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Abstract—Due to a wide range of Doppler frequency shift and low SNR existing in the AIS transmission, it is difficult to decode correctly in Satellite-based AIS receiver. In order to solve the problem, a new design of PLL is proposed based on the theory of conventional PLL. The new design is aided by ML estimating. In the algorithm, frequency offsets are reduced to the fast capture zone after compensations to received AIS signals and then the phase locked loop start to track surplus frequency offsets. The simulation results indicate that the algorithm requiring the symbol number of fast capture is less than 256(the number of AIS frame symbols) when the SNR is 1dB and tracking precision is within 2 Hz. And it can be used under conditions of low SNR with large frequency offsets.

Index Terms—Satellite-based AIS; Doppler frequency shifts; low SNR; phase locked loop

I. INTRODUCTION

Satellite-based Automatic Identification System (AIS) receives terminal signals through the satellite and feedbacks the information to monitoring centers of the shore to achieve the purpose of monitoring ships in the large range. However, influenced by the distance and electromagnetic interference, electromagnetic wave decay is larger and the Signal Noise Ratio (SNR) of the received AIS signal is low on the one hand, while on the other hand, great range of Doppler frequency shifts caused by the satellite motion has a significant impact on the AIS signals whose bandwidth are only a few thousand hertz (Hz). Because of that Doppler frequency shift of large range can easily lead to error decoding of the received signals in very low SNR, so frequency offsets must be compensated before decoding.

AIS signals are modulated by GMSK and the wireless channel can be approximated to Gauss white noise

(AWGN) channel from berths to the satellite. For frequency offsets estimation, there are many estimation methods^[1-4] such as blind estimation^[1], data aided estimation[2] and so on. The research of blind estimation is less because its obvious SNR threshold and estimation precision is low under the low SNR environment. The common algorithms are Kay algorithm and Kim algorithm [5]. As to data aided estimation, there are a lot of researches. In 1985, Steven A.Tretter proposed a approximate estimate algorithm [6] to change the frequency offsets estimation problem into frequency offsets estimation based on complex sequences phase, however it has the SNR threshold and it can only be applied to the high SNR conditions. In 1989, Kay raised a difference phase estimation method [7] on the basis of Steven A.Tretter, but the estimation accuracy is strongly influenced by noise and the algorithm is no longer applicable under the condition of low SNR with phase hopping. In 1995, M.Lulse and R.Reggiannini present a L&R algorithm [8] whose mean square error is close to Cramer Rao Bound (CRB) on Maximum Likelihood (ML) estimation basis, but its estimation range is limited. In 2007, Hua Fu put forward the ML algorithm [9-10] which joint amplitude and phase information. Because this algorithm can not only get the frequency estimated value, but also obtain the phase estimated value, it is the ideal algorithm and mean square error of frequency offset estimation is closed to CRB. However, fewer pilot symbols will lead to the estimation accuracy declining. When the Phase Locked Loop (PLL) is used for phase tracking process, large Doppler frequency shifts will reduce the capture speed and phase locked loop can not be finished capturing within a limited time. With the frequency lock loop[11] help, phase locked loop has a faster capture speed, but it is not suitable in the low SNR because the frequency discriminator requires that input SNR cannot be less than 7dB.

In order to solve the above problems, characteristic sequences (training sequences, start flag and end flag) in AIS frame are viewed as a pilot in this paper. The preliminary estimation of Doppler frequency offsets is

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completed through ML algorithm and phase is tracked by the PLL. This method can satisfy the requirements of tracking precision and capture time in the low SNR circumstance simultaneously.

II. ANALYSIS OF PHASE LOCKED LOOP PARAMETERS

In engineering applications, the PLL is almost working on time synchronization state and loop designs meet the requirement of tracking precision, however, the noise and interference will increase the loop capture difficulties, reduce the tracking performance and make the loop output phase produce random jitter. In the following, we start to analyze the influence of noise on the circuit performance.



Figure 1. Basic structure of phase locked loop

PLL which is composed by phase discriminator (PD), a loop filter (LF) and a voltage controlled oscillator (VCO) is a negative feedback control system and its principle is shown in Fig.1 where LF is a active proportional integral filter and its parameters[12] are as follows

The transfer function of LF is

$$F(s) = \frac{1 + s\tau_2}{s\tau_1} \tag{1}$$

The system transfer function is

$$H(s) = \frac{K(1+s\tau_2)}{s^2\tau_1 + sK\tau_2 + K} = \frac{2\xi w_n s + w_n^2}{s^2 + 2\xi w_n s + w_n^2}$$
(2)

where, *K* is the loop gain, w_n is the loop natural frequency, ξ is the loop damping coefficient and it meets the optimal value based on Wiener theory [13]. They are related to circuit parameters τ_1 and τ_2

$$w_n = \sqrt{K/\tau_1} \tag{3}$$

$$\xi = (\tau_2/2)\sqrt{K/\tau_1} \tag{4}$$

The digital domain transfer function $\frac{1}{1-z^{-1}}$ corresponds to analog integrator transfer function $\frac{1}{s}$ with

the impulse invariance. Taking into account that feedback operational results will be produced at least one symbol time delay when the digital loop hardware is implemented, the system transfer function of the NCO

is $\frac{z^{-1}}{1-z^{-1}}$. The digital phase-locked loop structure is as shown in Fig.2.



2. Digital phase-locked loop structure

The mathematical expression of digital loop is

$$H(z) = \frac{K_0 K_d (c_1 + c_2) z^{-1} - K_0 K_d c_1 z^{-2}}{1 + \left[K_0 K_d (c_1 + c_2) - 2 \right] z^{-1} + (1 - K_0 K_d c_1) z^{-2}} (5)$$

When $w_n T_b > 1$, parameter sets of the impulse response is easy to shock while the bilinear transformation can still work stably. Therefore, bilinear transformation is used to achieve conversion from analog phase locked loop $(2/T_b)(1-z^{-1})$

domain to the digital domain. When
$$s = \frac{(1 + b)((1 + b))}{1 + z^{-1}}$$
,

the result of equation (2) is

$$H(z) = \frac{4\xi w_n T_b + w_n^2 T_b^2 + 2w_n^2 T_b^2 z^{-1} + (w_n^2 T_b^2 - 4\xi w_n T_b) z^{-2}}{(4 + 4\xi w_n T_b + w_n^2 T_b^2) + (2w_n^2 T_b^2 - 8) z^{-1} + (4 - 4\xi w_n T_b + w_n^2 T_b^2) z^{-2}}$$
(6)

Considering the result of (7) is approximately equal to (8), the loop filter parameters are:

$$C_{1} = \frac{1}{K_{o}K_{d}} \frac{8\xi w_{n}T_{b}}{4 + 4\xi w_{n}T_{b} + w_{n}^{2}T_{b}^{2}}$$
(7)

$$C_{2} = \frac{1}{K_{o}K_{d}} \frac{4w_{n}^{2}T_{b}^{2}}{4 + 4\xi w_{n}T_{b} + w_{n}^{2}T_{b}^{2}}$$
(8)

where, K_d is the phase detector gain, K_o is the numerically controlled oscillator gain, T_b is the symbol period,

Phase locked loop for noise suppression ability can be reflected by loop SNR and the definition is

$$\left(\frac{S}{N}\right)_{L} = \left(\frac{S}{N}\right)_{i}\frac{B_{i}}{B_{L}}$$
(9)

in (3), $(S/N)_i$ is the input SNR, $(S/N)_L$ is loop SNR, B_i is the bandwidth of input signals, B_L is the bandwidth of loop noise.

The bandwidth of loop noise B_L relates to the loop damping coefficient ξ and the natural frequency w_n

$$B_{L} = \frac{w_{n}(1+4\xi^{2})}{8\xi}$$
(10)

For a given target, the corresponding loop SNR can be calculated and the upper limit of loop natural frequency can be inferred in the low SNR from (9) and (10):

$$w_n \le \frac{8\xi}{1+4\xi^2} \left(\frac{S}{N}\right)_L^{-1} \left(\frac{S}{N}\right)_i B_i \tag{11}$$

In order to make PLL work stably, loop SNR $(S/N)_L$ must be more than 13dB from reference [13]. Fast capture zone [12] of the ideal two order loop is

$$\Delta w_L = 2\xi w_n \tag{12}$$

When input frequency offsets Δw_o greater than fast capture zone Δw_L , the loop proceeds to capture frequency first and then capture phase, that is to say, the loop reduces frequency offsets to fast capture zones through cycle-skipping and then captures phase quickly. The loop capturing time [14] is

$$t_p \approx \Delta w_o^2 / 2\xi w_n^3 \tag{13}$$

III. DESIGN OF ML AIDED PHASE LOCKED LOOP FOR GMSK

A. Matched Filter for GMSK Signals

Composed of AIS frame structure, binary message sequences are encoded by NRZI and are modulated by GMSK. After that the corresponding complex envelope signal [15] can be expressed as $s(t) = e^{j\Psi(t)}$

where,

$$\Psi(t) = 2\pi\varepsilon \sum_{i} a_{i}q(t-iT_{b})$$

$$q(t) = \int_{-\infty}^{t} g(\tau)d\tau$$

$$g(t) = \frac{1}{2T_{b}} \left\{ Q \left[\frac{2\pi B}{\sqrt{\ln 2}} \left(t - \frac{L+1}{2}T_{b} \right) \right] \right\}$$

$$-\frac{1}{2T_{b}} \left\{ Q \left[\frac{2\pi B}{\sqrt{\ln 2}} \left(t + \frac{L-1}{2}T_{b} \right) \right] \right\}$$

$$Q(t) = \frac{1}{\sqrt{2\pi}} \int_{t}^{\infty} e^{-\tau^{2}/2} d\tau$$

 $\Psi(t)$ is the information-bearing phase, T_b is the symbol period, $a \triangleq \{a_i\}$ are the data symbols 1 or -1 in this case, and ε determines the rate of change in the symbol interval and its value is 0.5 for GMSK, g(t) is the frequency response which is equal to the response of a rectangular pulse with unit amplitude, q(t) is phase response, L is an integer number called correlation length and indicates the symbol duration of the phase transition (ISI).

GMSK (continuous phase modulation) signals whose correlation length is L can be divided into 2^{L-1} digital PAM signals with Laurent algorithm and its mathematical expression [16-17] is as follows

$$s(t) = \sum_{k=0}^{Q-1} \sum_{n} b_{k,n} c_k \left(t - nT_b \right)$$
(15)

However, when GMSK signals are decomposed, most of the energy is concentrated in the main pulse $c_0(t)$, thus, ignoring other pulse $c_k(t)$ $k = 1 \dots 2^{L-1} - 1$, its expression is

$$s(t) \approx \sum_{n} b_{0,n} c_0 \left(t - nT_b \right) \tag{16}$$

where,

(14)

$$b_{0,n} = e^{\left\{j\pi\varepsilon\sum_{m=0}^{a_m}a_m\right\}}$$
$$c_0(t) = \prod_{i=0}^{L-1} \lambda(t+iT_b)$$

n)

For AIS, $BT_{h} = 0.4$, L = 3 is considered.

AIS signals received by the satellite are with low SNR. In order to filter out noise, improve SNR and obtain large gains to the greatest degree, signal frequency offset estimation process need to go through the matched filter. The principle of matched filter is introduced in the following.

Matched filter which filter characteristics are consistent with signal characteristics is a linear filter. The ratio of filter output between signal power and the noise average power is the maximum and the transfer function is equal to signal spectrum conjugate. Based on matched filter features and GMSK signal characteristics, only $c_0(t)$ is taken as a matched filter impulse response for GMSK signals, $c_0(t)$ waveforms is shown as follows.



Figure 3. Matched filter impulse response

B. ML Aided Capture Frequency Offset Tracking Loop

In order to capture quickly under large frequency offsets, the frequency offsets must be reduced to the fast capture zone with the help of ML [10] (when the distance between pilot signals vector and received signals vector is the minimum value, the frequency offset of pilot signals vector is viewed as the estimation result) and then start the phase locked loop to track frequency. The structure diagram is shown in Fig.4.



8bit	24bit	8bit	168bit	16bit	8bit	24bit
Prefix code	Training sequence	Start flag	Data	FCS	End flag	Buffer

Figure 5. AIS frame structure

Frequency offsets estimated by ML is completed with the help of characteristic sequences in AIS signals, where the AIS frame structure [18] is shown in Fig.5. According to the actual requirements of frequency offset estimation precision, the characteristic sequence selected as pilot signals can be the training sequences, start flag, end flag and their combination forms. The end flag is not continuous with the training sequences and the start flag. They are all dispersed (interval) in the AIS frame and the information symbols separated is usually 184 symbols.

Removed the carrier, the received AIS signal vector (assuming the time synchronization) can be expressed as

$$r = v + w = [s * h]e^{j2\pi f_d n T_b} + w$$
(17)

where,

$$r = (r_1, r_2, \dots, r_N)^T$$
$$v = (v_1, v_2, \dots, v_N)^T$$
$$s = (s_1, s_2, \dots, s_N)^T$$
$$h = (h_0, h_1, \dots, h_L)^T$$
$$w = (w_1, w_2, \dots, w_N)^T$$

N is the number of received symbols, *s* is the GMSK signals with E_b being the energy per bit, Doppler frequency offset is f_d which range from (-3.8)kHz to (+3.8)kHz, *h* is the system impulse response, φ_0 is signal initial phase, *w* is the complex Gauss white noise whose expected value is $\mu = 0$ and power spectrum density is $N_a/2$.

Receivers need to decide which sending sequence is with a minimum probability of error, that is to say, it needs to calculate conditional probability $P[v_i/r]$ called posteriori probability for each possible sequences and the sequence with maximum probability is considered to be the sending sequence. By Bayesian equation

$$P[v_i/r] = \frac{f_R[r/v_i]P[v_i]}{f[r]}$$
(18)

where, f[r] is the one-dimensional probability density function of r, $P[v_i]$ is the probability of signal with i being the possible number, maximum a posteriori (MAP) and ML decision process is of equivalent for v_i having equal probability, the receiver has to calculate the conditional probability $f_R[r/v_i]$. Transmitting and receiving signals are discrete sequences and exist in the same vector space from reference [19]. When transmitting signal is v_i , the mathematical expectation of signal space component r_n can be obtained by vector calculation

$$E[R_{n}/v_{i}] = E[v_{in} + W_{n}] = v_{in} + E[W_{n}] = v_{in}$$
(19)

Capital letters are used to denote a random variable, and then the variance can be obtained

$$E\left[\left(R_{n}-v_{in}\right)^{2}/v_{i}\right]=E\left[\left(v_{in}+W_{n}-v_{in}\right)^{2}\right]=E\left[W_{n}^{2}\right]=\frac{N_{o}}{2}$$

Multidimensional conditional probability $r_1 r_2 \dots r_N$ on the condition v is

$$f[r_{1} r_{2} ... r_{N} / v_{i}] = \prod_{n=1}^{N} \frac{\exp\left[-(r_{n} - v_{in})^{2} / N_{o}\right]}{\sqrt{\pi N_{o}}}$$
(21)

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N is the length of the sequence *r* and the equation $f[r_1 \ r_2 \dots r_N / v_i]$ is only related to $\sum_{n=1}^{N} (r_n - v_{in})^2$. To ensure $f[r_1 \ r_2 \dots r_N / v_i]$ is the maximum value, $\sum_{n=1}^{N} (r_n - v_{in})^2$ shall take minimum value.

In (17), h and f_d are unknown. In order to estimate those two variables, 24bits training sequences, 8bits start flag and 8bits end flag are viewed as characteristic sequences and then ML estimation is done to make the estimation signals be closest to r in (17). i corresponds to the frequency offset f_d and with AWGN the cost function Λ can be expressed as

$$\Lambda = \left\| r - v \right\|^{2} = \sum_{n=1}^{N} \left| r_{n} - v_{in} \right|^{2} = \sum_{n=1}^{N} \left| r_{n} - \sum_{l=0}^{L} h_{l} s_{n-l} e^{j2\pi f_{d}(n-l)T_{b}} \right|^{2}$$
(22)

where, N = 40

To satisfy Λ minimum, $\hat{\gamma}$ can be written as

$$\hat{\gamma} = \operatorname*{arg\,min}_{\substack{\gamma \\ \gamma = \{f_d,h\}}} \Lambda = \operatorname*{arg\,min}_{\substack{\gamma \\ \gamma = \{f_d,h\}}} \sum_{n=1}^{N} \left| r_n - \sum_{l=0}^{L} h_l s_{n-l} e^{j2\pi f_d(n-l)T_b} \right|^2$$
(23)

In order to estimate $\hat{\gamma}$, functions Λ has to be partial derivatives of *h* and let the partial derivative is 0 at the same time, then

$$\frac{\partial}{\partial h} \sum_{n=1}^{N} \left| r_n - \sum_{l=0}^{L} h_l u_{n-l} \right|^2 = 0$$
(24)

where, $u_{n-l} = s_{n-l} e^{j2\pi f_d (n-l)T_b}$

Equation (24) can be sorted as

$$\sum_{n=1}^{N} u_{n-l} \left(r_n - \sum_{l=0}^{K} h_l u_{n-l} \right) = 0 \qquad l \in [0, ..., L] \quad (25)$$

Equation (25) can be easily expressed with matrix $A \cdot h = b$ where the individual elements A and b are as shown in (26). The system impulse response is $h = A^{-1}(f_d) \cdot b(f_d)$. Pilot signal vector modulated can be expressed as v = u * h with $u = (u_1, u_2, \dots, u_N)^T$ and then we can obtain the value Λ .

$$b_{m} = \sum_{n=1}^{N} u_{n-m}^{*} r_{n} \qquad m \in [0,...,L]$$

$$A_{m,k} = \sum_{n=1}^{N} u_{n-m}^{*} u_{n-k} \qquad m \in [0,...L], k \in [0,...L]$$
(26)

This process is repeated for all frequencies in the range. Among the possible values of f_d , the one that minimizes Λ is the frequency offset estimation result \hat{f}_d . Hence, $\hat{h}=A^{-1}(\hat{f}_d)\cdot b(\hat{f}_d)$, $\hat{\gamma}=\{\hat{f}_d,\hat{h}\}$.

Considering the impact of Gauss filter on the signal of the generalized channel, channel length is L+1. For AWGN channel, the channel length is 1.

A, b, \hat{h} and v are simplified into a number and then the initial phase estimation result is

$$\hat{\varphi}_0 = \arg(h) \tag{27}$$

Frequency offset \hat{f}_d and phase $\hat{\varphi}_0$ estimated by ML are compensated to the received AIS signals

$$r1 = \left\{ \left[s * h \right] e^{j2\pi f_d n T_b} + w \right\} e^{-j\left(2\pi f_d n T_b + \hat{\varphi}_0\right)}$$
(28)

Passing through the matched filter, the ideal signal phase information has not be changed and just the distributions of signal energy are more centralized. In view of only the phase information is useful to the tracking loop, therefore, the following theoretical derivation is based on the equation (28).

According to the derivation of QPSK signal phase characteristics in reference [20], phase characteristics of GMSK signal are derived. By probability theory, random variables variance is proportional to the signal amplitude square and variance is equal to AWGN power spectral

density in value, that is
$$\frac{|s(n)|}{|w(n)|} = \sqrt{\frac{2E_b}{N_o}}$$
. r1 in (28) is

proportional to r2 in (29).

$$r2(n) = \sqrt{2R}e^{j\left[\varphi(n)+2\pi\Delta \phi_n\right]} + e^{j(\Phi(n))}$$

where, $\varphi(n)$ is GMSK signals phase, $\Delta f = f_d - \hat{f}_d$, $\Delta \varphi_0 = \varphi_0 - \hat{\varphi}_0$, $\Phi(n)$ is complex AWGN phase whose mathematical expectation is $\mu = 0$ and power spectrum *E*

phase density is 1, SNR is
$$R = \frac{D_b}{N_0}$$
.

NCO output signal is

$$u(n) = e^{-j\theta(n)} \tag{30}$$

The result of multiplying compensated signals by the NCO output signals is

$$z(n) = r2(n)u(n) = \sqrt{2R}e^{j[\varphi(n) + 2\pi\Delta n T_b + \Delta \varphi_0 - \theta(n)]} + e^{j[\Phi(n) - \theta(n)]}$$
(31)

Denote $\theta_e(n) = 2\pi\Delta f n T_b + \Delta \varphi_0 - \theta(n)$, the above equation can be simplified as

$$(n) = \sqrt{2R}e^{j[\varphi(n) + \theta_e(n)]} + e^{j[\Phi(n) - \theta(n)]}$$
(32)

If the sampling rate is optimal and reserving sampling point is in the middle of symbol, GMSK can be approximately regarded as OQPSK and the corresponding phase is $\varphi(n) = \frac{\pi}{4}, \frac{3\pi}{4}, \frac{5\pi}{4}, \frac{7\pi}{4}$ during GMSK discrete process. Thus, $\sin(\varphi(n)) \in \left\{\frac{\sqrt{2}}{2}, -\frac{\sqrt{2}}{2}\right\}$,

 $\cos(\varphi(n)) \in \left\{\frac{\sqrt{2}}{2}, \frac{-\sqrt{2}}{2}\right\}$ can be obtained. Take

 $\cos(\varphi(n)) = \frac{\sqrt{2}}{2}$, $\sin(\varphi(n)) = \frac{\sqrt{2}}{2}$ for example, the real part and imaginary part of z(n) respectively are

$$I = \operatorname{Re}(z(n)) = \sqrt{R} \left[\cos(\theta_e(n)) - \sin(\theta_e(n)) \right] + \cos \left[\Phi(n) - \theta(n) \right]$$

(29)

 $Q = \operatorname{Im}(z(n)) = \sqrt{R} \left[\cos(\theta_e(n)) + \sin(\theta_e(n)) \right] + \sin[\Phi(n) - \theta(n)]$ (33)

Due to E[w] = 0, $E\left[e^{j(\Phi(n)-\theta(n))}\right] = 0$ is obtained.

Definition 1: $F(\theta_e(n)) = E[sgn(Q)I - sgn(I)Q]$ (34) where, sgn is a piecewise function

Definition2:
$$U_d(\theta_e(n)) = \frac{F(\theta_e(n))}{F(0)}$$
 (35)

Because the derivative of sgn function does not exist, we can use erf function instead of sgn function

$$U_{d}(\theta_{e}(n)) = \frac{1}{F'(0)} \left\{ \beta(\theta_{e}(n)) \operatorname{erf}(a(\theta_{e}(n))) \right\}$$

$$-\frac{1}{F'(0)} \left\{ a(\theta_{e}(n)) \operatorname{erf}(\beta(\theta_{e}(n))) \right\}$$
(36)

where,

$$\alpha(\theta_{e}(n)) = \sqrt{R} \left(\cos(\theta_{e}(n)) + \sin(\theta_{e}(n)) \right)$$
$$\beta(\theta_{e}(n)) = \sqrt{R} \left(\cos(\theta_{e}(n)) - \sin(\theta_{e}(n)) \right)$$
$$F'(0) = \sqrt{R} \left(2erf\left(\sqrt{R}\right) - \frac{4\sqrt{R}}{\sqrt{\pi}} \right)$$

When $\varphi(n) = \frac{3\pi}{4}, \frac{5\pi}{4}, \frac{7\pi}{4}$, the results of $U_d(\theta_e(n))$

are the same.

Assuming that loop middle variables is K(n), thus numerically controlled oscillator output phase is

$$\theta(n) = \Delta m(n) + \theta(n-1) \tag{37}$$

where,

$$\Delta m(n) = C_1 U_d(\theta_e) + K(n)$$

and

 $K(n) = C_2 U_d(\theta_2(n)) + K(n-1)$

IV SIMULATION RESULTS

As we know that when the radian is odd times $\pi/4$, phase shift φ has a worst effect on signal. M is AIS signal symbol length of 256. The symbol period is $T_b = 1/R_b$ and the symbol rate is $R_b = 9600$ bps. In order to make sure the signal demodulation has a little performance degradation, it requires $\varphi = |2\pi\Delta f_o M T_b| < \frac{\pi}{4} \times \frac{1}{2}$, namely $|\Delta f_o| < \frac{600}{256}$, $\Delta f_o = \Delta \hat{f} - \Delta f$, $\Delta \hat{f}$ is the tracking frequency offset and Δf is the signal frequency offset.

During tracking loop design, phase locked loop is a very important part and phase characteristics of the detector is directly related to the performance of PLL. In the low SNR, the equivalent phase characteristics are not only related to the phase error but also are affected by the SNR of the input signals. By equation (27), equivalent phase discriminators curve in different SNR is shown in Fig.6



In Fig.6, the solid lines describe $E_b / N_o = 100 dB$ phase characteristics, dotted lines stand for $E_b / N_o = 10 dB$ phase characteristics and dashed lines represent $E_b / N_o = 1 dB$ phase characteristics. What can be seen visually from the fig.6 is that the SNR is lower, the linear range of phase characteristic is narrower and loop locked is slower. In low SNR, due to the deterioration of demodulation performance with SNR decreasing, it requires the loop noise bandwidth smaller than the input signal bandwidth $B_L < B_i$. The smaller B_L , the loop to eliminating the noise is more ability.

Loop SNR can affect phase locked loop tracking accuracy and capture time (lock speed) relates to loop SNR, frequency offsets and input SNR. Taking that minimum loop SNR is 13dB, the damping coefficient $\xi \text{ is } \sqrt{2}/2$ and the data rate is 9600bps as an example, PLL capture time is calculated under different frequency and input SNR according to (9), (10), (11), and (13). It is shown in Tab.1

As we can see from Tab.1, with the input SNR decreasing or the frequency offset increasing, loop lock time will be changed to some extent. Take the minimum input SNR 1dB and frequency offset 1500Hz for example, capture time is 42.01 ms(more than 26.67ms) with a traditional phase-locked loop. The more frequency offsets, the more capture time. That is what the AIS system can not tolerate. In order to reduce the capture time, ML aided is used to reduce large Doppler frequency offsets to fast capture zone and then phase locked loop starts to track surplus frequency.

In the experiments, the normalized 3dB bandwidth is BT = 0.4 and the symbol rate is $R_b = 9600$ bps. The parameter L = 3 is referred to the correlation length. Doppler frequency shifts are 2000Hz and 3000Hz respectively. AIS datas in the frame assumed without filling datas are of 168 bits and input SNR is 1dB~10dB. Let $|\Delta f_o|$ is 2 Hz, that is to say, when the absolute value is within 2Hz, frequency offset is tracked. $K_d = 1$, $K_o = 2\pi$.

$f_d(Hz)$	3800	3000	2500	2000	1500	1000	100	10
(S/N);(dB)								
1	269.6	168.0	116.7	74.69	42.01	18.67	1.867×10^{-1}	1.867×10^{-3}
2	136.8	85.23	59.19	37.88	21.31	9.470	9.470×10^{-2}	9.470×10^{-4}
3	67.42	42.02	29.18	18.68	10.50	4.669	4.669×10^{-2}	4.669×10^{-4}
4	34.03	21.21	14.73	9.427	5.303	2.357	2.357×10^{-2}	2.357×10^{-4}
5	17.05	10.63	7.383	4.723	2.661	1.181	1.181×10^{-2}	1.181×10^{-4}
6	8.551	5.332	3.700	2.368	1.331	0.5921	5.921×10^{-3}	5.921×10^{-5}
7	4.281	2.674	1.854	1.187	0.6672	0.2973	2.973×10^{-3}	2.973×10^{-5}
8	2.146	1.341	0.9290	0.5947	0.3341	0.1493	1.493×10^{-3}	1.493×10^{-5}
9	1.077	0.6702	0.4659	0.2981	0.1683	0.07523	7.523×10^{-4}	7.523×10^{-6}
10	0.5383	0.3355	0.2331	0.1491	0.08421	0.03732	3.732×10^{-4}	3.732×10^{-6}

TABLE I. PHASE LOCKED LOOP CAPTURE TIME

The cost function Λ of frequency offset estimation error curve is as shown in Fig. 7~ Fig. 8. Fig. 9~ Fig. 10 shows the tracking phase locked loop curve under different conditions.



Figure 7. Calculation of the cost function in all Doppler range for a signal with $E_b/N_o = 8.5 dB$ and $f_d = 2000 Hz$



Figure 8. Mean error of ML estimation versus the SNR









It has been observed that the cost function value for the whole frequency range has several minima, representing different frequency candidates, as shown in Fig.7.The smallest one does not always correspond to the Doppler frequency we are looking for. For this reason, the three frequencies that produce the smallest minima are considered. It can be seen from Fig.8 that the estimation error of frequency offsets increases versus the input SNR decreasing. When the SNR is larger than 5dB, the error is within $|\Delta f_{a}| = 2Hz$ and it can meet the requirements of coherent demodulation. However, when SNR is less than or equal to 5dB, the error is greater than $|\Delta f_a| = 2Hz$ and frequency offsets need to estimate further. From Fig.9 we can obtain that when frequency offset is $\Delta f = 2000$ Hz $\Delta \hat{f} \in [1998 \text{Hz}, 2002 \text{Hz}]$ and input SNR is $(S/N)_i = 1dB$, $(S/N)_i = 22dB$ requiring number of symbols is about 150 and $(S/N)_r = 24dB$ requiring number of symbols within the range from 250 to 300. In Fig.10, when frequency offset is $\Delta f = 3000$ Hz $\Delta \hat{f} \in [2998 \text{Hz}, 3002 \text{Hz}] \text{ and input SNR is} (S/N)_i = 1 dB$ $(S/N)_L = 22dB$ capture time is less than 150 symbols and $(S/N)_{I} = 24dB$ capture time is more than 150 symbols. Fig.9~ Fig.10 show that when we use tracking loops to track frequency offsets based on the ML algorithm and set tracking error is less than $|\Delta f_o| = 2$ Hz, capture time of $(S/N)_L = 22 dB$ is less than $(S/N)_L = 24dB$. But its anti-noise performance is poor. Tracking time is no longer affected by frequency offsets but the error of ML preliminary estimating frequency offsets. To ensure that the required tracking time is short, the estimation error must be as small as possible.

V CONCLUSIONS

Frequency offset estimation methods in AWGN channel are analyzed firstly, and then a method of estimating frequency offsets precisely is proposed. The method is aided by ML algorithm and can be used under conditions of low SNR with large frequency offsets. Simulation results show that the method requiring symbol number of fast capture is less than 256(the number of AIS frame symbols) when the low SNR is 1dB and tracking precision is within 2 Hz. It can satisfy the requirements of AIS signals demodulation.

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