A Novel Channel Amplitude and Phase Correction Algorithm Based on Phased Array Space-based TT&C Communication

Yanqiu Yang*, Yi Zhao, Zhanfeng Chen, Xi Deng and Lili Wei
INFORMATION Center of Chongqing Public Security Bureau, China

*Corresponding author email: zhubl@cqu.edu.cn

Abstract. According to the characteristics of phased array space-based TT&C satellite antenna aperture size and transponder power limitation, this paper proposes a correction algorithm based on PN code and BPSK without data modulation. At the same time, combined with multi-period coherent accumulation, the channel amplitude and phase correction accuracy is guaranteed under low signal-to-noise ratio. From the simulation test, we can see that our method has the characteristics of low interception and anti-barrage jamming. Keywords: Phased array technology; Space-based TT&C communication; Calibration data processor; BPSK correction.

1. Introduction
With the rapid development of modern communication technology, new technical requirements have been put forward for police work, which has become the demand for police communication access services in the new era. Therefore, phased array measurement and control technology has increasingly become a research hotspot in police and military communication systems. Currently, it has been widely used in police unmanned reconnaissance aircraft. It is believed that it will be more widely used in other police communication industries in the near future. Channel amplitude and phase correction technology is one of the key technologies of phased array space-based TT&C communication. Its success will directly affect the beam forming quality and beam pointing accuracy of phased array antenna, and directly affect the establishment of communication link between relay satellite and spacecraft. Therefore, channel amplitude and phase correction is the key to the normal operation of phased array space-based TT&C communication system. In general, radar phased array systems guarantee channel amplitude and phase correction accuracy through high signal-to-noise ratio (about20-30dB): amplitude correction accuracy≤0.8dB(rms) and phase correction accuracy≤30(rms). However, for phased array space-based TT&C communication systems, the correction signal has low interception and anti-barrage jamming, and the signal-to-noise ratio of the correction signal is low (about-10~6dB) due to the limitation of satellite antenna aperture size and transponder power. Therefore, point frequency continuous wave correction technology is not suitable for phased array space-based TT&C communication.

In view of this, this paper proposes a BPSK correction algorithm based on PN code add no data modulation, and combines with multi-period coherent accumulation to ensure the channel amplitude and phase correction accuracy: amplitude correction accuracy ≤0.2 dB(rms) and phase correction accuracy ≤1.50 (rms).

2. System Model
The calibration model of phased array space-based TT&C communication is as follows figure 1:
Figure 1. Calibration model of phased array space-based TT&C communication.

When the forward/backward link is corrected, the correction antenna receives/transmits the correction signal, the calibration data processor CDPE (Calibration Data Process Equalizer) estimates the amplitude difference of each channel of the forward/backward link, and sends the estimated value to the beam control unit to correct the forward/backward beam weight coefficient.

The forward/backward correction signal adopts the form of PN code (m sequence) add no data modulation BPSK signal, namely:

$$S_0(t) = A_0 \cdot c_0(t) \cdot \cos(\omega t + \theta_0)$$

Where $A_0$ is the amplitude of the correction signal, $\theta_0$ is the frequency synthesizer phase difference, $c_0(t)$ is the PN code sequence, and the value is 1 or -1.

The following is a specific mathematical analysis of the backward link. The time when the correction signal reaches each array unit of the Mounted on TDRS after being transmitted by the correction antenna will be different.

Assuming that the delay to the $k$-th antenna element is $\delta_{0,k}$, the signal received by the $k$-th antenna element is:

$$E_k(t) = A_0 \cdot c_0(t - \delta_{0,k}) \cdot \cos[\omega(t - \delta_{0,k}) + \theta_0] + n_k(t)$$

Let the wave path difference between the two farthest units be:

$$\text{Diff}_{\text{MAX}} = \text{MAX}(\delta_{0,i} - \delta_{0,j}), \quad i, j = 0, 1, ..., 30$$

This value is about 2ns, that is, the wave path difference is an integer multiple of the wavelength (this value is negligible with respect to the PN chip width), and the resulting spatial difference $\phi_{0,i}$ is any value within $[0, 2\pi]$, so the above formula can be rewritten as:

$$E_k(t) = A_0 \cdot c_0(t - \delta_{0,k}) \cdot \cos[\omega t + \theta_0 + \phi_{0,k}] + n_k(t)$$

$k = 1, ..., 30$

received by the receiver and the internal thermal noise of the receiver. Both are stationary Gaussian random processes.

After being received, the correction signal is synthesized, amplified and converted by FDM and then transmitted back to the ground by Ka-band antenna. After the ground is converted, FDM separates and restores it into 30 channel signals and sends them to CDPE. After such a long link transmission, the
signal is added with delay, amplitude and phase distortion. The correction signal of channel k at the CDPE input is:

$$r_k(t) = \sqrt{G_k \cdot A_0 \cdot c_0 (t - \delta_{0,k} - \delta_k)} \cdot \cos[\omega_{IF} t + \theta_0 + \phi_{0,k} + \psi_k] + \sqrt{G_k} \cdot n_k(t)$$

(1)

Where \(\omega_{IF}\) is receive the intermediate frequency, \(G_k, \delta_k, \psi_k\) indicates the channel gain, delay, and phase shift introduced by the k-th antenna element. \(n_k(t)\) is a band-limited Gaussian white noise, which can be expressed as:

$$n_k(t) = n_{k,I}(t) \cos(\omega_{IF} t + \psi_k) - n_{k,Q}(t) \sin(\omega_{IF} t + \psi_k)$$

(2)

Where \(n_{k,I}(t)\) and \(n_{k,Q}(t)\) is baseband Gaussian white noise with equal power and independent of each other.

Theoretically, the beam forming quality and beam direction can be guaranteed only by estimating the channel gain \(G_k\) and channel phase \(\psi_k\), then offset \(G_k\) and \(\psi_k\) making corrections to the beam forming weight vector.

When the difference between different channels \(\delta_k\) is relatively large, channel delay estimation and delay equalization are required before channel amplitude and phase correction to eliminate the influence of channel delay inconsistency on amplitude and phase estimation accuracy.

3. Channel Amplitude and Phase Detection Algorithm

Compared with conventional channel amplitude and phase estimation algorithms, this algorithm has two significant advantages: first, PN code acquisition is performed before amplitude and phase estimation, and channel amplitude and phase estimation is started after acquisition; Second, the processing object of the channel amplitude and phase estimation algorithm is the correlation peak of PN code of I and Q branches, not the received signal itself. According to the corrected signal form, the algorithm can obtain SNR improvement from two aspects: 1) PN code correlation benefits; 2) Since there is no data modulation, the correlation peak of the corresponding PN code at the receiving end has no polarity change, thus multi-period coherent accumulation can be used to improve the signal-to-noise ratio during channel amplitude and phase estimation.

![Figure 2. Structure of channel amplitude and phase detection algorithm.](image)

The detection variables of channel amplitude and phase are:

$$\begin{align*}
A(k) &= \sqrt{Y_I^2 + Y_Q^2} \\
\Phi(k) &= \arctan\left(\frac{Y_Q}{Y_I}\right)
\end{align*}$$
Where YI and YQ is the correlation peaks of I and Q matched filter outputs respectively. Under the influence of noise, A(k) and $\Phi(k)$ is a random process. In order to evaluate the error range of amplitude and phase detection under a certain signal-to-noise ratio, A(k) and $\Phi(k)$ is necessary to find out the probability distribution.

The correction signal of the first channel at the CDPE input is recorded as:

$$r_i(t) = \sqrt{2s}c(t-\tau)\cos(\omega t + \theta) + n(t)$$

(3)

Where s, $\omega$ and $\theta$ are carrier power, frequency and phase difference respectively. $c(t)$ is the PN code sequence and $\tau$ is the signal delay. n(t) is an additive narrow-band Gaussian random process with a unilateral power spectral density of $n_0$. The matched filter output YI(k) and YQ(k) is a narrow-band Gaussian random process. Let $y_{ik} | \theta$ and $y_{qk} | \theta$ is the value of random process at time k under given conditions $\theta \in (-\pi, \pi)$, $y_{ik} | \theta$ and $y_{qk} | \theta$ is a Gaussian random variable and independent of each other.

$$E[y_{ik} | \theta] = \sqrt{s}M \cdot \cos \theta = y_0 \cdot \cos \theta$$

(4)

$$E[y_{qk} | \theta] = \sqrt{s}M \cdot \sin \theta = y_0 \cdot \sin \theta$$

$$D[y_{ik} | \theta] = D[y_{qk} | \theta] = MBn_0 = \sigma^2$$

Where m is the correlation length and b is the bandwidth of the baseband signal, then (YI, YQ) is a two-dimensional random variable with a joint probability density:

$$f_{YI,YQ}(y_{ik}, y_{qk} | \theta) = \frac{1}{2\pi\sigma^2} \exp\left\{-\frac{1}{2\sigma^2} \left[ \left(y_{ik} - y_0 \cos \theta\right)^2 + \left(y_{qk} - y_0 \sin \theta\right)^2 \right] \right\}$$

(5)

Applying the method of finding the function distribution of random variables, the rectangular coordinates are mapped to polar coordinates to obtain the two-dimensional distribution of (A, $\Phi$):

$$f_{\phi\theta}(a, \phi, \theta) = |a| \cdot f_{YI,YQ}(y_{ik}, y_{qk} | \theta) = \frac{a}{2\pi\sigma^2} \cdot \exp\left\{-\frac{1}{2\sigma^2} \left[ a_1^2 + y_0^2 - 2a_1y_0 \cos(\theta - \phi) \right] \right\}$$

$$a_1 \geq 0, -\pi < \phi, \pi$$

(6)

3.1. Edge Distribution of Amplitude Detection Value A $f_A(a_k)$

The output of the matched filter has obtained the spread spectrum gain. Combined with multi-period coherent accumulation, it can be considered that the detection variables A and $\Phi$ are carried out under the condition of large signal-to-noise ratio. At this time, the one-dimensional probability density of the amplitude A is approximately Gaussian distribution:

$$f_A(a_k) \approx \frac{1}{\sqrt{2\pi\sigma}} \exp\left\{-\frac{(a_k - y_0)^2}{2\sigma^2} \right\}, a_k \geq 0$$

(7)

According the $3\sigma$ rules, the value interval of random variables is $y_0 \pm 3\sigma$, error is $\pm 3\sigma = \pm 3\sqrt{Mb n_0}$, Normalized and logarithmically obtained:
\[ A_{\text{Error}(\text{db})} = 10\log_{10} \left( \frac{y_n + 3\sigma}{y_0} \right) = 10\log_{10} \left[ 1 \pm \frac{3\sqrt{M\sigma^2}}{sM} \right] = 10\log_{10} \left[ 1 \pm \frac{3}{\sqrt{2BN_0}M} \right] \] (8)

Assuming the signal-to-noise ratio at the receiver input \( SNR = \frac{s}{2BN_0} \), the above formula can be rewritten as:

\[ A_{\text{Error}(\text{db})} = 10\log_{10} \left[ 1 \pm \frac{3}{\sqrt{2\cdot SNR_{re}\cdot M}} \right] \]

3.2. Edge Distribution of Phase Detection Value \( \Phi \)

Under the condition of large signal-to-noise ratio, the phase detection value is approximately Gaussian distribution.

\[ f_{\Phi}(\phi | \theta) \approx \frac{\rho}{\sqrt{2\pi}} \exp \left\{ -\frac{\rho^2}{2} (\theta - \phi_k)^2 \right\} \] (9)

Where mean value is \( \theta \), variance is \( \frac{1}{\rho^2} \), \( \rho = \frac{y_0}{\sigma} \) represents the ratio of signal amplitude to standard deviation of narrowband noise, that is, the output signal-to-noise ratio of matched filter.

According to rule \( 3\sigma \), the range of random variables is

\[ \Phi_{\text{Error}} = \frac{\theta \pm 3}{\rho} = \frac{\pm 3\sigma}{sM} = \pm \frac{3\sqrt{MBn_0}}{\sqrt{sM}} = \pm \frac{3}{\sqrt{2\cdot SNR_{re}\cdot M}} \] (10)

3.3. Numerical Analysis

Assuming that PN code is m-sequence, code length is 1023, code clock is 3.1 MChip/s, and one-cycle PN code is correlated, the simulation results of channel amplitude A detection error band and channel phase \( \Phi \) detection error band are respectively:

**Figure 3.** Single-period channel amplitude detection error.

**Figure 4.** Single-period channel phase detection error.

Assuming that PN code is m-sequence, code length is 1023, code clock is 3.1 MChip/s and 16-cycle
PN code is correlated, the simulation results of channel amplitude A detection error band and channel phase \( \Phi \) detection error band are respectively:

![Figure 5. Error of channel amplitude detection after 16 cycles accumulation.](image1)

![Figure 6. 16-periods accumulated channel phase detection error.](image2)

### 4. Conclusion

For the satellite-to-ground reverse link, the inconsistency of device parameters and selective fading of wireless transmission channels between receiving channels of each antenna unit make the channel phase and amplitude distortion of received signals different in each channel, which leads to the difference of amplitude between signals received by different channels. If the channel amplitude and phase are not corrected before beamforming, it will inevitably affect the performance of digital beamforming, and even make the beam pointing error. Thus, channel amplitude and phase correction is the key to the normal operation of phased array space-based TT&C communication system. In this paper, a channel amplitude and phase correction algorithm is proposed, and the differences between the proposed algorithm and the existing algorithms are analyzed in detail. Firstly, the PN code is captured before the amplitude and phase estimation, and then the channel amplitude and phase estimation is started after the acquisition. Secondly, the processing object of the channel amplitude and phase estimation algorithm is the correlation peak of the PN code of branch I and Q, not the connection. Receive the signal itself. After detailed theoretical deduction, the calculation formulas of channel amplitude detection error band and channel phase detection error band are obtained. The simulation results show that the channel amplitude and phase correction technology has the characteristics of low interception and anti-blocking interference. At the same time, the output signal-to-noise ratio of the matched filter is effectively improved by combining the phase coherent accumulation. The method proposed in this paper can guarantee high channel amplitude and phase correction accuracy under low signal-to-noise ratio, and is very suitable for phased array space-based TT&C communication system.

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