GPS Signal Joint Acquisition Method of Mean Function and Autocorrelation Function under Multiplicative and Additive Noise

CHAO WU*, LUPING XU, HUA ZHANG
School of Aerospace Science and Technology
Xi'dian University
Xi'an 710126
China
wuchaoid@126.com

Abstract: GPS signal detection method based on high order cyclostationarity (DMHOC) is the state-of-the-art method for acquiring the GPS signal under multiplicative and additive noise. However, since the method only uses autocorrelation function to obtain the peak and estimate the received signal frequency, the peak may be buried in the noise when power spectrum of the multiplicative noise (PSMN) is low. To solve the problem, the GPS signal joint acquisition method of mean function and autocorrelation function under multiplicative and additive noise has been proposed. The proposed method uses the energy ratio threshold to determine which function (mean function or autocorrelation function) to be used for estimation of the received signal frequency. The simulation results show that the proposed method is more robust to the change of multiplicative noise than DMHOC.

Key-Words: GPS acquisition, Cyclostationarity, Multiplicative and additive noise, Mean function, Autocorrelation function, Energy ratio

1 Introduction

Global Positioning System (GPS) signal [1] is the direct sequence spread spectrum modulation signal, and every satellite transmits a particular pseudorandom noise (PRN) code. Satellites are acquired by correlating the received signal with local code signals and comparing the results against a threshold. In practice the local replica of the transmitted code signal differs from the received code signal by a code phase shift and a Doppler shift. Both have to be determined simultaneously in a two-dimensional search [2]. The results of this search, which is usually called acquisition, are required for presetting subsequent stages of the GPS receiver [3].

To improve the acquisition probability of satellite navigation signal, accumulating the received data[4] could be used to increase the signal to noise ratio (SNR). For example, the non-coherent detection method [5] has been proposed, which can reduce the influence of the bit-transition on signal accumulation but make the noise amplitude squared at the time of detection. The coherent detection method [6] has been proposed, which improve the SNR but is sensitive to bit-transition. The differential combination method [7,8] have been proposed, which can suppress the effect of the Doppler offset and bit-transition but the correlation peak was attenuated due to the multiplication of the noisy signal. The models of the method above in received GPS signal are almost assumed to be additive.

However, when GPS signal passes through the ionosphere [9,10], each electron will cause a phase change in the carrier. The carrier phase fluctuation caused by the abundant electrons can be modeled as a multiplicative noise process. Moreover, the comprehensive carrier phase changing effect in a huge number of paths [11,12] can also be modeled as the multiplicative noise. The problem of estimating the frequencies of harmonics under multiplicative and additive noise has been discussed in [13,14]. Due to the feature of cyclostationarity of harmonics, the mean function is called cyclic mean, and autocorrelation function is called cyclic autocorrelation. Many methods[15-17] only use the mean function to estimate the signal frequency in low power spectrum of the multiplicative noise (PSMN). In other words, the signal with PRN code stripped off is processed by frequency domain method, such as Fast Fourier Transformation (FFT), which is not proper for high PSMN [18,19,20].

Since there is little discussion on how to acquire GPS signals degraded by both multiplicative and additive noises, the signal model has been set up in

Ref [20]. However, the part of terms in signal model has been omitted, and the signal model is written as:

$$x(n)=\sum_{k=1}^{K}\frac{A_k C_k(\tilde{\tau}_k)d(n)}{\cos(2\pi f_k\tilde{\tau}_k)w_*(n)}+w_+(n) \quad (1)$$

where A_k is the amplitude of k-th GPS signal. $d(\,)$ is the bit sign. $\tilde{\tau}_k=nT_s-\tau_k$. τ_k is the code phase. T_s is sampling interval. $w_*(n)$ denotes multiplicative noise process. $w_+(n)$ denotes additive Gaussian noise process. The correlation function can be written as:

$$r(n,m)=\sum_{k=1}^{K}\begin{matrix}A_k^2C(\tilde{\tau}_k^+)C(\tilde{\tau}_k^-)\\ d_k(n^+)d_k(n^-)\\ \cos(2\pi f_k\tilde{\tau}_k^+)\\ \cos(2\pi f_k\tilde{\tau}_k^-)\sigma_*^2\delta(m)\end{matrix}+\sigma_+^2\delta(m) \quad (2)$$

where $\tilde{\tau}_k^+=n^+T_s-\tau_k$. $\tilde{\tau}_k^-=n^-T_s-\tau_k$. $n^+=n+\frac{m}{2}$. $n^-=n-\frac{m}{2}$. If the multiplicative noise is large enough, the correlation result is related to time nT_s. Therefore, under the condition of large multiplicative noise, the noise processes of the received GPS signal cannot be modelled as wide sense stationary and the conventional GPS signal acquisition method based on correlation may not always work. Based on the signal model, GPS signal detection method based on high order cyclostationarity (DMHOC) has been proposed. However, the method only uses the autocorrelation function to acquire the GPS signal. This may not acquire the signal in low PSMN, which is analyzed in the following discussion.

To acquire the GPS signal under the multiplicative and additive noise, this paper has proposed the GPS signal joint acquisition method of mean function and autocorrelation function. Firstly, the received GPS is proved to be cyclostationary. Due to the property, the Fourier series coefficient of mean function and autocorrelation function can be used to estimate the frequency and code phase by Fourier transformation (FT) process. Since the proposed method can choose the detection scheme based on mean function or autocorrelation function by the energy ratio threshold, the proposed method is more robust to the change of multiplicative noise than DMHOC. The simulation results shows that the proposed method can have better detection performance than DMHOC under low PSMN circumstance.

The rest of this paper is organized as follows: In Section 2 the received signal model is given and proved to be cyclostationary. In Section 3, the conventional method (GPS signal detection method based on high order cyclostationarity) has been analyzed. Based on the analysis, the paper proposes the GPS signal joint acquisition method of mean function and autocorrelation function in Section 4. The simulation results show the effectiveness of the proposed method compared with the DMHOC in Section 5. Finally, the conclusions are drawn in Section 6.

2 Signal Model

The received signal from a specific satellite vehicle (SV) can be written as

$$S(t)=\frac{C(t-\tau)d(t)\left(A+w_*(t)\right)}{\cos\left(2\pi(f_{IF}+f_d)t\right)+w_+(t)} \quad (3)$$

where τ is the code phase delay. f_d is Doppler frequency. A is the signal amplitude. $c(\,)$ is the pseudo random noise (PRN) code. $w_*(\,)$ is the multiplicative noise. $w_+(\,)$ is the additive noise. $d(\,)$ is the bit sign.

It is assumed that T_0 is the period of received signal $S(t)$.

$$\begin{aligned}E\left(S(t+T_0)\right)&=C(t_0-\tau)d(t_0)A\cos(2\pi ft_0)\\&=E\left(S(t)\right)\end{aligned} \quad (4)$$

$$\begin{aligned}&R_S\left(t+T_0,t_\tau\right)\\&=E\left(S(t+T_0+t_\tau)S(t+T_0)\right)\\&=\frac{\left(A^2+\sigma_*^2\delta\left(t_\tau\right)\right)\Gamma}{\cos(2\pi ft_1)\cos(2\pi ft_0)+\sigma_+^2\delta\left(t_\tau\right)}\\&=E\left(S(t+t_\tau)S(t)\right)=R_S\left(t,t_\tau\right)\end{aligned} \quad (5)$$

where $\Gamma=C(t_1-\tau)C(t_0-\tau)d(t_1)d(t_0)$. $f=f_{IF}+f_d$. $t_0=t+T_0$. $t_1=t+T_0+t_\tau$. $E\left(S(t)\right)$ is the mean function of $S(t)$. $R_S\left(t,t_\tau\right)$ is the autocorrelation function of $S(t)$. σ_*^2 and σ_+^2 are multiplicative noise variance and additive noise, respectively. As the equations are shown, both mean function and autocorrelation function are periodic. So it proves that received signal is cyclostationary [18].

3 GPS signal detection method based on high order cyclostationarity (DMHOC)

GPS signal detection method based on high order cyclostationarity (DMHOC) can be drawn as follows [20]:

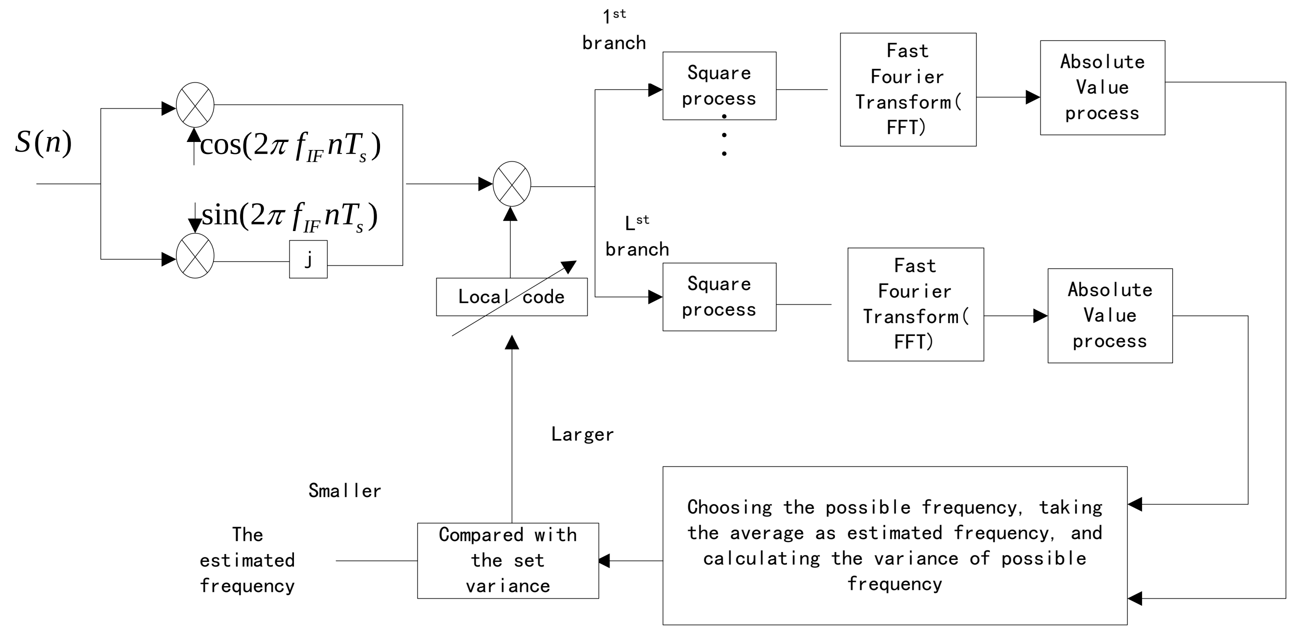


Fig 1 The block diagram of DMHOC

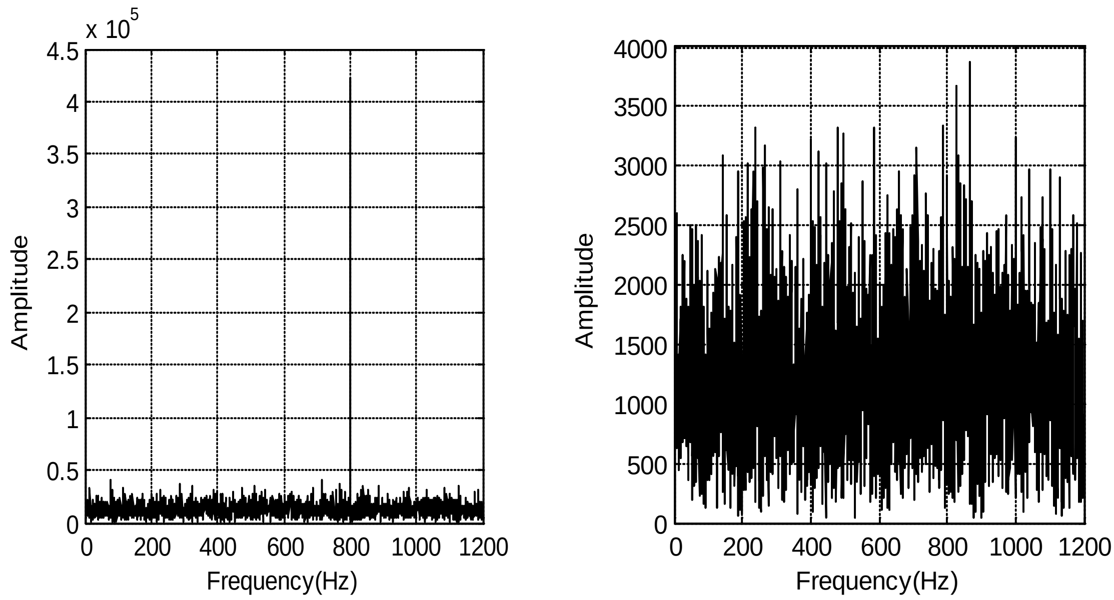
Based on the block diagram of DMHOC, the signal through the square process can be written as:

$$S_s^i(n) = \left\{ \begin{matrix} C(n-\tau')C(n-\tau)d(n)(A+\omega_*(n)) \\ \exp(j2\pi f_d nT_d)+W_+(n) \end{matrix} \right\}^2$$
$$= (A+\omega_*(n))^2 \exp(j4\pi f_d nT_d) + (W_+(n))^2 + 2W_+(n)C(n-\tau')C(n-\tau)d(n)(A+\omega_*(n))\exp(j2\pi f_d nT_d) \quad (6)$$

where $S(n)$ in Fig 1 represents the digital signal of $S(t)$ at sampling interval T_s. $i=1,...,L$. τ' is the local code phase. τ is the received code phase. $d(n)$ is the data bit of received signal. Both real part and imaginary part of $W_+(n)$ are additive Gaussian noise with zero mean and variance σ_+^2, $\omega_*(n)$ is the multiplicative noise with mean 0 and variance σ_*^2. Since the mean function of $S_s^i(n)$ can be written as:

$$E\left(S_s^i(n)\right) = \left(A^2+\sigma_*^2\right)\exp(j4\pi f_d nT_d) + 2\sigma_+^2 \quad (7)$$

Due to the different frequencies, we can distinguish the signal from additive noise and estimate the signal frequency by FFT process. This is ideal situation. As Fig 1 shows, in practice $S_s^i(n)$ instead of $E\left(S_s^i(n)\right)$ is performed on FFT process. Since the square process is performed on the received signal, the estimated frequency corresponding to the peak is 2 times than the frequency f_d. When and $\sigma_*^2 + A^2$ is small (i.e PSMN is low) [20], the peak may be buried in the additive noise. The Fig 2 below shows the amplitude under different PSMN when signal power is -160dBW and the power spectrum of the additive noise (PSAN) is -180 dBm/Hz:



(a) PSMN=-170 dBm/Hz (b) PSMN=-200 dBm/Hz

Fig 2 The Amplitude of FFT of $S_s^i(n)$ when f_d =400Hz

Comparing Fig 2 (a) with Fig 2 (b), the obvious peak disappears in Fig 2 (b). This is due to the fact that PSMN decreases, the peak is buried in the additive noise, and the correct peak is hard to detect. This will lead to the increase of variance, and the performance of estimated frequency will be degraded. This will be proved in Section 5.

4 The proposed method

Since DMHOC only uses autocorrelation function to estimate the frequency and is hard to detect the signal under low PSMN circumstance, this paper proposes the joint acquisition method of mean function and autocorrelation function, and uses the energy ratio threshold to determine which function (mean function or autocorrelation function) to be used for estimation of the received signal frequency.

To get rid of the disturbance caused by two periods (the period of PRN code and the period of the frequency), the period of PRN code should be removed. The generalized Fourier series coefficient of the mean function $E\left(S(t)C(t_0-\tau)\right)$ and autocorrelation function $R_S\left(t,0\right)$ can be written as, respectively:

$$
\begin{aligned}
F_E\left(\alpha\right) &= \lim_{N\to\infty}\frac{1}{N}\sum_{t=0}^{N-1} E\left(S(t)C(t-\tau)\right)\exp(-j\alpha t) \\
&= \frac{D}{2}\left(\delta(\alpha-\omega_0)+\delta(\alpha+\omega_0)\right)
\end{aligned}
\tag{8}
$$

$$
\begin{aligned}
F_R\left(\alpha\right) &= \lim_{N\to\infty}\frac{1}{N}\sum_{t=0}^{N-1} R_S\left(t,0\right)\exp(-j\alpha t) \\
&= \frac{A^2+\sigma_*^2}{4}\begin{pmatrix}\delta(\alpha-2\omega_0)+\\ \delta(\alpha+2\omega_0)\end{pmatrix}+\left(\frac{A^2+\sigma_*^2}{2}+\sigma_+^2\right)\delta(\alpha)
\end{aligned}
\tag{9}
$$

where it is assumed that D is constant regardless of the influence of data period on the coefficients. $\omega_0 = 2\pi f_d T$. T is the sampling interval. $F_E\left(\alpha\right)$ and $F_R\left(\alpha\right)$ can be calculated by Fourier transform (FT), and can be used to estimate the code phase and frequency of the received signal. The specific steps of the proposed method can be described as follows:

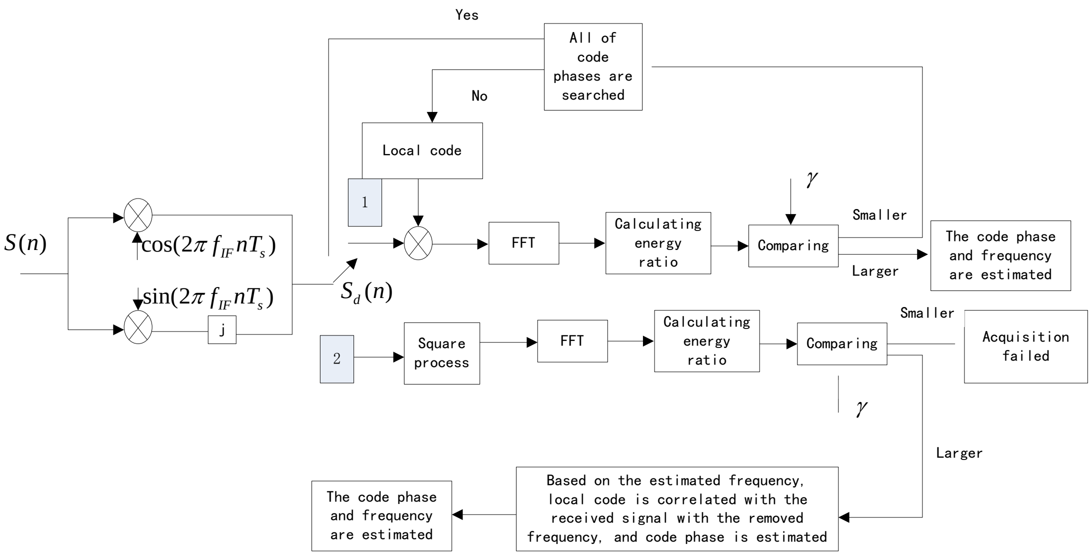


Fig 3 The block diagram of the proposed method

Firstly, the $S(t)$ is converted to the digital signal $S(n)$:

$$S(n) = \begin{matrix} C(nT_s - \tau)d(nT_s)\left(A + w_*(n)\right) \\ \cos(2\pi(f_{IF} + f_d)(nT_s - \tau)) + w_+(n) \end{matrix} \quad (10)$$

Then, multiplied by two orthogonal reference sinusoids, $S(n)$ can be converted to the signal $S_d(n)$:

$$S_d(n) = \begin{matrix} C(nT_s - \tau)d(nT_s)\left(A + w_*(n)\right) \\ \exp(2\pi j f_d nT_s) + W_+(n) \end{matrix} \quad (11)$$

There are two steps to process the $S_d(n)$:

Step1: $S_d(n)$ is correlated with the local code. Since the PRN code has been removed, it is assumed that the sampling interval is $T_d\,(T_d > T_s)$. Then the signal goes through the fast Fourier transform (FFT):

$$S_{F1}(n) = \mathrm{FFT}\left(S_d(n)C(n - \tau_c)\right) \quad (12)$$

where τ_c is the local code phase. Calculating the energy ratio E_{r1}:

$$E_{r1} = \frac{\left(\left|S_{F1}(\tilde{n}1)\right|^2 - \frac{1}{N-1}\sum_{\substack{n=0 \\ n \neq \tilde{n}1}}^{N} \left|S_{F1}(n)\right|^2 \right)}{\frac{1}{N-1}\sum_{\substack{n=0 \\ n \neq \tilde{n}1}}^{N} \left|S_{F1}(n)\right|^2} \quad (13)$$

where $\tilde{n}1$ is the index corresponding to the peak of $\left|S_{F1}(\tilde{n}1)\right|^2$. Comparing the E_{r1} with the set threshold γ. If E_{r1} is larger than the γ, the signal is acquired. If not, the code phase τ_c needs to be updated.

Step2: if all of code phases are searched, the energy ratio of step 1 is still smaller than the threshold. $S_d(n)$ goes through the square process and then is performed on FFT:

$$S_{F2}(n) = \mathrm{FFT}\left(\left(S_d(n)\right)^2\right) \quad (14)$$

Calculating the energy ratio:

$$E_{r2} = \frac{\left(\left|S_{F2}(\tilde{n}2)\right|^2 - \frac{1}{N-1}\sum_{\substack{n=0 \\ n \neq \tilde{n}2}}^{N} \left|S_{F2}(n)\right|^2 \right)}{\frac{1}{N-1}\sum_{\substack{n=0 \\ n \neq \tilde{n}2}}^{N} \left|S_{F2}(n)\right|^2} \quad (15)$$

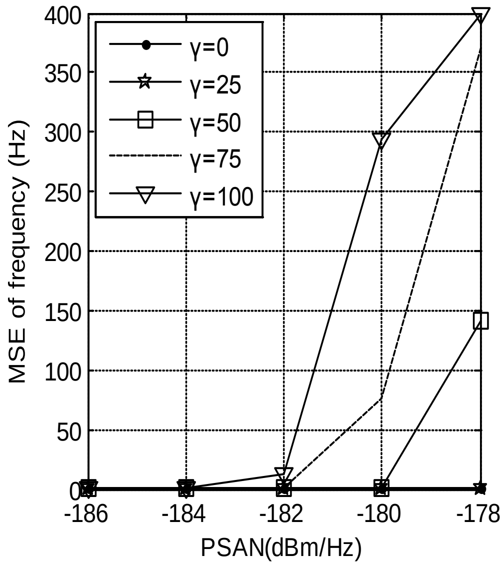
where $\tilde{n}2$ is the index corresponding to the peak of $\left|S_{F2}(\tilde{n}2)\right|^2$. Comparing the E_{r2} with the set threshold γ. If E_{r2} is larger than the γ, the signal is acquired. The estimated frequency can be calculated by $\tilde{n}2$. Then, local code is correlated with the received signal with the removed frequency:

$$S_C(n)=\mathrm{IFFT}\begin{pmatrix}\mathrm{FFT}\left(S_d(n)\exp\left(-2\pi j f_{\tilde{n}2} n T_s\right)\right)\\ \left(\mathrm{FFT}\left(C(n)\right)\right)^*\end{pmatrix} \quad (16)$$

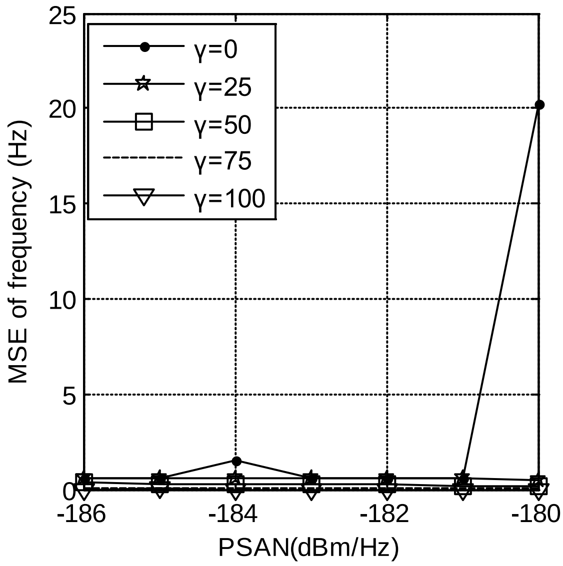
where $f_{\tilde{n}2}$ is the estimated frequency of $S_d(n)$. $(\ )^*$ represents the conjugation operation. $\mathrm{IFFT}(\ )$ is inverse fast Fourier transform (IFFT). The code phase can be estimated by comparing the energy ratio of the peak of $S_C(n)$ with the set threshold β. If the energy ratio is larger than β, the code phase is successfully estimated. Based on the analysis above, the frequency and code phase can be estimated. To evaluate the performance of the proposed method, the mean square error (MSE) of the frequency and code phase is calculated:

$$MSE_T=\sqrt{\frac{1}{N}\sum_{i=1}^{N}\left(\tilde{T}_i-T\right)^2} \quad (17)$$

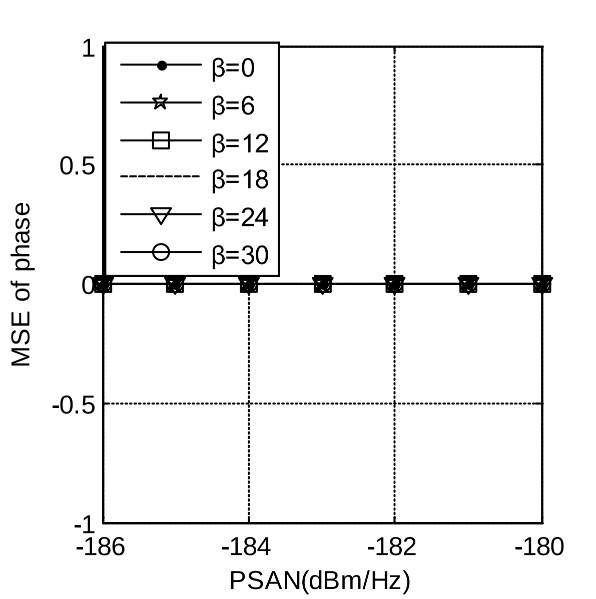
where $T=f$ or C. $T=f$ represents the MSE of frequency. $T=C$ represents the MSE of code phase. $\tilde{T}_i$ represents the i-th estimation of T.

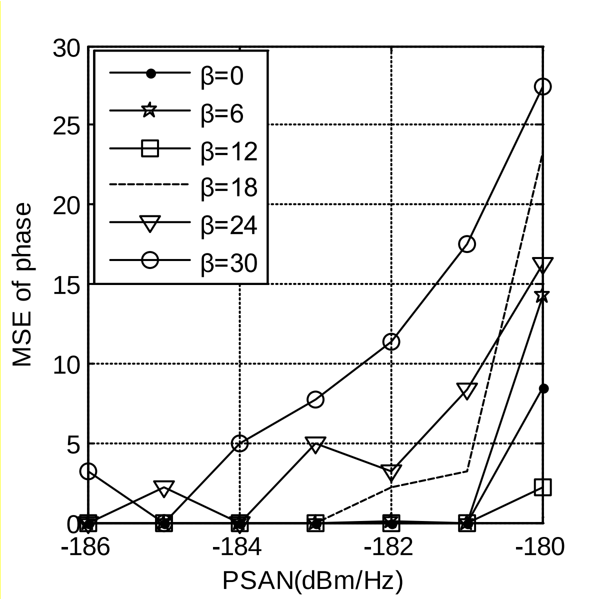


(a) PSMN=-200dBm/Hz, it is assumed that the local code is synchronized with received signal code



(b) PSMN=-177dBm/Hz, it is assumed that the local code is synchronized with received signal code





(c) PSMN=-200dBm/Hz, it is assumed that the received signal frequency is known

(d) PSMN=-180dBm/Hz, it is assumed that the received signal frequency is known

Fig 4 MSE when the signal length is 1ms and N =2000

To avoid the interaction effects between β and γ, we assume that the received signal frequency is known when MSE of phase is calculated, and assume that the local code is synchronized with received signal code when MSE of frequency is calculated.

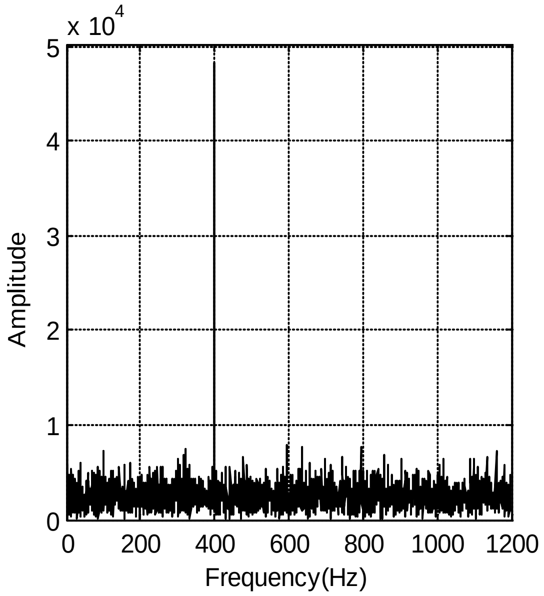
The Fig 4 shows the MSEs of frequency and code phase with different γ values and β values, respectively. The parameter corresponding to minimum of MSE should be selected. So $\gamma = 25$, and $\beta = 6$.

5 Simulation Results

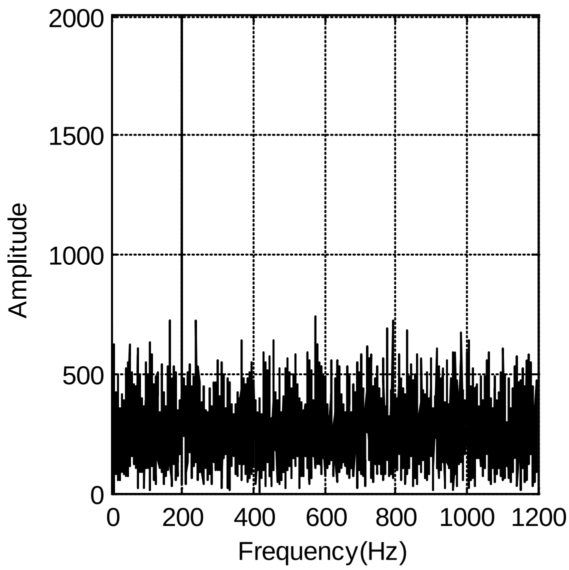
To prove that the proposed method is more robust to the change of multiplicative noise than DMHOC, the simulation results have been made. The parameters are set as follows:

Table 1 Simulation parameters.

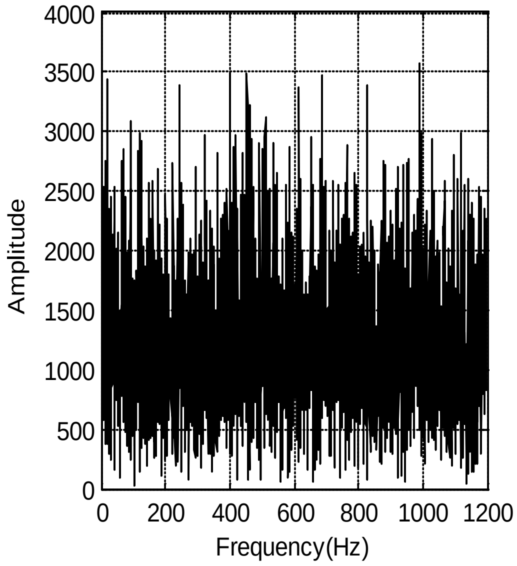
Frequency f_d	200Hz
The signal length	1ms
The signal power	-160dBw
The threshold γ	25
The threshold β	6
Sampling frequency $1/T_d$	1200Hz
The PRN rate	1.023Mc/s
The variance of DMHOC	0.3
The number of branches of DMHOC L	100
The number of Monte Carlo simulations	2000



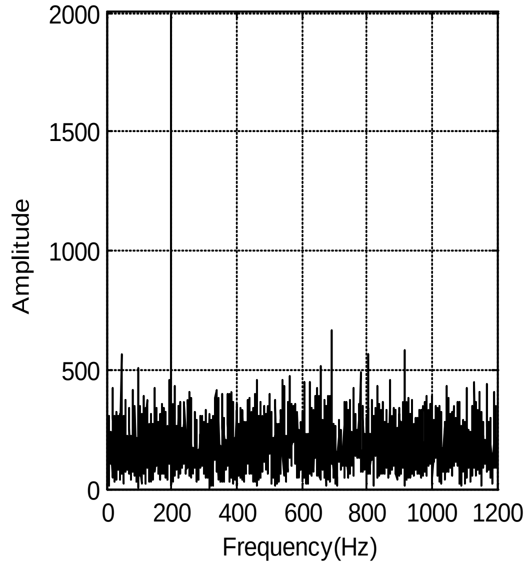
(a)DMHOC when PSMN=-180dBm/Hz,



(b)The proposed method when PSMN=-180dBm/Hz

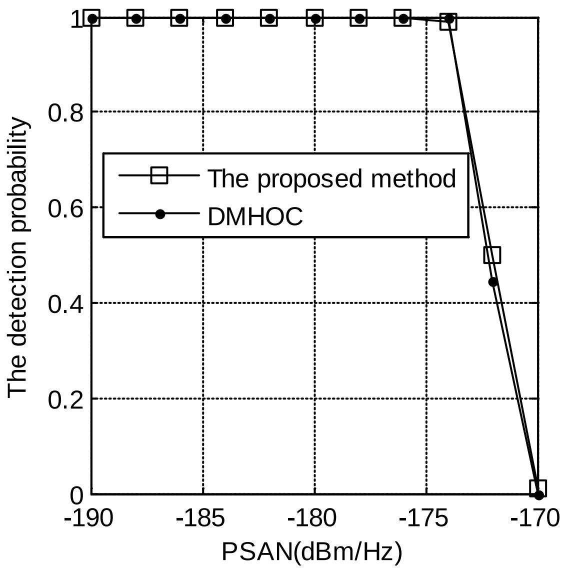


(c)DMHOC when PSMN=-200dBm/Hz,

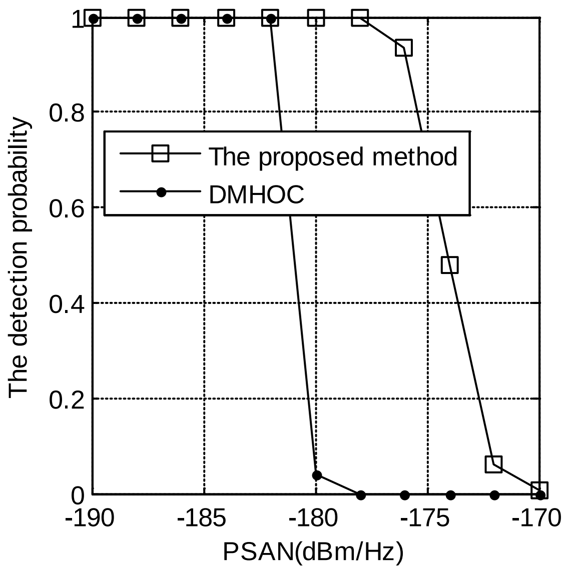


(d)The proposed method when PSMN=-200dBm/Hz,

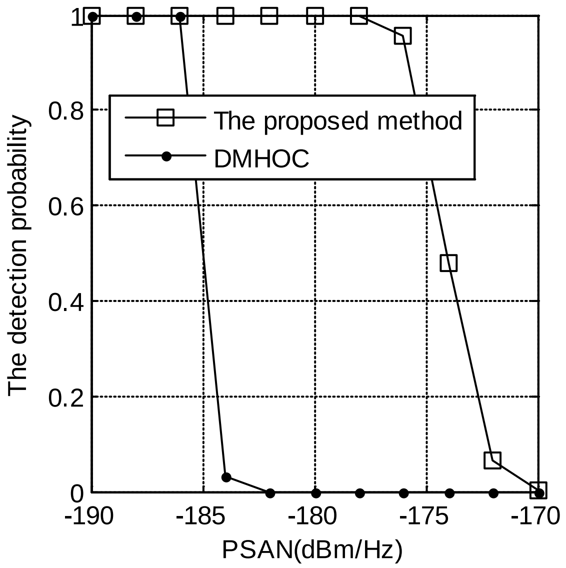
Fig 4 The amplitude for frequency estimation obtained by two methods when PSAN=-180dBm/Hz



(a) PSMN=-180dBm/Hz



(b) PSMN=-190dBm/Hz



(c) PSMN=-200dBm/Hz

Fig 5 The detection probability comparison

Fig 4(a) and Fig 4(c) show the one branch signal amplitude through FFT process of DMHOC. Fig 4(b) and Fig 4(d) shows the signal amplitude through FFT process of the proposed method when the energy ratio is larger than or equal to the set threshold. Comparing Fig 4(a) and Fig 4(b), the obvious peaks can be seen from both figures under the same noise condition (PSMN=-180dBm/Hz and PSAN=-180dBm/Hz). However, the peak is buried in the noise in Fig 4(c) when PSMN=-200dBm/Hz and PSAN=-180dBm/Hz, and clear peak still can be seen in Fig 4(d). This is because DMHOC only uses autocorrelation function to obtain the peak and correct peak may be buried in low PSMN. So the variance calculated by peak detection increases in low PSMN, and the detection probability of frequency of DMHOC will decrease.

To further illustrate the fact, the detection probability comparison is made in Fig 5 by Monte Carlo simulation. The detection probability is the probability that the method estimates the correct frequency unit. As the Fig 5 (a) shows, the detection curve of proposed method almost overlaps with the detection curve of DMHOC. However, when PSMN decreases, the detection performance of DMHOC degrades badly compared with the detection performance of the proposed method in Fig 5(b) or Fig 5(c). So it proves that the proposed method is more robust to the change of multiplicative noise than DMHOC.

6 Conclusion

For GPS signal acquisition under the multiplicative and additive noise, this paper has proposed the joint acquisition method based on mean function and autocorrelation function. The signal model is set up. The mean function and autocorrelation function is utilized to prove that the received signal is cyclostationary. Based on cyclostationary feature, the energy ratio is calculated by the Fourier series coefficient of the mean function or autocorrelation function, and the proposed method use the energy ratio threshold to determine which function (mean function or autocorrelation function) to be used for estimation of the acquisition parameters (including code phase and frequency). The simulation results show that the proposed method has better detection performance than DMHOC under low PSMN circumstance. In other word, the proposed method is more robust to the change of multiplicative noise than DMHOC.

We demonstrated that the proposed method has better frequency estimation performance than DMHOC. The proposed method is applied in high dynamic environment[15]. This is a topic of our further research.

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