Forward Transmission over a Multipath Faded Channel in a DS CDMA System

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Abstract—the pseudorandom spreading sequence in the forward transmission in the IS-95 system is assigned only to individual base stations. To separate users, the system assigns orthogonal (Hadamard) sequences to each of them. In this paper, a different scheme for the forward transmission is considered. In this scheme pseudorandom sequences are used, instead of employing orthogonal spreading sequence for individual users. The effect of fade amplitudes, phase shifts, time-mismatching and AWGN on the detection process of Gold Sequence Spread Spectrum Signal (GSSSS) is considered.

Key-words: Gold sequence, Rake receiver, Time-mismatching, QPSK-GS SS receiver.

1 Introduction

Spread-spectrum access techniques allow multiple signals to occupy the same radio frequency (RF) bandwidth to be transmitted simultaneously without interfering with each other. One of the applications of spread spectrum (SS) communication technology is cellular mobile radio, exploits fourth and fifth benefits to provide resistance to signal interference from multiple transmission paths and potentially higher bandwidth efficiency in multiple access communication than in other technologies. In GS/CDMA systems, the narrowband message signal is multiplied by a very large-bandwidth signal called spreading signal [1].

In IS-95, the code sequences from individual users and the pilot sequence are multiplied by the base station-specific in-phase and the Quadrature of spreading sequences. The separation of users is achieved by employing orthogonal signals. This, in turn, is done by assigning one of the N Hadamard sequences with duration N to spread each codeword of the inner code. When the base station receives and recognizes a newly accessing user it will assign it as a Hadamard sequence.

2 Transmission of GSSSS

In this system [2], it assumed that K active users simultaneously send digital information by using a modulated SS signal $s(t)$ carrying digital information and this signal can be represented as a sequence of signal waveforms $\tilde{s}_n(t-\tau_n)$, where $...\tau_{-1} < \tau_0 < \tau_1 < ... < \tau_n < ...$ and

$$s(t) = \sum_{n=-\infty}^{\infty} \tilde{s}_n(t-\tau_n) \qquad (1)$$

In GS/CDMA applications number of pulses can be used in modulation formats. Given the input modulator symbol $v_n, n = ...-1,0,1,2,..., v_n \in \{1,-1\}$, the signal waveform in the nth time instant $\tau_n, ... < \tau_{-1} < \tau_0 < \tau_1 < \tau_2 < ... < \tau_n < ...,$ is [2]

$$\tilde{s}_n(t-\tau_n) = \frac{(v_n+1)}{\sqrt{2}} h_{T_P}(t-\tau_n) \qquad (2)$$

where $h_{T_P}(t)$ is a pulse with duration T_P, $\tau_n = nT_P$, and v_n is a code sequence that defined by:

$$v_n^{(k)}(t) = \sum_{n=-\infty}^{\infty} v_n h_{T_P}(t-nT) \qquad (3)$$

The kth transmitter's output signal $s^{(k)}(t)$ is given by:

$$s^{(k)}(t) = \sum_{n=-\infty}^{\infty} \frac{(v_n^{(k)}+1)}{\sqrt{2}} h_{T_P}(t-nT_f - a_n^{(k)}\Delta) \qquad (4)$$

where Δ is the duration of the addressable time delay bin (practically it can be equal to T_P) and T_f is the frame time ($T_f >> T_P$) $a_n^{(k)}$ is called the addressable pulse position shift, which provides an additional time shift to each pulse in the pulse train. The elements $a_n^{(k)}$ of the sequence are chosen from a finite set {-(Q-1)/2, (Q-3)/2… (Q-1)/2], where $Q = T_f / \Delta$ and Q>>1 [2].

Suppose now that user sends a data signal $u(t)$ that represents a pulse train of positive and negative pulses $u(t) = \{1,-1,1,1,-1,...\}$ as shown in fig 1.

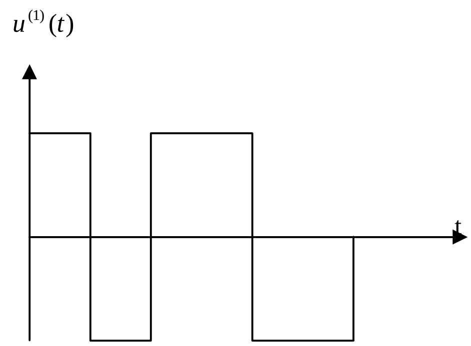

Fig 1. User data sequence

The spread signal $a_n^{(k)}$ can be generated by mapping the spreading sequences $a^{(1)} = a_0, a_1, \cdots, a_n^{(1)} \in \{1,-1\}$ for one user. Then the transmitted signal for one user (see formula 3 and 4) without repeating of each symbol is:

$$s^{(k)}(t) = \sum_{n=-\infty}^{\infty} u(t) h_{T_P}(t - T_f - a_n^{(k)}\Delta) \qquad (5)$$

A Gold sequence is constructed by the XOR of two m-sequences with the same clocking [3, 4, 5, 6].

Using equation 5 and spreading Gold sequence code from pseudo noise generator where the number of chips N=31 and using a shift register implementation as shown in fig 2.

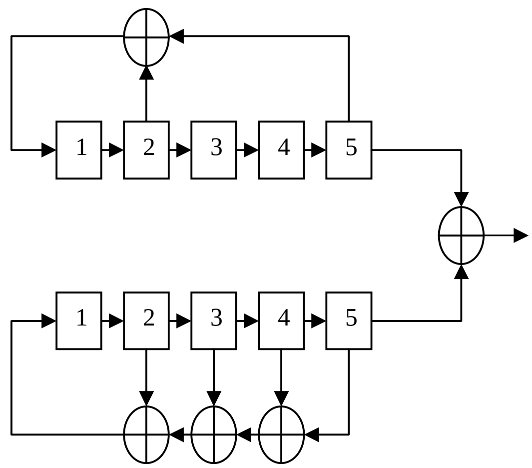

Fig 2 Gold sequence pseudonoise generators

The resulting GS CDMA for bit stream $u_n^{(k)} = v_n^{(k)} a_n^{(k)} = \{ 1,-1, 1, 1,-1,-1\ldots \} a_n^{(k)}$ using PPH CDMA for positive one and of gold sequence is $a_{31}^{(k)} = \{1000010001000101000110001101011\}$

The transmitter's output signal $s^{(1)}(t)$ for binary one is given by:

$$s^{(k)}(t) = \sum_{n=0}^{30} \frac{(v_n^{(k)}+1)}{\sqrt{2}} h_{T_P}(t - nT_P - a_n^{(k)}\tau_n)$$

$$\times \cos(2\pi f_c n + \phi^{(k)}) \quad (6)$$

where $T_P = T_f$ for single user.

We suppose that the rate $r = 1/N$ repetition code is used, that is, $v_n = u_{\lfloor n/N \rfloor}$, and that the phases are random variables uniformly distributed on the interval $[0,2\pi]$ and known to both transmitter and receiver.

Without loss of generality we assume that the phase $\phi^{(0)}$ of the pilot signal is equal to zero. The pilot signal is:

$$s^{(0)}(t) = \sqrt{2} \sum_{n=0}^{30} h_{T_P}(t - nT_P - a_n^{(0)}\tau_n)$$

$$\times \cos(2\pi f_c n) \quad (7)$$

Assuming that pilot's spreading sequence is shared by all users, we multiply the user sequence by the pilot sequence to identify the user with the particular base station that is handling the call.

Let $a_n^{(0)}$ be the pilot's BPSK sequence [2]. The user's sequences are the product of those of the pilot and of the user -specific sequence $\tilde{a}_n^{(k)}$. That is:

$$a_n^{(k)} = a_n^{(0)} \tilde{a}_n^{(k)} \quad (8)$$

The transmitted signal is:

$$s(t) = \gamma^{(0)} s^{(0)}(t) + \sum_{k=1}^{K} \gamma^{(k)} s^{(k)}(t) \quad (9)$$

where $s^{(0)}(t)$ is defined by Formula (7) and $s^{(k)}(t), k = 1,2,\cdots,K$, is defined by (6). The weight coefficients $\gamma^{(k)}$ should be chosen such that all users have the same SNR at the receiver input.

3 Receiving of GSSSS

The received signal contains signals of K users and a pilot signal all are originating from a common base station, is:

$$r(t) = r^{(0)}(t) + \sum_{k=1}^{K} r^{(k)}(t) \quad (10)$$

here

$$r^{(0)}(t) = \gamma^{(0)} \sum_{l=1}^{L} \alpha_l \sum_{n=0}^{30} h_{T_p}(t - nT_p - a_n^{(0)} - \delta_l)$$

$$\times \cos[2\pi f_c(n - \delta_l) + \varphi_l] + \sum_{l=1}^{L} \xi_l(t) \quad (11)$$

and

$$r^{(k)}(t) = \gamma^{(k)} \sum_{l=1}^{L} \alpha_l \sum_{n=0}^{30} \frac{(v_n^{(k)} + 1)}{\sqrt{2}}$$

$$\times h_{T_p}(t - nT_p - a_n^{(k)}\tau_n)$$

$$\times \cos[2\pi f_c(n - \delta_l) + \varphi_l + \phi^{(k)}] \quad (12)$$

where $\xi_l(t)$ are AWGN processes.

Because the transmission is from one base station, the fade amplitudes α_l, delays δ_l, and phase shifts φ_l do not depends on k. The amplitudes delay are assumed to be IID Rayleigh random variables that have the probability density functions:

$$P_b(\alpha_1, \alpha_2, \cdots, \alpha_L) = \frac{1}{2}\exp\left(-T_p N \sum_{l=1}^{L} \alpha_l^2 / N_0\right) \quad (13)$$

4 Pilot-Aided Rake Receiver

The quadrature outputs of a pilot signal despreader dependent on the estimated time delay of a given path are:

$$\hat{z}_{nl} = \gamma^{(0)} \alpha_l \cos\varphi_l + \varphi_l$$

$$\times \sum_{k=1}^{K} \gamma^{(k)} \frac{v_n^{(k)} + 1}{\sqrt{2}} h_{T_p}(t - nT_p - \tilde{a}_n^{(k)}\tau_n)$$

$$\times \cos(\varphi_l + \phi^{(k)}) + \hat{\xi}_{nl} \quad (14)$$

and

$$\breve{z}_{nl} = \gamma^{(0)} \alpha_l \sin\varphi_l + \varphi_l$$

$$\times \sum_{k=1}^{K} \gamma^{(k)} \frac{v_n^{(k)} + 1}{\sqrt{2}} h_{T_p}(t - nT_p - \tilde{a}_n^{(k)}\tau_n)$$

$$\times \sin(\varphi_l + \phi^{(k)}) + \breve{\xi}_{nl} \quad (15)$$

The first terms are due to the pilot signal, the second terms are due to given cell user's signal, $\hat{\xi}_{nl}$ and $\breve{\xi}_{nl}$ are due to the AWGN and the other-cell interference. The value of v should be as large as possible without exceeding the period over which α_l and φ_l remain relatively constant. Then

$$\hat{\mu}_{Il} \approx \frac{1}{31}\sum_{n=0}^{30}\hat{z}_{nl}^{(k)} \tag{16}$$

and

$$\hat{\mu}_{Ql} \approx \frac{1}{31}\sum_{n=0}^{30}\breve{z}_{nl}^{(k)} \tag{17}$$

The Neyman-Pearson criterion uses $\hat{\sigma}_l^2, l = 1,2,\cdots,L,$ for calculating the decision statistics y_l. Although, the variances σ_l^2 are not known to the receiver, we can estimate them directly from (14) and (15):

$$\sigma_l^2 = \frac{1}{2(31-1)}\sum_{n=0}^{30}(\hat{z}_{nl} - \hat{\mu}_{Il})^2 + \frac{1}{2(31-1)}\sum_{n=0}^{30}(\breve{z}_{nl} - \breve{\mu}_{Ql})^2 \tag{18}$$

The pilot-aided coherent Rake receiver for the *kth* user consists of L fingers [2]. The pilot sequence tracking loop estimates the time delay of a given path to remove the pilot BPSK spread and to give rise to the quadrature components $\hat{z}_{nl}$ and $\breve{z}$. The decision statistics at the output of *L*-path demodulator can be considered as independent and Gaussian because of the central limit theorem:

$$y_l^{(k)} = \frac{2}{\sigma_l^2}\sum_{n=0}^{30} z_{nl}^{(k)},\ l = 1,2,\cdots,L \tag{19}$$

Taking $\hat{\alpha}_l = \alpha_l$, $\hat{\varphi}_l = \varphi_l$ and $\hat{\sigma}_l^2 = \sigma_l^2$, the statistics y_l are independent conditionally Gaussian given α_l with expectation:

$$E(y_l^{(k)}) = \begin{cases} 4\rho_{cl}^{(k)}, & \text{if } H_1 \text{ is true,} \\ -4\rho_{cl}^{(k)}, & \text{if } H_2 \text{ is true} \end{cases} \tag{20}$$

Where:

$$\rho_{cl}^{(k)} = \frac{(\gamma^{(k)})^2\alpha_l^2}{\alpha_l^2[\psi - (\gamma^{(0)})^2] + N_0/T_p + N_{oc}^{(k)}/T_p} \tag{21}$$

is the *kth* user, *lst* path SNR per chip, and

$$\psi = \sum_{k=0}^{K}(\gamma^{(k)})^2 \tag{22}$$

is the normalized average power of the signal (5), $N_0/2$ is the two sided power spectral density of AWGN and $N_{oc}/2$ is two-sided power spectral density of the other cell interference noise on the input of the *lst* finger of the *kth* user Rake receiver. The decision rule is:

$$y^{(k)} = \sum_{l=1}^{L} y_l^{(k)} \underset{H_2}{\overset{H_1}{\gtrless}} 0 \tag{23}$$

Where: H_1 and H_2 are hypotheses of decision rule.

The kth user conditional bit error probability is upper bounded by:

$$P_b^{(k)}(\alpha_1,\alpha_2,\cdots,\alpha_L) < \frac{1}{2}\exp\left(-30\sum_{l=1}^{L}\rho_{cl}^{(k)}\right) \qquad (24)$$

5 General Experemental Results

Experimental results were obtained successfully as shown below:

5.1 The effect of fade amplitudes on $P_b^{(k)}$

Assuming that perfect synchronization $\varphi_l = 0$, zero delays $\delta_l = 0$ and perfect synchronization, the quadrature outputs of the pilot signal despreader [compared with (14) and (15)] become

$$\hat{z}_{nl} = \gamma^{(0)}\alpha_l \qquad (25)$$

and

$$\breve{z}_{nl} = 0 \qquad (26)$$

In our case ($k = 1$) for one user equation 16 and 17 become:

$$\hat{\mu}_{Il} \approx \frac{1}{31}\sum_{n=0}^{30}\gamma^{(0)}\alpha_l \qquad (27)$$

and

$$\hat{\mu}_{Ql} = 0 \qquad (28)$$

The variance σ_l^2 can be written as:

$$\sigma_l^2 = \frac{1}{60}\sum_{n=0}^{30}(\gamma^{(0)}\alpha_l - \sum_{n=0}^{30}\gamma^{(0)}\alpha_l) \qquad (29)$$

The decision statistics at the output of first path demodulator (for the first user) is:

$$y_l^{(1)} = \frac{2}{\sigma_l^2}\sum_{n=0}^{30}\gamma^{(1)}\alpha_l^2 + \tilde{\xi}_{nl}^{(1)} \qquad (30)$$

where the decision rule is:

$$y^{(k)} \overset{def}{=} \frac{2}{\sigma_l^2}\sum_{n=0}^{30}\gamma^{(1)}\alpha_l^2 \underset{H_2}{\overset{H_1}{\gtrless}} 0 \qquad (31)$$

As mentioned above, all weight coefficients should be chosen such that all users have the same SNR at the receiver input. The first user conditional bit error probability is upper- bounded (see formula 24) by

$$P_b^{(1)} < \frac{1}{2}\exp\left(\frac{-30}{\psi} + \frac{\alpha_1^2}{N_0/T_p + N^{(k)}{}_{oc}/T_p}\right) \qquad (32)$$

From equation 32, it is clearly shown that the probability of error depends on the value of the fade amplitudes.

Table 1 shows the values of normalized energies for different fade amplitudes and fixed probability of bit error $P_b = 0.1$.

L	α_l^2	ψ
1	0.1	20
2	0.2	14
3	0.3	11
4	0.4	10

TABLE I. PROBABILITY OF ERROR FOR DIFFERENT FADE AMPLITUDES AND NORMALIZED ENERGIES.

These results can be explained as follows:

- To get a fixed probability of error depending on the fade amplitudes, we should decrease the normalized energy.
- The fade amplitudes and normalized energy are reversed proportional.

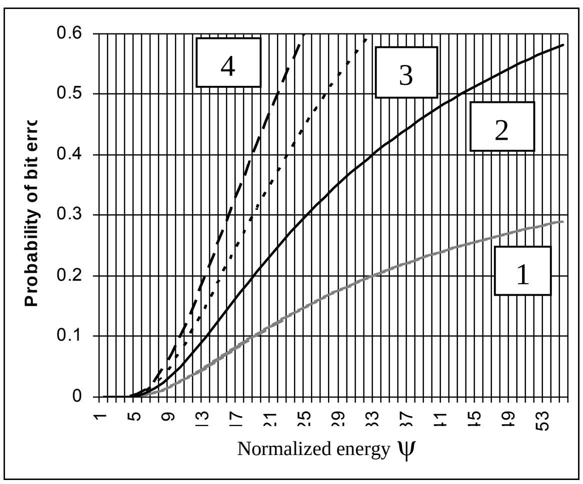

Fig 3. Upper bounds of the error probability P_b for a downlink GS CDMA system that operates in an L-path Rayleigh channel, 1-$\alpha_1^2 = 0.1$, 2-$\alpha_2^2 = 0.2$, 3-$\alpha_3^2 = 0.3$ and 4-$\alpha_4^2 = 0.4$

5.2 The effect of phase shifts on $P_b^{(k)}$

Assuming that zero delays $\delta_l = 0$, equal amplitudes $\alpha_l^2 = const$, zero effect of AWGN and interference $\hat{\xi}_{nl} = \breve{\xi}_{nl} = 0$, the quadrature outputs of the pilot signal despreader [compared with (14) and (15)] for the first user become

$$\hat{z}_{nl} = \gamma^{(0)}\alpha_l \times \gamma^{(1)} \frac{v_n^{(1)}+1}{\sqrt{2}} h_{T_p}(t - nT_p - \tilde{a}_n^{(1)}\tau_n)\cos(\phi^{(1)}) \quad (34)$$

and

$$\breve{z}_{nl} = \gamma^{(1)} \frac{v_n^{(1)}+1}{\sqrt{2}} h_{T_p}(t - nT_p - \tilde{a}_n^{(1)}\tau_n)\sin(\phi^{(1)}) \quad (35)$$

where

$$\hat{\mu}_{Il} \approx \frac{1}{31}\sum_{n=0}^{30} \hat{z}_{nl}^{(k)} \text{ and } \hat{\mu}_{Ql} \approx \frac{1}{31}\sum_{n=0}^{30} \breve{z}_{nl}^{(k)} .$$

The variance σ^2 can be computed using equation (18). The decision statistics is:

$$y_l^{(k)} = \frac{2}{\sigma_l^2}\sum_{n=0}^{30}\sqrt{(\hat{z}_{nl}^{(1)} + \breve{z}_{nl}^{(1)})^2}, \quad (36)$$

Assuming

$$E_n^{(1)} = \gamma^{(1)} \frac{v_n^{(1)}+1}{\sqrt{2}} h_{T_p}(t - nT_p - \tilde{a}_n^{(1)}\tau_n) \quad (37)$$

the decision statistics can be written as

$$y_l^{(k)} = \frac{2}{\sigma_l^2}\sqrt{[\gamma^{(0)}\alpha_l \cos(\phi^{(1)}) + \sin(\phi^{(1)})]^2}$$

$$\times \sum_{n=0}^{30} \sqrt{E_n^{(1)}} \qquad (38)$$

where $E_b = \sum_{n=0}^{30} (E_n^{(1)})^2$ is the bit energy of the resulting DS CDMA for bit stream $u_n^{(k)} = v_n^{(k)} a_n^{(k)} = \{1,-1,\ 1,\ 1,-1,-1\ldots\} a_n^{(k)}$ using PPH CDMA for positive one and a sequence of gold sequence.

The final decision statistics of the proposed GS CDMA depends on the right choice of PN sequence to get maximum bit energy, which is the result of correlation process at the output of L-path demodulator, that is

$$y^{(1)} = \frac{2}{\sigma^2}\sqrt{[\gamma^{(0)}\alpha\ \cos(\phi^{(1)}) + \sin(\phi^{(1)})]^2} \times \sqrt{E_b} \qquad (39)$$

The ideal case, when the phase shifts=0, then

$$y^{(1)} = \frac{2\gamma^{(0)}\alpha\sqrt{E_b}}{\sigma^2} \qquad (40)$$

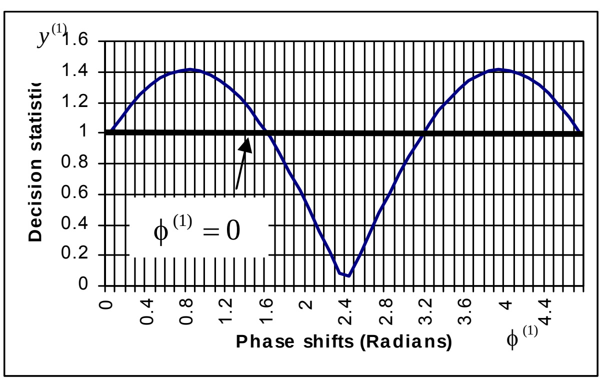

Fig 4. The effect of phase shifts on the decision statistics

Fig 4 shows that we can get a maximum decision statistics if we have a phase shift that equals to (0.8 ± 3.14) radians ($45^0 \pm 180^0$ degrees). The value of decision statistics in the range (0-1.6) radians is satisfying and giving values larger than one.

6 Analysis of Proposed DS Spread Spectrum Signal

To demodulate the received DS signal we use QPSK GS/CDMA receiver [7] (See fig. 5).

First we should calculate the E_b as a result of coherent detection process of the received signal (Assuming zero delays $\delta_l = 0$, equal amplitudes $\alpha_l^2 = const$ and zero effect of AWGN and interference $\hat{\xi}_{nl} = \breve{\xi}_{nl} = 0$) [compared with equation (12)] in our case its binary one is to eliminate the carrier frequency.

$$\Big(r^{(1)}(t) = \gamma^{(1)}\alpha \sum_{n=0}^{30} \frac{(v_n^{(k)} + 1)}{\sqrt{2}}$$

$$\times h_{T_p}(t - nT_p - a_n^{(k)}\tau_n)$$

$$\times \cos[2\pi f_c(n) + \phi^{(k)}] \Big) \qquad (41)$$

And the locally generated in-phase and quadrature signals (at front of the receiver):

$$\begin{array}{l} \cos(2\pi f_c t) \\ \sin(2\pi f_c t) \end{array} \qquad (42)$$

Now, suppose that the sampling process at T_c is perfect and impulse response of the matched filter is the mirror image of the input signal, equations (15) and (16) become at the output of matched filters with one sampling period T_c:

$$z_{In}^{(1)} = \sum_{n=-30}^{30} \frac{(v_n^{(1)}+1)}{T_c\sqrt{2}} h_{T_P}(t - nT_P - a_n^{(1)}\tau_n)$$

$$\times \cos^2(2\pi f_c n) h_c(-t) a_n^1 \qquad (43)$$

And:

$$z_{Qn}^{(1)} = \sum_{n=-30}^{30} \frac{(v_n^{(1)}+1)}{T_c\sqrt{2}} h_{T_P}(t - nT_P - a_n^{(1)}\tau_n)$$

$$\times \sin^2(2\pi f_c n) h_c(-t) a_n^1 \qquad (44)$$

Using equations (41), (43) and (44) and assume that the transmitted signal is binary one (+1), we get:

$$y^{(1)}{}_{(one)} = \frac{2}{\sigma^2}\sqrt{[\gamma(0)\alpha\cos(\phi^{(1)}) + \sin(\phi^{(1)})]^2}$$

$$\times \sum_{n=-30}^{30} \frac{\sqrt{2}(v_n^{(1)}+1)}{T_c} h_{T_P}(t - nT_P - a_n^{(1)}\tau_n) h_c(-t) a_n^1 \qquad (45)$$

Where:

$$\cos^2(2\pi f_c n) + \sin^2(2\pi f_c n) = 1$$

6.1 The effect of phase shifts

The output of matched filter is a signal that is shown in fig 6. The main bit normalized energy is accumulated

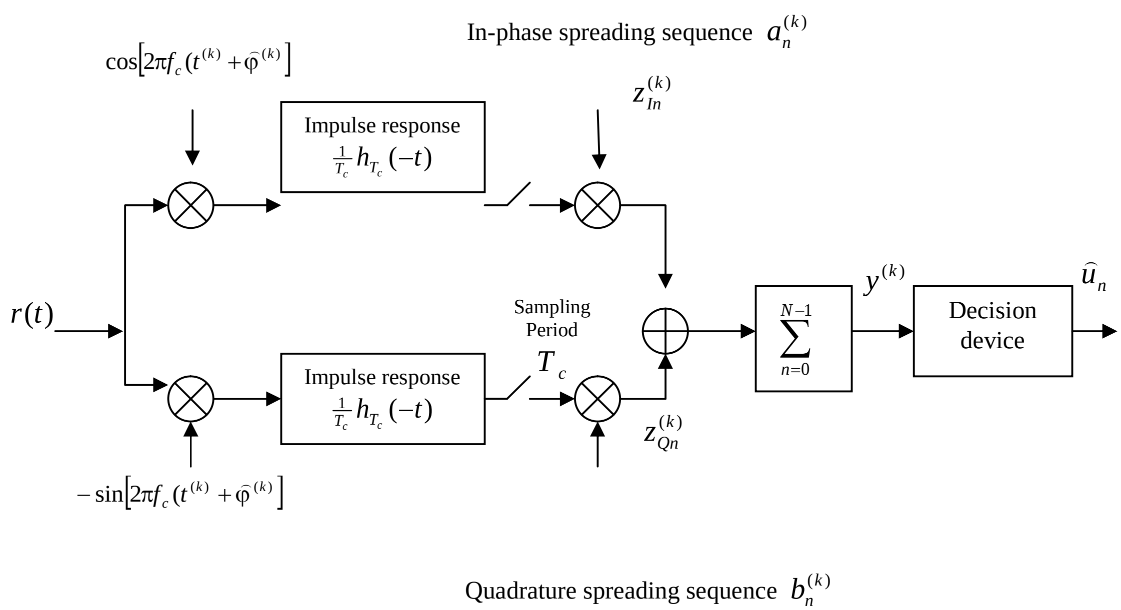

Fig 5. QPSK GS/ SS receiver

In the main beak ($n = 31$) and equals to $E_{acc(main)} = 0.89$. Where the maximum normalized accumulated energy in other peaks $E_{acc(sub)} = 0.64$. These results show that the threshold level should be equal to $E_{acc(sub)}$ to get right decision.

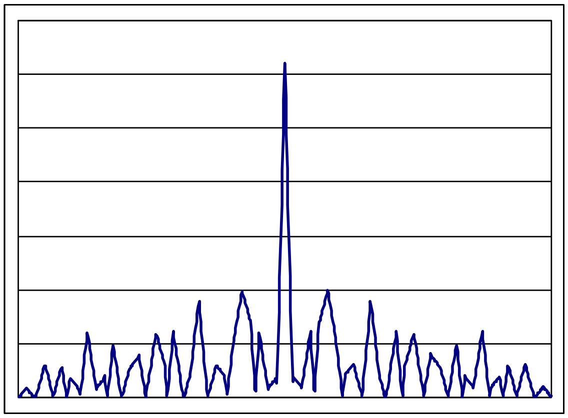

Fig 6. The signal at the output of ideal matched filter

6.2 The effect of phase shifts and time-mismatching:

The following table shows the effect of time-mismatching on $E_{acc(main)}$, $E_{acc(sub)}$ and $y^{(1)}_{(one)}$ using gold sequence at the output of matched filter. We assumed that $\phi^{(1)} = 0$.

TABLE II. THE EFFECT OF TIME-MISMATCHING

Timemismatching τ_n (%)	$E_{acc(main)}$	$E_{acc(sub)}$	$\frac{E_{acc(sub)}}{E_{acc(main)}}$
0	0.89	0.31	0.348
10	0.81	0.27	0.333
20	0.73	0.27	0.370
30	0.70	0.24	0.343
40	0.65	0.22	0.338
50	0.60	0.21	0.350
60	0.54	0.21	0.389
70	0.53	0.19	0.358
80	0.50	0.17	0.340
90	0.48	0.18	0.375
100	0.40	0.16	0.400

The results from table 2 show that the effect of time-mismatching is very small and give us a constant probability of bit error.

If we considered the effect of time-mismatching and phase shifts, then of course the probability of bit error will be affected as shown in fig 7.

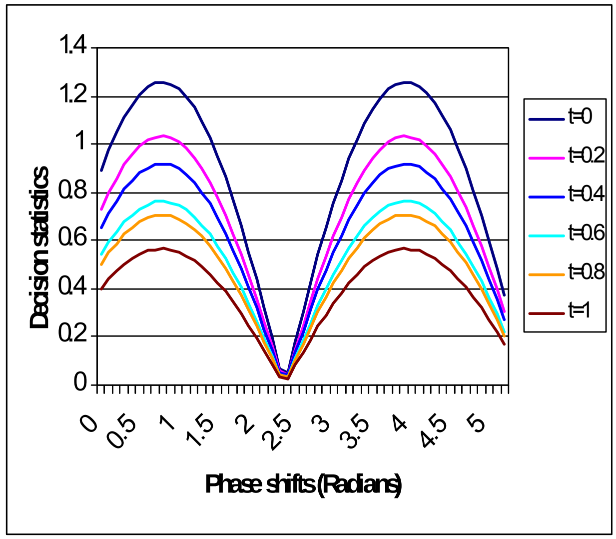

Fig 7.The effect of phase shifts and time-mismatching on the decision statistics.

6.2 The effect of AWGN and time-mismatching

Assuming zero phase shifts $\phi^{(1)} = 0$, equation (45) can be written as

$$y^{(1)}_{(one)} = \frac{2\gamma^{(0)}\alpha}{\sigma^2} \times \sum_{n=-30}^{30} \frac{\sqrt{2}(v_n^{(1)}+1)}{T_c}$$

$$\times h_{T_P}(t - nT_P - a_n^{(1)}\tau_n)h_c(-t)a_n^1 + \xi^{(1)} \qquad (46)$$

where $\xi^{(1)}$ is the AWGN process.

Fig 8 shows the main energy decreases for different variances. But, still max if the time-mismatching equals to 40% of τ.

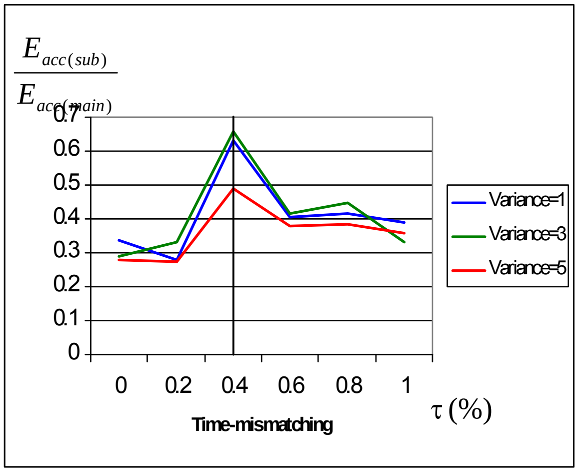

Fig 8. The effect of AWGN and time-mismatching on $\frac{E_{acc(sub)}}{E_{acc(main)}}$

7 Conclusions

The transmitted gold sequence spread spectrum with number of chips consists of two main components:

- Pilot signal to achieve synchronization between transmitter and receiver.
- User information signal [Equation (8) and (9)]

General analysis of received signal has been introduced and it was shown that the main sources of ambiguous is fade amplitudes α_l, delays δ_l, phase shifts φ_l AWGN of the system and interference from other users that use the same DSSS communication system.

The uses of pilot-aided Rake receiver gives a conditional bit error probability that dose not depend on the number of users [equations (21) and (24)].

The effect of fade amplitudes decreases $P_b^{(k)}$ and this phenomenon can be compensated by increasing the normalized energy per one user [Fig.3].

The effect of phase shifts on the decision statistics shows that we can get a maximum decision statistics if we have a phase shift that equals to (0.8 ± 3.14) radians ($45^0 \pm 180^0$ degrees). The value of decision statistics in the range (0-1.6) radians is satisfied and gives values larger than one.

The result of using QPSK GS/ SS receiver to detect the proposed gold sequence spread spectrum is an information signal with energy that depends on the correct choice of GSSS signal at the transmitter.

The effects of time-mismatching, time-mismathcing and phase shifts and AWGN and time-mismatching are shown on table 2, fig. 7 and fig. 8 respectively.

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