, Rockville, MD, 1985)

3 WALD, A.: 'Sequential analysis' (Wiley, New York, 1947)

4 SU, Y.-T., and WEBER, C.L.: 'A class of sequential tests and its applications', *IEEE Trans. Commun.*, 1990, **38**, (2), pp. 165–171