Self-calibration of Spaceborne Membrane Phased Array

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***Abstract*—Spaceborne membrane phased array is prone to distortion due to lack of mechanical rigidity. Distortion introduces phase errors between array elements which will significantly degrade antenna performance, such as directivity and sidelobe level. An innovative self-calibration method was invented to compensate phase errors with minimum equipment involved. Using the un-calibrated array itself to transmit and receive ground reflective signal, the phase errors between elements can be estimated and then be compensated. Simulation demonstrated the principle of the algorithm and show promising results for further investigation.**

Keywords—membrane phased array; spaceborne antenna; calibration

I. INTRODUCTION

Future space missions need low mass, low cost and high packaging efficiency structures to reduce launch cost, stow volume and production costs. The innovation of highly deployable membrane phased array antenna satisfies the requirements and promise wide application in future space missions.

Spaceborne membrane phased array antenna (Figure 1) is manufactured on light weight membrane structures. This technology enables the use of larger antennas for high speed space communication and low frequency earth observation. Some earth remote sensing applications, such as soil moisture and ocean salinity, need the antenna to be physically large in order to obtain the necessary resolution at the frequencies of interest (low frequency). Membrane antenna also makes it possible to place huge Synthesized Aperture Radar (SAR) systems on Medium or Geosynchronous Earth Orbits (MEO or GEO) to compensate signal space loss and improve the field of view for earth observation which would lead to much shorter revisit times in comparison to satellites in a Low Earth Orbit (LEO). Moreover, membrane antenna is a key enabling technology for miniaturized satellite, for example, CubeSat, for its super light weight and high stow efficiency.

The forerunner in this research field is Jet Propulsion Laboratory (JPL) where a membrane phased array demonstrator has been built including structure, radiator and electronics design. The primitive demonstrator has shown huge reduction of mass in comparison with traditional phased array antenna [1][2].

When applying membrane phased array to real system, there are many technical problems need to be solved. One of them is as the result of lack of mechanical rigidity; membrane structure is prone to distortion. Such distortion may be introduced during launch vibration, during folding and expending or caused by long term thermal effect. Besides, as fabricated on such highly elastic structure, antenna array will vibrate when spacecraft maneuvers, as happened in dual side looking SAR operations. Both of mechanical distortion and structure vibration generate relative displacements between array elements which will significantly degrade radiation performance of the array as a whole. Therefore, these displacements need to be calibrated, that is, to estimate and compensate the relative phase differences of antenna elements.

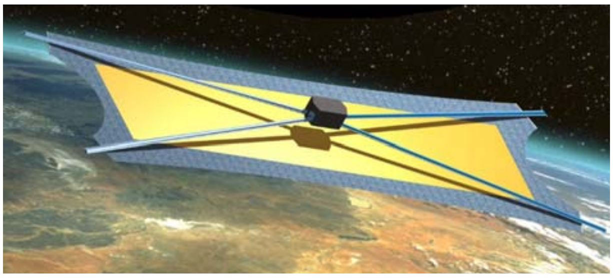

Figure. 1. membrane phased array in space (imaginery picture)

There are some existing technologies to calibrate array element displacement and radiation pattern. As introduced in [3][4], the phase differences between array elements can be accurately measured by near field probe antenna attached on the spacecraft. The advantage of this technology is it can correct the phase errors with high accuracy. The major disadvantage is extra mass and cost, increased inherent technical risk to the spacecraft as the near field probe requires more mechanical structure.

Another method measure the phase differences between array elements with pre-installed ground transmitters, receivers or corner reflectors [5][6]. The process is time consuming and costs a lot of labor.

A more convenient scheme use reflective signal from illuminated patch to calibrate the antenna [7]. This kind of technology is more convenient because no ground facilities are needed. It processes signal reflected from noncoherent ground scatterers and takes advantage of the statistical properties of the backscattered clutter signal to estimate phase errors in the array [8]. The key idea is that a homogeneous clutter scene will make the phase of the ensemble correlation between signals received by adjacent array elements equal to zero. Thus, any variation from this zero phase correlation attributes to relative phase error between the two elements and therefore can be used to calibrate the array.

A large number of nonrigid scatterers, which are uniformly distributed and with random relative motions, can provide the required noncoherence, for example, sea waves and rainforest. And a well calibrated transmitting antenna installed on the spacecraft is also needed to illuminate the patch for the required homogeneous clutter scene.

Our algorithm is based on such noncoherent scatterer algorithms, the difference from former ones is that the independent well calibrated transmitting antenna has been removed to reduce support structure complexity. Although adding some computation complexity, our method minimized the structure demand, reduced cost, and achieved real self-calibration by using uncalibrated antenna array to calibrate itself. Since the transmitting antenna and the receiving antenna is the same array. This algorithm can calibrate transmit channel and receive channel at the same time. Theoretical description of the algorithm is in Section II, followed by simulation demonstration in Section III, and finally concluded in Section IV.

II. ALGORITHM

Unlike the well calibrated illuminating antenna, transmitting signal with the uncalibrated antenna array can't generate the homogeneous clutter scene we need. But when randomize transmitting phase in each pulse and accumulate different pulses, we can achieve equivalent result as required from homogeneous clutter scene. The underlying principle of our algorithm is presented here.

The signal received by i th element is:

$$\begin{aligned} e_i'(n) &= e_i(n)e^{j(\phi_i + \frac{2\pi}{\lambda}d_i)} + \eta_i(n) \\ &= e_i(n)e^{j\delta_i} + \eta_i(n) \end{aligned} \quad (1)$$

where $e_i(n)$ is the received signal at each sampling time when there is no phase error between adjacent elements. Since phase errors exist, the signal phase will change. ϕ_i is for electronics delay of different channel, d_i is physical displacement of the array element deviated from its nominal position and η_i is measurement noise. Since λ is the wavelength, $2\pi d_i/\lambda$ is the phase difference from displacement. The phase delay from electronics and displacement can be added together to δ_i. As it will be showed in following description, our algorithm can actually estimate electronics and displacement phase change at the same time which is favorable for real system application. If we set the phase of one element as reference phase, then we only need to estimate the phase difference between reference element and other elements $\Delta\delta_{ij} = \delta_i - \delta_j$. Conjugate multiply can achieve the phase difference as

$$e_i'(n)e_j'^*(n) = e_i(n)e_j^*(n)e^{j(\delta_i-\delta_j)} + noise\ term \quad (2)$$

In order to reduce the noise, we need to average the signal

$$E\left[e_i'(n)e_j'^*(n)\right] = E\left[e_i(n)e_j^*(n)\right]e^{j(\delta_i-\delta_j)} \quad (3)$$

let $\psi_{ij} = \arg\{E(e_i'e_j'^*)\}$, $\beta_{ij} = \arg\{E(e_ie_j^*)\}$, then $\psi_{ij} = (\delta_i - \delta_j) + \beta_{ij}$. If β_{ij} is 0, then $\Delta\delta_{ij} = \psi_{ij}$, thus phase difference can be estimated from measured signal $e_i'(n)$, $e_j'(n)$. The condition of $\beta_{ij} = 0$ can only be satisfied when $E\left[e_i(n)e_j^*(n)\right]$ is real.

For convenience, we give the theoretical analysis of one-dimensional situation to explain our algorithm. In this situation, e_i can be modeled as:

$$e_i = \int_{-\infty}^{+\infty} f(x)e^{-j\frac{4\pi}{\lambda}r_i} w_T(x)w_R(x)dx \quad (4)$$

where $f(x)$ is radar backscattering coefficient. r_i is distance from array element to ground scatterers as show in Figure 2. Figure 2 visualized the one-dimensional situation, where two array element are showed in this figure, among them, one will be taken as reference. The vertical distance from antenna array to ground is h. $w_T(x)$ is ground project of antenna array transmitting pattern and $w_R(x)$ is the receiving pattern of individual element. 4π is for two-way distance delay include both transmitting and receiving path. Since the unknown array element displacement, we have no idea of the radiation pattern of the transmitting array. Nevertheless, we can randomize the transmitting array phase at each pulse and average the receiving signal from pulse to pulse. As draw in Figure 2, transmitting beam pattern is completely randomized.

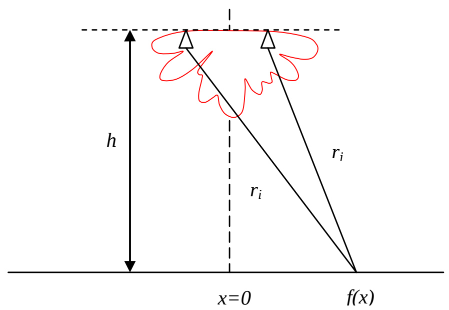

Figure. 2. visualization of randomised radiation pattern

From (4). The conjugate correlation of received signal is:

$$\begin{aligned} E\left[e_ie_j^*\right] = E\{&\int_{-\infty}^{+\infty}\int_{-\infty}^{+\infty} f(x)f^*(x)e^{-j\frac{4\pi}{\lambda}(r_i-r_j)} \\ &w_T(x)w_T(x')w_R(x)w_R(x')dxdx'\} \end{aligned} \quad (5)$$

where $E\{f(x)f^*(x')\} = E\{|f(x)|^2\delta(x-x')\} = \sigma_0\delta(x-x')$ for noncoherent ground scatterers. $w_T(x)$ is a random process. Notice in (5), $r_i - r_j$ can be simplified by binomial expansion:

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$$r_i - r_j = \sqrt{h^2 + (x + D/2)^2} - \sqrt{h^2 + (x - D/2)^2}$$
$$\approx h\left(1 + \frac{(x + D/2)^2}{2h^2} - 1 - \frac{(x - D/2)^2}{2h^2}\right) = \frac{Dx}{h}$$

Then, (5) reduced to:

$$E\left[e_i e_j^*\right] = \sigma_0 \int_{-\infty}^{+\infty} e^{-j\frac{4\pi D}{\lambda h}x} p(x) q(x) dx \quad (6)$$

where $p(x) = E\left[w_T(x + D/2) w_T(x - D/2)\right]$, $q(x) = E\left[w_R(x + D/2) w_R(x - D/2)\right]$. Let $\omega = \frac{4\pi D}{\lambda h}$, we got:

$$E\left[e_i e_j^*\right] = \sigma_0 \int_{-\infty}^{+\infty} e^{-j\omega x} p(x) q(x) dx \quad (7)$$

which is in the form of Fourier transform. According to the properties of Fourier transform, if $p(x)q(x)$ is a real and even function, then $E[e_i e_j^*]$ will be real. Therefore $\beta_{ij} = \arg\left\{E(e_i e_j^*)\right\}$ will be zero and phase difference can be estimated $\Delta\delta_{ij} = \psi_{ij}$. Since $p(x)$ is the expectation of completely randomized radiation pattern, which can be taken as a random process, $p(x)$ will be even as the expectation will show symmetry pattern around central axis. $q(x)$, as the function related to array radiation element, will submit to the property of element radiation pattern. Here, we take a typical radiation pattern, microstrip patch antenna element to analyze. The pattern can be described through approximation as:

$$w_R'(\theta) = \frac{\sin\left(\frac{kh_e}{2}\sin\theta\right)}{\frac{kh_e}{2}} \frac{\sin\left(\frac{kw}{2}\cos\theta\right)}{\frac{kw}{2}} \quad (8)$$

which is showed in Figure 3. Figure 3 present the definition of θ and the pattern for a patch antenna element. In (8), $k = 2\pi/\lambda$, h_e is patch height, w is width of the patch.

With some substitutes, $w_R'(\theta)$ can be transformed to

$$w_R(\theta) = \frac{\sin\left(\frac{kh_e}{2}\frac{x}{\sqrt{x^2+h^2}}\right)}{\frac{kh_e}{2}} \frac{\sin\left(\frac{kw}{2}\frac{h}{\sqrt{x^2+h^2}}\right)}{\frac{kw}{2}\frac{h}{\sqrt{x^2+h^2}}} \quad (9)$$

according to (9), $q(x) = w_R(x + D/2) w_R(x - D/2)$ is an even function. Then in (7), $p(x)q(x)$ will be even, therefore, $E\left[e_i e_j^*\right]$ will be a real function $\beta_{ij} = \arg\left\{E\left[e_i e_j^*\right]\right\} = 0$. This means if antenna array elements are microstrip patches, our algorithm will guarantee the phase difference between reference element and uncalibrated element can be estimated. In fact, the analysis above also guarantee that as long as array element radiation pattern is symmetric which is real for most applications, the algorithm will work well.

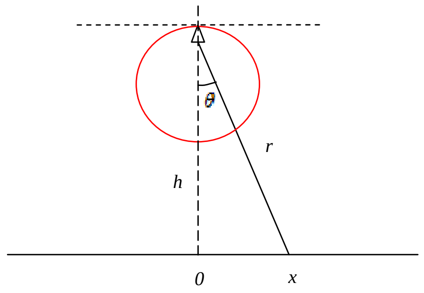

Figure. 3. approximate radiation pattern of a patch antenna element

III. SIMULATION

We need some simulation to demonstrate the algorithm and first to model the received signal. Assume the radar transmits signal pulse $s_T(t)$

$$s_T(t) = g(t)\exp(j2\pi f_0 t) \quad (10)$$

where $g(t)$ is the complex envelop and f_0 is carrier frequency. The received signal from scatterer i will be:

$$s_i(t) = f_i(\theta,\beta) w_T(\theta,\beta) g\left(t - \frac{2r}{c}\right) \exp\left(j2\pi f_0\left(t - \frac{2r}{c}\right)\right) \quad (11)$$

c represents the speed of light, r is distance between radar and a scatterer, then the round-trip time delay is $2r/c$. The received signal is multiplied by factor $f(\theta,\beta)$, determined by the scatterer reflectivity as well as the elevation angle θ and azimuth angle β, and w_T, the beampattern. When striped off the carrier frequency, the echo from scatterer i would be:

$$s_i(t) = f_i(\theta,\beta) w_T(\theta,\beta) g\left(t - \frac{2r}{c}\right) \exp\left(-j\frac{4\pi r}{\lambda}\right) \quad (12)$$

where λ is the wavelength. By combining echoes from each scatterers, the composite signal is given by

$$e(t) = \sum_{i=1}^{N} s_i(t) \quad (13)$$

where N is the number of scatterers.

The i th element delivers measured field value of $e_i'(n)$ to the signal processor. In order to get $E\left[e_i' e_j'^*\right]$, time sample average is applied to approach the assembly expectation. $E_T = \frac{1}{M}\sum_{m=1}^{M} e_i'(m) e_j'^*(m)$. There are two strategies to realize the sample average. One is just averaging echoes from multiple pulses; the other is averaging from both different pulses and

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successive range bins. The latter will work when the geometrical features are statistically homogeneous over the whole area, thus echoes from different range bins can be treated as sample function drawn from the same random process.

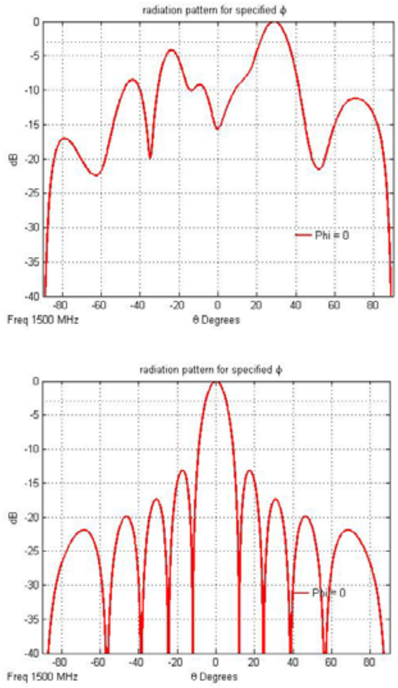

Figure. 4. radiation pattern before (up) and after (down) compensation 2D

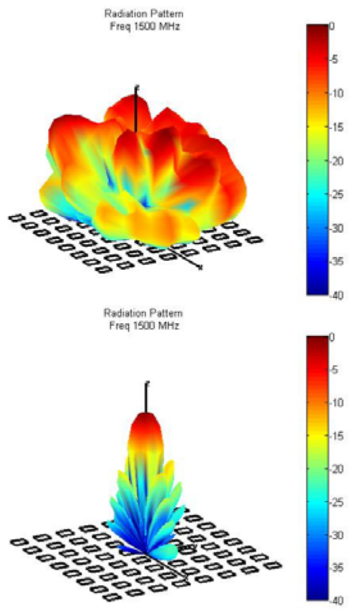

Figure. 5. radiation pattern before (up) and after (down) compensation 3D

Figures 4 and 5 showed the simulation results. The array was constructed by 64 elements. Given working frequency 1.5GHz, we design the patch firstly. The up sides of Figure 4 and 5 is the radiation pattern before calibration, the down sides are after calibration.

Simulation results proved the efficiency of the algorithm. Phase errors between elements can be estimated and calibrated with high accuracy.

IV. CONCLUSION

A problem when applying membrane phased array to real system is that the light weighted non-rigid membrane structure is prone to distortion. As the result of distortion, relative position of array radiation element will change and cause relative phase errors which dramatically degrade antenna performance and imaging capability. Therefore, these phase errors need to be calibrated and compensated. By measuring the correlation of backscattered clutter signal from noncoherent ground homogeneous scatterers, unknown phase errors between array elements can be estimated. The problem of non-uniform illumination arise from unfocused array can be tackled by average pulses with randomized transmitting phase. Simulation result showed even when the array is completely unfocused, the algorithm can still work.

Randomized radiation pattern will reduce imaging power in real system which may generate some other problems. Since the array will not completely unfocused at the first time, the decrease of transmitting power will not affect the efficiency of our algorithm, however, phase error estimation accuracy may be affected which need further investigation.

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