FAST ACQUISITION OF A PHASE-COHERENT DSFH SATELLITE SIGNAL

XIAO CHEN, GUANGXIA LI, ZHIQIANG LI, WENMING ZHU

Key Lab of Satellite Communication, College of Communications Engineering, PLA University of Science and Technology, Nanjing 210007, China
Email:ty_chenxiao@163.com

Abstract:

A hybrid direct sequence and frequency hopping (DSFH) spread spectrum signal can greatly improve ranging precision and jamming resisting capability in spacecraft TT and C(Tracking, Telemetry and Command). The frequency hopping adds another searching dimension to acquisition and increases the acquisition complexity. Therefore, this paper designs a phase-compensating-based acquisition algorithm. First, the model DSFH signal is introduced. Then the auto-correlation function and the acquisition problem are analyzed. Finally, a phase-compensating-based acquisition algorithm is proposed, its computation complexity is analyzed, and the signal-to-noise ratio (SNR) losses due to necessary approximations are shown. Simulations validate the advantageous performance of our algorithm. The proposed method can be modified to adapt to different PN chip rates and frequency hopping rates.

Keywords:

Satellite TT&C; Acquisition; anti-jamming; DSFH signal;

1. Introduction

Communication systems of military spacecraft suffer from hostile jamming and spying attempts, which are particularly dangerous for the TT&C subsystem. Although current TT&C systems use the direct sequence, (DS) spread spectrum to make the signal spectrum low-lying and unnoticeable [1], the importance of TT&C link calls for stronger anti-jamming capability.

DSFH signal modulates the DS signal to a frequency-hopping radio frequency and spreads the signal spectrum even wider. Combining frequency hopping and PN code randomizing schemes, we see two main advantages for TT&C. First, the hybrid DSFH spread spectrum technique can provide a secure satellite TT&C link with high jamming-resisting capability. Second, the ranging precision of DSFH signal largely surpasses that of a PN code system. Inspired by these advantages, a phase coherent DSFH signal model is developed for military satellite TT&C. In satellite TT&C systems, usually phase coherent signal is transmitted and consistent carrier phase tracking is performed to achieve higher ranging and velocity measuring resolution.

An X/X band spread spectrum TT&C system was developed by Alcatel Alenia Space-Italy. In this system, slow frequency hopping and direct sequence spread spectrum are applied but not combined [2].

Like the PN signal, the acquisition of DSFH signals is a two-dimensional searching process of time and Doppler [3]. In the time dimension and the Doppler dimension, the receiver accomplishes the acquisition process by searching for the correct PN code phase and the correct Doppler and obtaining the maximum correlation. The searching process can be sequential or parallel, depending on the acquisition time requirement and the calculation resources available. Whatever the searching strategy is, we can save acquisition time and calculation resources at the cost of SNR loss.

The main contributions of this work are, we introduce a practical DSFH TT&C signal model based on the research of Cherubini and Milstein [4]. In this signal model fast frequency hopping (FFH) is combined with cryptographic pseudo-noise codes, phase coherence is guaranteed at the transmitter and the receiver performs coherent accumulation among hops to acquire the signal.

This paper is organized as follows. In Section II, we introduced the phase-coherent DSFH TT&C signal. Section III analyzed the auto-correlation traits of DSFH signal and the searching complexity in acquisition. In Section IV the PC correlation is proposed, the digital signal processing of auto-correlation of DSFH signal is described the resulting SNR loss caused by approximations are discussed. Section V concludes the paper.

Here we claim that the PN signal and the DSFH signal discussed later use the same PN code, the frequency hopping bandwidth far exceeds the bandwidth of PN signal

978-1-4673-8266-3/15/$31.00 ©2015 IEEE

and the hopping rate far exceeds the data symbol rate, in other words, fast frequency hopping is discussed.

2. DSFH TT&C Signal Model

The aim of a TT&C signal is to monitor the range, velocity and working status of space vehicles. In a FFH and phase-coherent DSFH signal, the hopping rate is so high that we have to accumulate the energy from dozens of hops at the receiver. The phase continuity between consecutive hops is maintained to guarantee high precision ranging of satellite.

Since the TT&C information bits is constant in an integration period during acquisition, the sending signal can be written as

$$s(t) = \sqrt{2S}\sum_{i=1}^{\infty} \mathrm{Re}\left[s_{l,i}(t) e^{j2\pi f_i (t - iT_h)} \right].$$

$s_{l,i}(t)$ denotes the complex envelope of the IF frequency hopping signal, and

$$s_{l,i}(t) = g(t - iT_h)PN(t)$$

S is the average power of the signal, $PN(t)$ is the PN sequence with a chip rate f_{chip}, f_i is the hopping radio frequency, and T_h is the time period of each hop.

To guarantee integral numbers of carrier cycles in a T_h, the hopping frequency is designed as follows, according to Cherubini and Milstein [3].

$$f_i = \frac{1}{T_h}(A + b), \quad b \in (1, 2, ..., B)$$

b is a number which is picked according a random frequency hopping pattern. For simplicity, we assume the frequency hopping pattern and PN code phase have the same cycle. A is a constant number that decides the RF frequency band of the signal. f_is are chosen from a set of the harmonics of frequency hopping rate R_{hop} to form integral RF carrier cycles in the time period T_h, The frequency hopping bandwidth is B_h=BR_{hop}.

The received signal is modeled as follows

$$r(t) = \sqrt{2S'}\sum_{i=1}^{\infty} \mathrm{Re}\left[s_{l,i}'(t) e^{j\left(2\pi(f_i + f_{d,i})(t-\tau)+\phi\right)} \right] + n(t)$$

$$s_{l,i}'(t) = g(t - iT_h')PN'(t - \tau)$$

$n(t)$ represents the additive white Gaussian noise, S' is the received signal power, τ is the propagation delay which is identical for all hops and $f_{d,i}$ is the Doppler frequency corresponding to f_i. We mark these different Dopplers in N frequency hops as a vector

$$\mathbf{f_d} = \left[f_{d,1}, f_{d,2}, \dots, f_{d,N} \right].$$

The DSFH signal suffers the Doppler Effect so the time length of a hopping period changes as

$$T_h' = T_h\,(1 + f_{d,i}/f_i)\cdot$$

3. Principles of Acquisition

Acquisition is to search through the range where the signal may potentially be and calculate the correlation between the received signal and its local replica. Therefore, the complexity of acquisition is the multiple of the searching complexity and correlation complexity.

3.1. AutoCorrelation traits of DSFH signal

The searching of DSFH signal correlation peak is much more complex than that of pure PN spread spectrum signal, because the correlation peak of DSFH signal is narrower than that of PN signal. Without considering channel impairments, the autocorrelation function of DSFH signal in 1 bit (N hops) is

$$R(\tau, \mathbf{f_d}) \stackrel{def}{=} \left| \int_0^{NT_h} PN(t)PN'(t-\tau) e^{j\left(2\pi(f_i + f_{d,i})(t-\tau)\right)} e^{-j2\pi f_i t} \mathrm{d}t \right| \quad (1)$$

The maximum of $R(\tau,\mathbf{f}_d)$corresponds to the optimized estimate of $(\tau,\mathbf{f}_d)$.

We can rewrite the autocorrelation function as the modulus of the summary of a cascade of complex hop integrals as

$$R(\tau, \mathbf{f_d}) = R_{DS}(\tau, \mathbf{f_d}) \left| \sum_{i=1}^{N} R_i(\tau, f_{d,i}) \right| \quad (2)$$

where

$$R_i(\tau, f_{d,i}) = \int_{(i-1)T_h}^{iT_h} e^{j\left(2\pi(f_i + f_{d,i})(t-\tau)\right)} e^{-j2\pi f_i t} \mathrm{d}t$$

and

$$R_{DS}(\tau, \mathbf{f_d}) = \int_0^{NT_h} PN(t)PN'(t-\tau)\mathrm{d}t\ .$$

For convenience in following discussion, we define

$$\theta_i = \arg R_i(\tau, f_d)$$

as the phase angle corresponding to the complex-valued integration result in the ith hop. We mark the phase angles in N hops as a vector

$$\mathbf{\Theta} = [\theta_1 \quad \theta_2 \quad \dots \quad \theta_N]$$

When the PN code phase error is small enough, say, less than 0.25 chip, the maximum correlation result can be approximated as

$$R(0,0) = R_{DS}(0,0) \left| \sum_{i=1}^{N} R_i(\tau, f_{d,i}) \cdot e^{-j\theta_i(\tau, f_{d,i})} \right|$$

To obtain the maximum of correlation, the relation between phase angles θ_i and $(\tau, f_{d,i})$is described below.

We can assume the phase angle θ_1=0, then the phase angle difference caused by time delay is

$$\Delta\theta_{\tau,i} = \theta_i - 0 = 2\pi f_i \tau \tag{3}$$

When τ increases, the τ-θ_i difference in each hop increases and the complex integral I_i+jQ_i in each hop differs, so the N-hop integrations in both I branch and Q branch cancels out. If τ exceeds the reciprocal of FH bandwidth, the bit integration is neutralized severely.

Under none-Doppler assumption, when the time delay between the incoming signal and the local replica approaches zero, $R(\tau,0)$ reaches the maximum. It is the same for DSSS signal and DSFH signal that when τ increases, $R(\tau,0)$ drops. The auto correlations of PN signal and DSFH signal are shown in Fig.1.

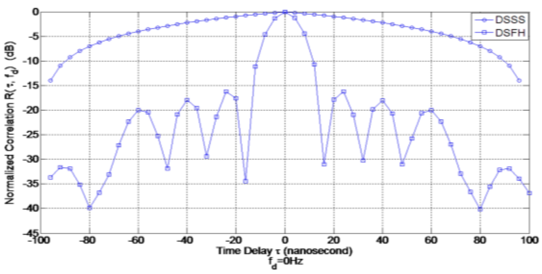

Fig.1 $R(\tau,0)$ of DSSS signal and DSFH signal
DSSS: f_{chip}=10Mcps, 1023 gold sequence RF frequency 2.2GHz, R_b =1ksps
DSFH: f_{chip} =10Mcps,hopping band 2.2~2.3GHz,random hopping.

The $R(\tau,0)$ of DSFH signal is sharper than DSSS signal since τ-θ_i differences in different hops increase rapidly with τ and lower the I,Q branch integral. This is reasonable because the FH bandwidth goes beyond the DSSS bandwidth.

The correlation peak of our DSFH signals respecting f_d is different from that of DSSS signals, because Doppler also neutralize the predetection integral.

With the Doppler Effect, the frequency hopping period T'_h at the receiver changes into

$$T_h^{'} = T_h / (1 + f_{d,1} / f_1)$$

So the phase angle difference caused by Doppler is calculated as

$$\Delta\theta_{f_d,i} = \theta_i - 0 \approx 2\pi f_i \frac{f_{dop,1}}{f_1} (i-1) T_h + 2\pi f_i \tau \tag{4}$$

The comparison of $R(\tau,0)$ between DSSS and DSFH is shown in Fig.2.

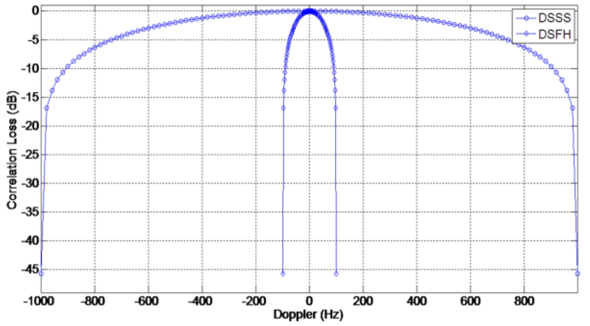

Fig.2 $R(0,f_d)$ of DSSS Signal and DSFH Signal

3.2. Searching Complexity in Acquisition

Acquisition of DSFH signal is to find the maximum match in both time and frequency scopes which may contain the signal. The searching space of DSFH signal is two dimensional if the frequency hopping pattern and the PN code phase can both be ascribed to time uncertaintyτ. From Equation 1, we found that the calculation of autocorrelation function can be transformed into the correlation of PN code multiplying the correlation of frequency hopping waves, so we expands the two dimensional search in a 3D form to search for the correct $\mathbf{\Theta}$.

The searching space and granularity of searching are illustrated in Fig.3. k represents the searching time of Doppler frequency, m_1 represents the PN code phase and m_2 represents the time delay within $T_c/2$ [5]. The searching for Doppler frequency and time delay is designed as below:

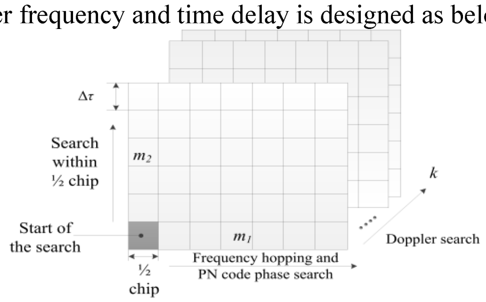

Fig.3 Searching Process of DSFH Signal Acquisition

$$\tau(m_1, m_2) = m_2 \cdot \Delta\tau + m_1 \cdot M_2$$

$$m_1 = 1, 2, ..., M_1, m_2 = 1, 2, ..., M_2$$

$\Delta\tau$ is the time searching resolution we choose. T_c is the reciprocal of PN chip rate and τ_{RG} is the time uncertainty range in acquisition. Obviously we have

$$\tau_{RG} = \Delta\tau M_1 M_2$$

$$M_2 = \frac{T_c}{2\Delta\tau}$$

$$M_1 = \frac{2\tau_{RG}}{T_c}$$

To a sequential searching algorithm, the maximum acquisition time is the multiplication of predetection integration time t_d and the number of searching steps [6]. Whether the searching steps are carried out in a parallel way or sequential way does not affect the acquisition complexity. If the correlation complexity is reduced, the saving in acquisition process will be quite remarkable.

4. Phase-Compensating Acquisition

In this section, we analyze the complexity of the classical acquisition algorithm, propose our PC-based correlation, and calculate the SNR losses caused by approximations in this algorithm.

4.1. Correlation of Classical Acquisition

In a classical acquisition algorithm, the receiver search the uncertainty bins one by one and calculates the correlation. The correlation function can be written in time-discrete form as

$$R(m_1, m_2, k) = R_{DS}(m_1, m_2, k)\left\langle \mathbf{R}, e^{-j\boldsymbol{\theta}} \right\rangle \quad (5)$$

where

$$\left\langle \mathbf{R}, e^{-j\boldsymbol{\theta}} \right\rangle = \sum_{i=1}^{N} R_i(m_1, m_2, k) \cdot e^{-j\theta_i(\tau(m_1, m_2), f_{d,i}(k))} .$$

The maximum of $R(m_1,m_2,k)$ gives the acquisition result. $\mathbf{R}$ and $\boldsymbol{\theta}$ are both column vectors with N elements. $f_{d,i}(k)$ is the Doppler frequency of the ith hop, and

$$\mathbf{f_d}(k) = \left[\frac{f_i}{f_1} f_{d,1}(k), ..., \frac{f_N}{f_1} f_{d,1}(k) \right], k = 1, 2, \cdots, K .$$

The Nyquist sampling rate is f_s, the sampled data is complex signal, that is, with an I branch and a Q branch. So the sampling rate is the bandwidth of the signal plus the PN chip rate.

$$f_s = B_h + f_c$$

The searching step contains the Doppler frequency and time delay, so the times of complex multiplication in calculating the correlation function is

$$K \cdot M_1 \cdot M_2 \cdot t_d \cdot f_s .$$

The times of add is

$$K \cdot M_1 \cdot M_2 \cdot (t_d \cdot f_s - 1) .$$

4.2. Phase-Compensating Correlation

In order to carry out the correlation efficiently, we propose our PC correlation algorithm. We first describe the complexity reduction of our algorithm, then, we give the signal processing structure of our algorithm. Two improvements are made in this algorithm. 1)A low pass (LP) filter is usually applied to lower the speed of signal processing at the cost of an affordable SNR loss, since usually Doppler frequency is far less than the frequency-hopping rate, or we can search the Doppler in multiple branches to reach this scope. 2)the searching for time delay is through phase compensating which operates on each hop, not each sampled data.

Therefore, the times of multiplication is reduced through transforming Equation (3) into the following calculation

$$R(m_1, m_2, k) = R_{DS}(m_1, 0, 0) \cdot C(m_1, m_2, k) \quad (6)$$

where

$$C(m_1, m_2, k) = \left| \sum_{i=1}^{N} R_i(m_1, 0, 0) \cdot e^{-j\theta_i(\tau(m_1, m_2), \mathbf{f_d}(k))} \right| .$$

Taking into account the calculation of $R_1(m_1,0,0)$ and $R_{ds}(m_1,0,0)$, the times of complex multiplication is reduced to

$$K \cdot M_1 \cdot M_2 \cdot (N+1) + M_1 \cdot t_d \cdot f_s ,$$

and the times of adding is changed into

$$K \cdot M_1 \cdot M_2 \cdot (N-1) + M_1 \cdot t_d \cdot f_s .$$

The above analysis shows that using our method, the calculation complexity is reduced at the ratio of

$$\frac{K \cdot M_1 \cdot M_2 \cdot (N+1) + M_1 \cdot t_d \cdot f_s}{K \cdot M_1 \cdot M_2 \cdot t_d \cdot f_s} .$$

To present the digital signal processing implementation of this correlation method, we divide the correlation calculations above into two functional processes: Carrier and Code Stripping (CCS) and Phase Compensating (PC). These two units are shown in Fig. 4 and Fig. 5.

In Fig.4 the DSFH signal is downconverted to digital intermediate frequency (Digital IF), carrier stripping is performed via complex downconverting and then code stripping is performed with a local PN code generator. The sampling rate of the input DSFH signal is Nyquist sampling rate f_s .Both the PN code generator and frequency-hopping pattern are controlled by the acquisition process.

Fig.5 illustrates the bit integration through compensating the $\boldsymbol{\Theta}$ caused by time and Doppler uncertainty. The integrals in different hops are compensated with an array of phase angles which changes with Doppler and time searching values, then the compensated integrals in N hops are summed up. The predetection integral is obtained through the compensated result of PC unit. The correct Doppler and time delay areacquired via deciding the maximum correlation value.

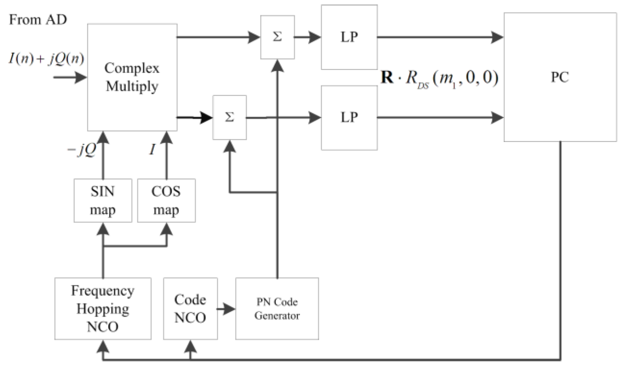

Fig.4 Carrier and Code Strippin

The phase-compensating shows a connection to compressed sensing in that when carrier and code stripping is done, the compensating factor Θ can be seen as a sparse representation of the received DSFH signal.

4.3. SNR Losses

This part concludes the SNR loss caused by the necessary approximation in PC correlation. Applying the conclusions of Kaplan [7] to this signal model, the SNR loss in acquisition can be calculated below.

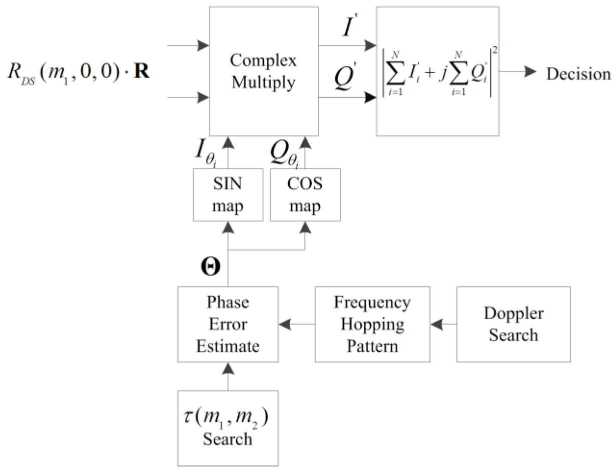

Fig.5 Phase Compensating

1) The SNR loss (in dB) in $R_i(m,0)$ caused by Doppler
approximation is

$$10\lg\left(\operatorname{sinc}^2\left(\frac{f_{d,i}}{R_{hop}}\right)\right)$$

2) The loss caused by Doppler frequency on $R_{ds}(m_1,0,0)$ is calculated as

$$10\lg\left(\int_0^{\frac{f_{d,i}f_{chip}}{f_{RFi}}t_d}(1-u)\mathrm{d}u\right)$$

3) The average SNR loss from $R_{ds}(m_1,m_2,0)$ to $R_{ds}(m_1,0,0)$ is

$$10\lg\left(\int_0^{\frac{1}{2}}(1-u)\mathrm{d}u\right)$$

5. Simulations

Assume the noise is white Gaussian noise with variance σ^2, we apply the Neyman-Pearson criterion to the simulation of acquisition. With the false-alarm probability P_F and detection probability P_D, the threshold V_t is calculated as

$$V_t = \sigma\sqrt{-2\ln P_F}$$

The simulation settings are given in Table 1.

Table 1 Simulation Settings

P_F	0.001
V_t	2.6283
σ^2	1
Doppler (Hz)	0
1 Chip (ns)	10

Monte-Carlo simulation is carried out to verify the detection performance of our algorithm. We adjusts the time searching step $\Delta\tau$ and the E_s/N_0 of sending signal. The probabilities of detection under different searching steps are shown in Fig.6.

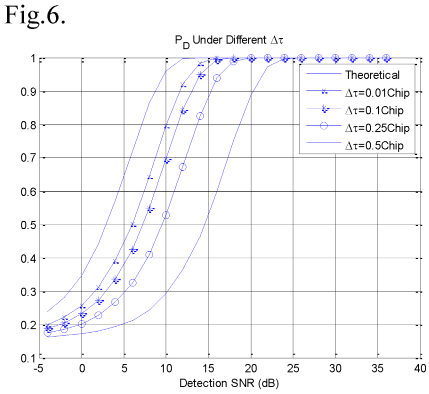

Fig.6 Phase Compensating

The simulation verifies that our algorithm can effectively acquire the DSFH signal and the theoretical P_D curve is 3dB away from our practical curve with $\Delta\tau$=0.01chip due to digital signal processing loss and searching SNR loss. When enlarging the time searching step the SNR loss increases. If the searching step exceeds 0.5chip, the correlation peak vanishes.

6. Conclusion

The DSFH signal possesses a steeper correlation peak than pure PN signal, which makes DSFH signal a promising anti-jamming signal system in space ranging and navigation. Our phase-compensating acquisition algorithm provided a practical scheme for the acquisition of DSFH signal.

References:

[1] CCSDS “Pseodo-Noise Ranging Systems” Draft Recommendation for Space Data System Standards, Aug. 2008

[2] Simone L, Salerno N, Maffei M, “Frequency-hopping Techniques for Secure Satellite TT&C: System Analysis and Tradeoffs,” IEEE Proceedings of IWSSC 2006, Madrid, pp.13-17, 2006.

[3] L1 Jian, CHEN Jie. “An Improved Parallel Acquisition Method for BDS Signals,” Journal of Astronautics,vol.11, pp1299-1305, Nov. 2014

[4] Giovanni Cherubini and Laurence B. Milstein, “Performance Analysis of Both Hybrid and Frequency-Hopped Phase-Coherent Spread- Spectrum Systems Part I: A Hybrid DS/FH System,” IEEE Transactions on Communications, vol. 37, no. 6, pp.600-611, Jun. 1989.

[5] Thales Alenia Space, “Cryptographic Pseudo-Noise Codes and Related Acquisition Techniques for Direct-Sequence Spread Spectrum Transponders-Final Report,” RPT-RFP-ESA-00013-AASI, Tech. Rep. Issue 2, Sep. 2011.

[6] Marco Maffei, Lorenzo Simone and Giovanni Boscagli, “On-Board PN Ranging Acquisition Based on Threshold Comparison with Soft-Quantized Correlators” IEEE Transactions on Aerospace and Electronic Systems, vol. 48, No. 1, Jan. 2012.

[7] Elliot D. Kaplan and Christopher J. Hegarty, Understanding GPS, Principles and applications Second Edition, ARTECH HOUSE, Inc., MA. 2006, chap. 5, pp.153-173.