Wireless Pers Commun (2011) 61:143–159
DOI 10.1007/s11277-010-0015-7

Broadband Frequency Invariant Beamformer

Ning Xie · Hui Wang · Hongwei Liu

Published online: 22 May 2010

Abstract In the broadband signal processing, the array has different relative aperture for the different frequency bins, which results in waveform distortion. Moreover, the greater the bandwidth is, the more serious the distortion becomes. It is valuable to study the Frequency-invariant beam patterns (FIBPs) for receiving broadband signals without distortion. Based on the array dimensions, this paper will summarize some new methods to design a broadband beamformer with an FIBP. There will be two categories: One-dimensional arrays and Multi-dimensional arrays. For one-dimensional array, there are sampling rate method, minimax frequency invariant beamforming, etc. For multi-dimensional array, there are Bessel function method, Bessel function and phase mode method, and so on. Finally, we will discuss the pros and cons of every method.

Keywords Broadband signals · Smart antenna · Frequency-invariant beam patterns

1 Introduction

In the array signal processing, the beamforming techniques are often used to receive the signal from the desired direction while to prevent the direction of interference. For the narrow-band signals, the time delay corresponds to a certain phase shift and the complex weight can simultaneously adjust the signal amplitude and phase. Therefore, the conventional beamformer achieves the array gain and directivity by a set of complex weights. The performance of beamformer, such as spatial resolution, depends on the relative aperture of the array, which is the ratio between absolute aperture and the wavelength of signal, instead of the absolute aperture.

N. Xie (✉) · H. Wang · H. Liu
College of Information Engineering, Shenzhen University, 518060 Shenzhen, Guangdong, China
e-mail: kensouren@yahoo.com.cn

N. Xie
National Mobile Communications Research Laboratory, Southeast University,
210096 Nanjing, Jiangsu, China

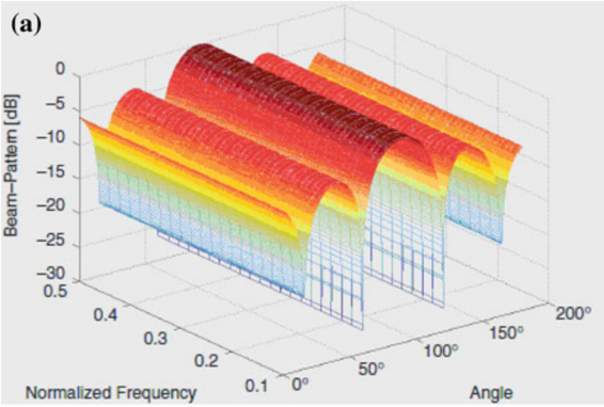
Some typical applications, such as audio conferencing systems, sonar, and underwater communication, usually need to deal with broadband signals. The maximum delay of the narrow-band signals is much smaller than the inverse of the bandwidth B, so the signal envelope delay along the array can be negligible, while the broadband signal is different, for example, B is much large, and the delay can't be ignored. Therefore, many researchers began to study broadband array for processing broadband signals, such as Widrow et al. [1] firstly proposed the tap delay broadband adaptive array, and the time delay lines (TDL) structure proposed later by Frost was the conventional broadband beamformer [2]. At the same time researchers studied the broadband signal beamforming under the different array structure. For example, Yunhong [3] researched the circular array and Wang [4] dealt broadband signal with the conformal array.

Because the bandwidth of the broadband signals is large, the difference of response at different frequencies can't be ignored. Therefore, the array has the different gains, resulting in waveform distortion. Moreover, the greater the bandwidth is, the more serious the distortion becomes. In order to receive broadband signals without distortion at all the frequency bins, it is necessary to research the frequency-invariant beamformer (FIB), which is very useful for ultrasonic detection, seismic survey, especially in underwater acoustic communication and signal spectrum detection [5]. FIB can make all frequency components of the interferers point to the nulls, which will increase the antenna gain and SINR, save transmission power and extend battery life. Figure 1a, b show an example of a frequency-invariant beam-pattern and a frequency-variant beampattern in the normalized frequency band of [0.1, 0.5], respectively, which is normalized over the sampling frequency [5].

For the one-dimensional array, researchers often use the structure of a tapped delay line transversal filter to deal with broadband signals. For the far-field signals, Ward et al. [6] proposed two methods of implementing FIR filters for a frequency invariant beamformer: one method applied the multirate processing technique, and the other was based on the single sampling rate. Both adopted a single underlying set of filter coefficients obtained directly from the desired beamformer response. The former method requires fewer filter coefficients but at the expense of a higher sampling rate compared to the latter one. Wei Liu et al exploited the Fourier transform relationship between the array's spatial and temporal parameters, then through starting from the desired frequency-invariant beam pattern, adopted a series of substitutions and an inverse Fourier transform to achieve a frequency-invariant beamformer design [7,8]. I.D.Dotlic optimized the vector of the beamformer coefficients in order to minimize the norm difference between the desired and realized pattern in the minimax sense, numerical examples showed this method had good results in shaping the sidelobes' envelope while keeping the main lobe width and direction equal to user-predefined values over the desired frequency range [9].

Although the case of aperture $D < \lambda$ (λ denotes wavelength of the signal) is unfavorable and quite uncommon, it occurs in several important applications, such as underwater communication, microphone array. For those arrays with closer element spacing, the superdirectivity technique produces a beam pattern with an acceptable profile and a useful directivity value. The superdirective beam pattern was achieved by synthesizing a specific apodization window, which was more robust than random array imperfections [10,11].

For multi-dimensional arrays, researchers are more likely to deal with broadband signals using circular arrays since it is symmetric and provides nearly invariant beampattern for 360° azimuthal coverage. Circular arrays can consist of either a single ring or multiple concentric rings. Yunhong and Chen achieved FIB by multiple rings [3,12,13], which



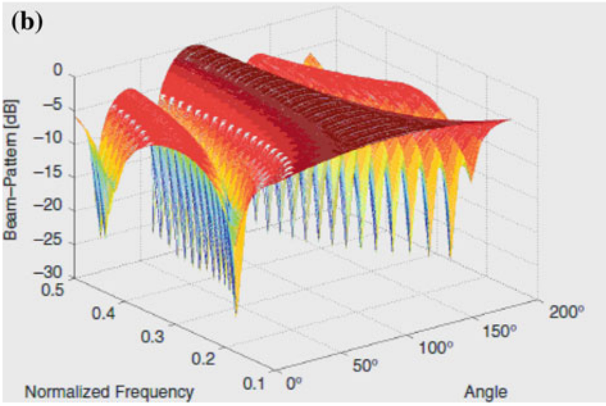


Fig. 1 Frequency-invariant beam-pattern and frequency-variant beam-pattern in the normalized frequency band of [0.1, 0.5], **a** frequency-invariant beam-pattern; **b** frequency-variant beam-pattern [5]

used fast Fourier transform(FFT) to separate the array signal into narrow-band components. By utilizing the multiring structure of a concentric ring array (CRA), the beamformer weights were decomposed into two sets: intra-ring weights and inter-ring weights. The intra-ring weights were chosen to be delay-and-sum weights and the inter-ring weights were deduced from Fourier-Bessel series expansion. The array patterns over a range of frequencies were able to approximate the same desired array pattern [3]. Chen transformed the received signals to the phase mode and removed the frequency dependency of the individual phase mode through the use of a digital beamforming or compensation networks. As a result, the far-field pattern of the array was approximately invariant over design frequencies range [12,13]. Besides, Wang proposed a uniform methods for the element polarized pattern transformation of arbitrary 3D conformal arrays, and a space-time-polarisation filter structure for frequency-invariant and low cross-polarisation pattern synthesis of conformal arrays [4].

The rest of this paper is organized as follows: In Sects. 2 and 3, we introduce the methods using one-dimensional arrays and multi-dimensional array, respectively. Moreover, we discuss the pros and cons of every method. Finally, we give a conclusion about this paper in Sect. 4.

2 The Frequency-Invariant Methods using One-Dimensional Arrays

2.1 The Frequency-Invariant Methods for Far-Fields Wave

2.1.1 Sampling Rate Method

The response of a linear continuous aperture to planar waves from an angle θ measured relative to broadside is

$$\tau(\theta, f) = \int_{0}^{X_{\max}} e^{j2\pi f x c^{-1} \sin\theta} \rho(x, f) dx \tag{1}$$

Where $\rho(x, f)$ is the aperture illumination, which is considered to be a continuous function of both location x and frequency f, and c is the speed of wave propagation. It necessary to assume that $\rho(x, f) = 0$ for $x > x_{\max}$, where $x_{\max}$ is the maximum location of the antenna. It is straightforward to show that the response of a broadband aperture remains constant if the aperture illumination in (1) is given by$\rho(x, f) = fG(xf)$, where $G(\bullet)$ is an arbitrary absolutely integrable function. Therefore, the filter of an FIB can be broken into two parts: (1) the primary filter response, $H_x(f) = G(xf)$ and (2) the secondary filter response, f, which is independent of position. As Fig. 2 shows.

For a set of $N + 1$ sensor locations with the zeroth sensor located at $x_0 = 0$ with frequency limited to the range$[f_L, f_U]$, the frequency invariant response equal to $f \sum_{n=0}^{N} g_n H_n(f) e^{j2\pi f x_n c^{-1} \sin\theta}$, where g_n is a spatial weighting term to compensate for the possibly nonuniform sensor spacing, and $H_n(f) = G(x_n f)$ is the primary filter response of the nth sensor.

The set of sensor locations can be determined by minimizing the number of required sensors while avoiding spatial aliasing [6]. The aperture length is P half-wavelengths, and there are thefollowing two design methods.

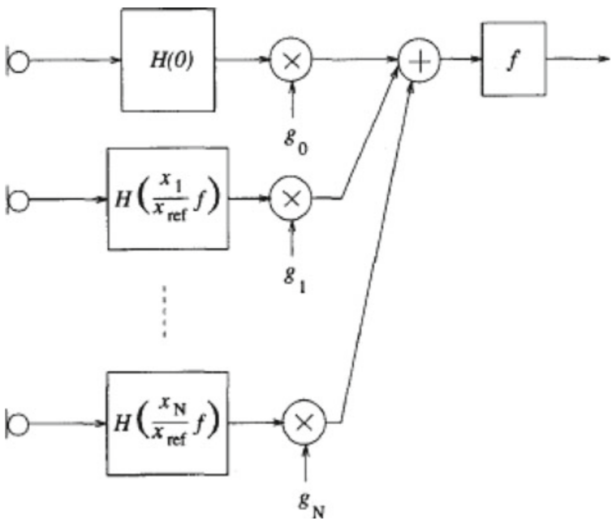


Fig. 2 Block diagram of a general linear FIB [6]

A. Multirate Method

A common method of multirate sampling is to sample every sensor at the highest rate and then to use decimation to achieve the desired sampling rate. Thus, each of the primary filter would be implemented by downsampling by $\gamma_n = x_n/x_{ref}$, applying the reference primary filter, and then upsampling by γ_n. The reference primary filter is located at $x_{ref} = c/(2f_U)$, where f_U is signal bandwidth.

B. Single Rate Method

If the input signal is band-limited to f_U, the minimum sampling rate is $f_s = 2f_U$, the location of the reference sensor should be chosen $x_{ref} = Pc/(2f_U)$.

Two examples are presented to demonstrate the effectiveness of each method. In both examples, the aperture length is $P = 4$ half-wavelengths, and the design bandwidth is 200–3,400 Hz, requiring 17 sensors and a total array size of 3.4 m. A secondary filter with 12 coefficients and a uniform aperture illumination is used in both cases.

Figure 3a shows the response of the multirate FIB with a maximum sampling rate of 30 kHz and a reference filter with nine coefficients. Figure 3b shows the response of the single sampling rate FIB with a sampling rate of 8 kHz. It is seen that, in both cases, the beam pattern is relatively FI over the entire design bandwidth. The multirate method requires fewer filter coefficients but at the expense of a higher sampling rate.

2.1.2 Minimax Frequency Invariant Beamforming

The physical structure that is being addressed here is a wideband beamformer with N antenna array elements and an M complex coefficient FIR filter on each array element. For this

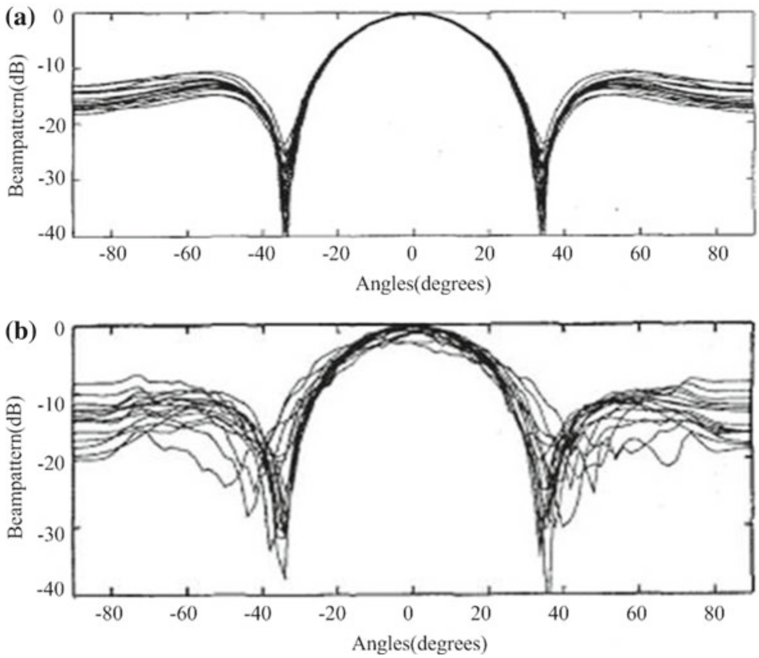


Fig. 3 Beam pattern of a 17-element at 15 frequencies within the design bandwidth: **a** Using multirate processing; **b** using a single sampling rate [6]

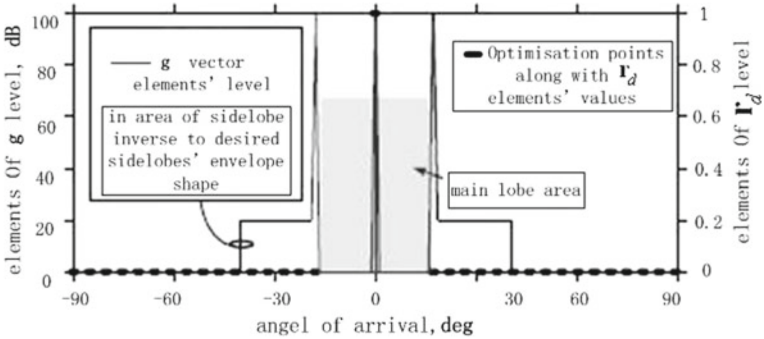


Fig. 4 Optimization points distribution along angular axis with corresponding $\mathbf{r}_d$ vectors elements' values [9]

purpose, no constraints are being put on the antenna array geometry, i.e., the array element' positions and patterns are arbitrary. For the set of discrete optimization points over angle and frequency, the problem is to find the vector of the complex beamformer coefficients w, which makes the norm of the difference between the realized beamformer pattern and desired one minimal.

In view of the general array theory, the beamformer response for the CW (Continuous Wave) signal from direction θ with a frequency f and amplitude equal to one, $r(\theta, f)$ can be represented as the product of steering vector' transpose and weights. For the set of discrete optimization points $\{(\theta_i, f_i)\}$, $1 \leq i \leq P$, the array output $\mathbf{r}$ can be arranged the product of signals vector $\mathbf{x}$ and weights $\mathbf{w}$. If the vector $\mathbf{r}_d$ is defined as the desired array response for the same set of optimization points, and the vector $\mathbf{e}$ as the difference between $\mathbf{r}$ and $\mathbf{r}_\mathrm{d}$. Then the optimal coefficients vector of the complex beamformer is defined as one that minimizes the norm of the $\mathbf{e}$, i.e. $(\min_{\mathbf{w}}\{\max|\mathbf{Xw} - \mathbf{r}_d|\})$. The minimax algorithm which gives the designer various possibilities in pattern synthesis is designed to solve this problem. Here the case of the FIB is discussed. In this case, the optimization should be performed for the set of discrete frequencies, uniformly distributed between minimum and maximum frequency of optimization. For each frequency of optimization, the optimization points should be placed over angular dimension as Fig. 4 shows. Since in the main lobe area optimization points are placed only in the directions of the main lobe maximum and its borders, so the $\mathbf{e}$ element magnitudes at the end of the optimization is very low. In this way the main beam width is precisely controlled and the frequency selectivity of the main beam maximum is minimized [14, 15].

2.1.3 Substitutions Method

For a one-dimensional sensor array aligned with the x axis, the beam response of a planar wave with angular frequency w from an angle θ measured from broadside is given by

$$P(w, \theta) = \int_{-\infty}^{\infty} e^{-j\frac{w\sin\theta}{c}} D(x, w)dx \tag{2}$$

Where c is the propagation speed and $D(x, w)$ is the frequency response of the sensor at location x. By substitution $w_1 = (w\sin\theta/c)$, we have $P(w_1, w) = \int_{-\infty}^{\infty} e^{-jw_1x} D(x, w)dx$. It is clear that $P(w, \theta)$ can be obtained by first applying a 1-D Fourier transform to $D(x, w)$ and then resubstituting $w_1 = (w\sin\theta/c)$. In order to achieve frequency invariant pattern,

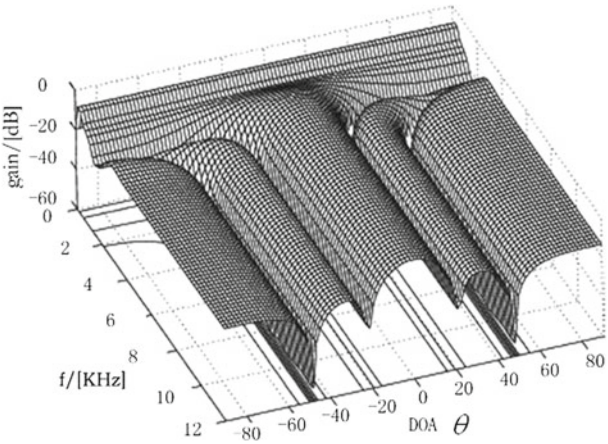


Fig. 5 Resultant frequency invariant beam pattern for the linear array design. [7]

$P(w_1, w)$ must be a function of only θ, or more preciselysin θ, let $F(\sin\theta)$ be such a frequency invariant beam pattern. In order to match this desired beam pattern, $P(w_1, w)$ must, after resubstituting $w_1 = (w\sin\theta/c)$, be identical to $F(\sin\theta)$. In order to achieve this, the variables w_1 and w must obey a specific dependency in the expression of $P(w_1, w)$ for w to disappear. Note that if $P(w_1, w)$ is a function of $c(w_1/w)$, then after the resubstitution, $c(w_1/w)$ will change to $\sin\theta$, Thus elimination any dependency on w. Therefore, note $P(w_1, w) = F(w_1c/w)$ for a desired frequency-invariant beam pattern $F(\sin\theta)$. Based on the above analysis, the idea of designing a frequency invariant beamformer can be summarized as follows. Given a desired frequency-invariant beam pattern $F(\sin\theta)$, which can be any function of the $\sin\theta$, with any shape, and can be obtained by a narrowband beamformer design method. For $P(w_1, w) = F(w_1c/w)$, in order to get the desired frequency response $D(x, w)$ for any position x. applying an inverse Fourier transform to $P(w_1, w)$ with respect to the variable w_1. $D(x, w)$ can then be realized by either an analog filter or digital filter, which can be obtained by an appropriate filter design method.

There is one example for the design of the frequency invariant beamformer with potential applications to microphone arrays. The frequency range of the interest is set to be $0 < f \leq$ 12 kHz and the propagation speed is 340 m/s. This example is for a linear array with 21 sensors with an adjacent sensor spacing of (340 m/s)/ (2 × 21 kHz) = 1.42 cm. They form a desired frequency-invariant response as Fig. 5 shows.

2.2 The Frequency-Invariant Methods for Near-Fields Wave

Now we discuss the case that aperture is shorter than the wave length ($D < \lambda$). Let us consider a linear equispaced array composed of M omnidirectional pointlike sensors, each connects to an FIR filter composed of L taps. The far-field beamformer response is a function of the direction of arrival and of the frequency and can be expressed as follows

$$\mathrm{BP}(\theta, f; \mathbf{W}) = \sum_{m=0}^{M-1}\sum_{l=0}^{L-1} w_{m,l}\exp\left[-j2\pi f\left(\frac{md\sin\theta}{c} + lT_c\right)\right] \tag{3}$$

where f is the frequency, θ is the arrival angle which belongs to the interval $[-90°, 90°]$, c is the speed of the acoustic waves in the medium, T_c is the sampling interval of the FIR filters,

d is the interelement spacing, and $w_{m,l}$ represents the lth tap coefficient of the mth filter. The L coefficients of the M FIR filters are independently adjustable, resulting in a $M \times L$ matrix denoted by $\mathbf{W}$. The mth line of such a matrix represents the impulse response of the mth FIR filter, and its Fourier transform is the frequency response of the same filter.

In order to directly synthesize the FIR filters, it is necessary to produce an FIBP by using a superdirective array, [10] proposes a stochastic approach based on the simulated annealing (SA) algorithm. Let P be the number of the points used in discretizing the direction-of-arrival axis from $-90°$to $90°$, Q the number of points used in discretizing the frequency axis over the desired bandwidth, and $\mathrm{BP}_{p,q}(\mathbf{W})$ the (p,q)-element of an $P \times Q$ matrix representing the broadband beam pattern in θ_p and f_q, compute by (3), applying the tap coefficients contained in the matrix **W.** a cost function well tailored to our aim can be defined as follows:

$$f(\mathbf{W}) = \sum_{p=0}^{P-1} \sum_{q=0}^{Q-1} a_q \cdot \left| \mathrm{BP}_{p,q}(\mathbf{w}) - \mathrm{BP}d_{p,q} \right| \tag{4}$$

Where $a_q$is the qth element of a $1 \times Q$ vector representing a weighting function for assigning different importance to the matching of the desired beam pattern at specific frequency values. In order to synthesize FIR filters having the same linear phase and to reduce the number of variables to optimize, it necessary to impose the symmetry relation on the weights [10]. To minimize (4), a stochastic approach based on SA has been adopted in [10]. As an example of application of this method, a linear array, made up of eight equally spaced pointlike omnidirectional microphones with a spatial aperture D equal to 12 cm and an interelement spacing d equal to 1.714 cm, has been taken into account. It has been designed to work in the air with acoustic waves traveling at the speed 340 m/s.

In design, each sensor feeds a 70-order FIR filter with a sampling frequency equal to 8 kHz. The frequency interval considered for the design of the FIBP ranges from 513 to 3,591 Hz and is discretized by using $Q = 100$ equally spaced points. The direction-of-arrival angle θ ranges between $-90°$ and $90°$, it is discretized by using $P = 31$ points that are equally spaced in the domain of $\sin\theta$. The steering direction is fixed at $\theta = 0°$. It is very important to note that the ratio D/λ is below unity up to 2,835 Hz and above unity at the higher

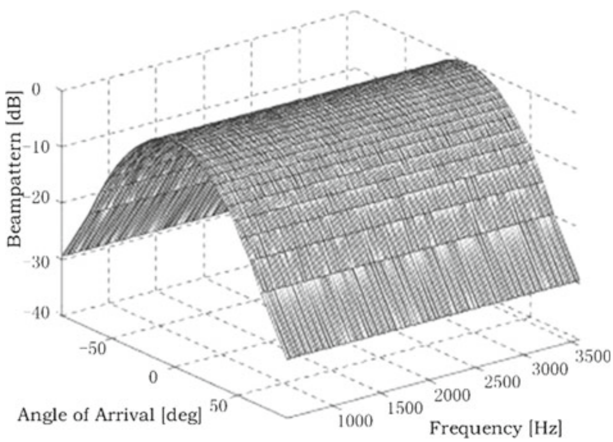


Fig. 6 Desired beam -pattern profile for an eight-sensor linear array with a spatial aperture$D = 0.12$ m working in the air. [10]

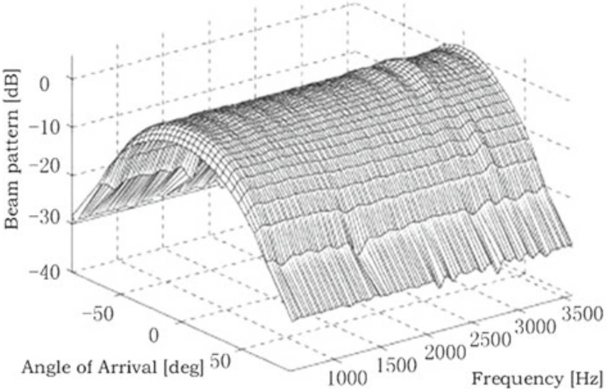


Fig. 7 Desired beam-pattern profile for an eight-sensor linear array with a spatial aperture$D = 0.12$ m working in the air. [10]

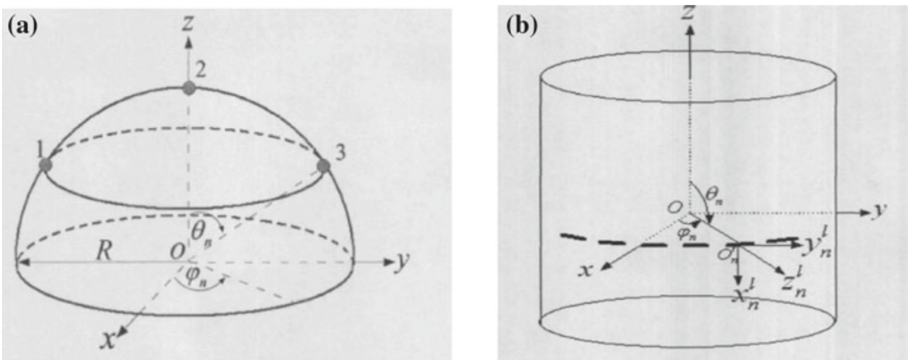


Fig. 8 conformal array, **a** cylindrical array **b** hemispherical array

frequency. Moreover, due to the characteristics of the desired beam pattern shown in Fig. 6, the simulation result is shown in Fig. 7. It clearly finds the pattern is frequency-invariant.

3 The Frequency-Invariant Methods using Multi-Dimensional Arrays

Multi-dimensional array here refers to the two-dimensional or three-dimensional arrays. If elements were placed on the two-dimensional array, it could be called as the planar array antenna. Array elements can also be arranged on a curve or surface. This array is called as the conformal array antenna. Conformal array can overcome the linear array and planar array's shortcomings of less scanning range. It can scan 360° azimuth. Usually, the conformal array antenna includes circular array, cylindrical surface array, hemispherical array, etc. as shown in Fig. 8

3.1 The Frequency-Invariant Methods using Bessel Function

Circular array can consist of either a single ring or multiple concentric rings; the latter is called as concentric ring array (CRA). The configuration of a CRA is shown in Fig. 9.

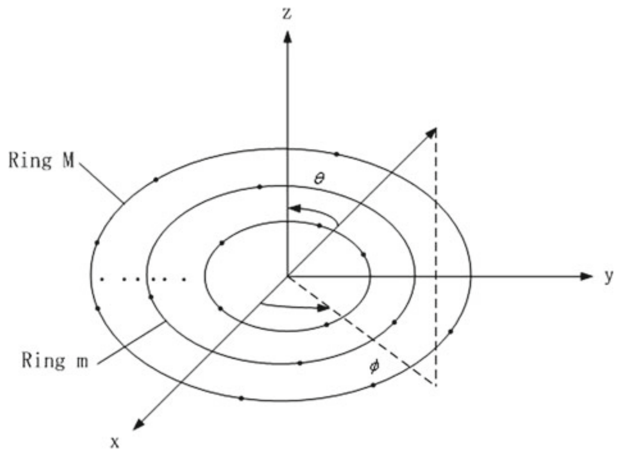


Fig. 9 configuration of a CRA [3]

Where ϕ denotes the azimuth and θ the elevation angle with respect to the z axis. The array consists of M rings, and the numbering of the rings starts from the innermost one. The mth ring has N_m equally spaced array elements. The total number of the array elements is $K = \sum_{m=1}^{M} N_m$.

For a signal impinging on the CRA, the beamformer output at time instant l is the product of weight vectors' complex conjugate transpose and signals vector [3]. The method in [3] does not obtain weight vectors directly, but rather decomposing weight vectors into two different sets of the weights: intra-ring weights and inter-ring weights, that will be chosen differently. It firstly gets the each ring output with the help of intra-ring weights. The final array output of the CRA is obtained as a weighted sum of the output from each ring, inter-weight governs the contribution of the individual rings to the final output.

When the source signal becomes broad-band input, it is usually decomposed into many narrow-band components through FFT, and different frequency components are processed by narrow-band beamformers.

In the exemplary CRA, the highest operating frequency is set to $f_0 = 2{,}000$ Hz, and the number of rings M in the CRA is chosen to be 12. Assuming the sound speed is 343 m/s. The largest radius is 0.5 m, and the smallest one is 0.03 m. The number of the array elements on each ring is $N = [5\ 8\ 11\ 15\ 18\ 21\ 24\ 27\ 30\ 34\ 37\ 40]$, from the innermost to the outermost ring. The total number of the array elements is 270. The 3-D beampattern obtained from this method is shown in Fig. 10.

3.2 The Frequency-Invariant Methods Using Bessel Function and Phase Mode

Chan and Chen proposed a digital beamformer for uniform concentric circular arrays (UCCAs), the obtained pattern had nearly frequency-invariant (FI) characteristics [12,13]. The basic principle was to transform the received signals to the phase mode and removed the frequency dependency of the individual phase mode through the use of a digital beamforming or compensation network.

Figure 11 shows a UCCA with P rings and each ring has K_p omnidirectional sensors located at $\{r_p \cos\phi_{k_p}, r_p \sin\phi_{k_p}\}$, where r_p is the radius of the pth ring, $p = 1, \ldots, P$, $\phi_{k_p} = 2\pi k_p/K_p, k_p = 0, \ldots K_p - 1$. In UCCAs, the intersensors spacing in each ring is

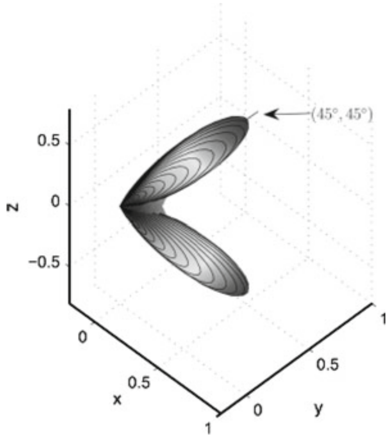


Fig. 10 3-D array pattern at 2,000 Hz; the look direction is $(45^0, 45^0)$ [3]

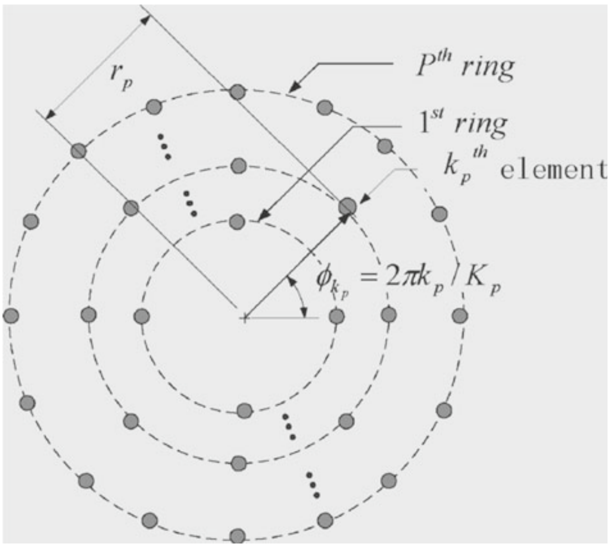


Fig. 11 UCCA with P rings, the pth ring has K_p equally spaced sensors [13]

fixed at $\lambda/2$, where λ is the smallest wavelength of the array to be operated and is denoted by $\lambda_{\min}$. The radius of the pth ring of the UCCA is given by $r_p = \frac{\lambda_{\min}}{\left(4\sin\left(\frac{\pi}{K_p}\right)\right)}$. For convenience, this radius is represented as its normalized version $\tilde{r}_p = r_p/\lambda_{\min} = 1/(4\sin(\pi/K_p))$.

Let α denote the ratio of the sampling frequency f_s to the maximum frequency $f_{\max}(\alpha = f_s/f_{\max})$, the phase difference between the k_p^{th} sensor and the center of the UCCA is$\chi_{k_p} = 2\pi\tilde{r}_p\alpha\sin\theta\cos(\phi - \phi_{k_p})$. For notation convenience, let $\hat{r} = \tilde{r}\cdot\alpha$. The corresponding frequency response of the phase difference can be expressed as $e^{jw\hat{r}_p\sin\theta\cos(\phi-\phi_{k_p})}$, where w is the frequency variable, ϕ and θ are the azimuth angle and the elevation angle respectively. Figure 12 shows the structure of the FI beamformer for a UCCA with P rings. After appropriate downconversion, lowpass filtering, and sampling the sampled signals from the antennas of the pth ring, which is called a

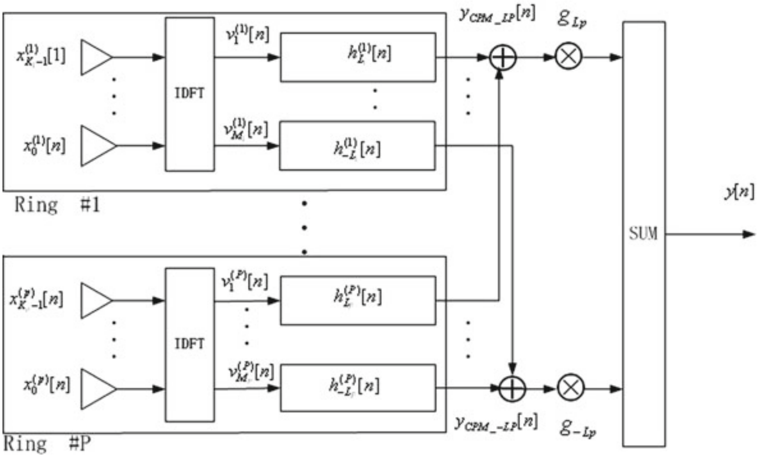


Fig. 12 UCCA-FIB block diagram [13]

snapshot at sampling instance n. This snapshot is IDFT to a set of Fourier coefficients; each coefficient is called a phase mode. Each branch of the IDFT output is then filtered (to compensate for the frequency dependency of the phase mode), multiplied with the variable beamformer weights before combining to give the beamformer output. To obtain the spatial-temporal transfer function of the beamformer, it necessary to assume that there is only one source signal $s(n)$ with spectrum $S(w)$. Taking the discrete-time Fourier transform (DTFT) of transformed snapshot and the beamformer output. Compare these two transform, it can get the spatial-temporal response of the pth ring of the UCCA. To obtain an FI response, the spatial-temporal response of the pth ring of the UCCA which is a function of both ϕ and w, should be independent of the frequency variable w. In a UCCA, the outer rings have more phase modes than the inner ones. For simplicity, it necessary to assume that the weighting vectors of the rings are identical. To this end, after making use of the Jacobin-Anger expansion and the property of Bessel functions [12], it gets the FI response.

3.3 The Frequency-Invariant Methods for Conformal Array Antenna

Wang et al. [4] proposed a uniform method for the element polarized pattern transformation of arbitrary 3D conformal arrays and a space-time-polarization filter structure (as shown in Fig. 13). This method established a unified frame for the frequency-invariant and low cross-polarization pattern synthesis of conformal array. Simulation results demonstrated its effectiveness.

Consider the space-time-polarisation filter structure, as shown in Fig. 13. Here the FIR-based temporal is used to acquire frequency-invariant pattern, and the polarization diversity in array global coordinates is used to deal with the serious cross-polarisation effect of conformal arrays.

This method is first established with the application of the Euler rotation. The rotation transformations of the polar coordinates of the global array to the local elements and the element polarized pattern rotation of the local coordinates to the global coordinates are realized subsequently. The Euler rotation method is verified with the simulation of the inherent shield

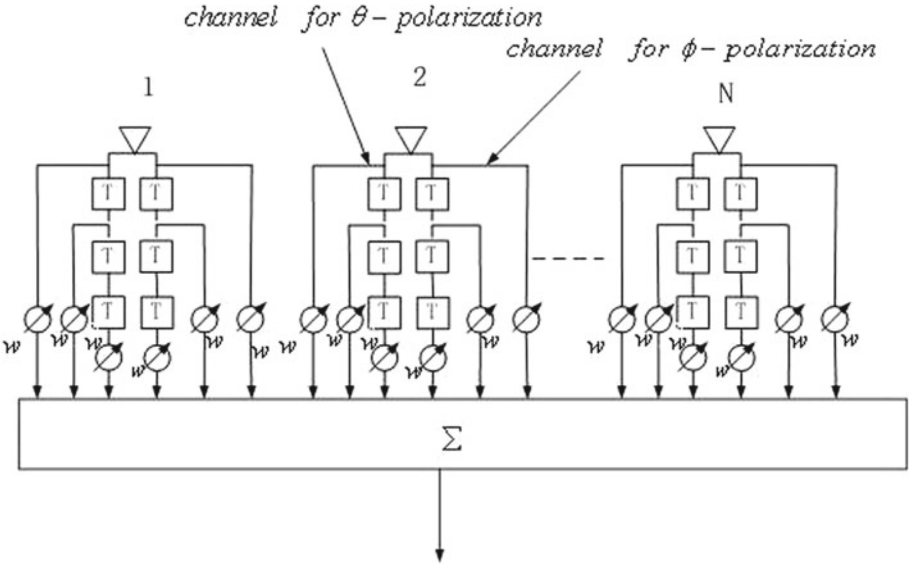


Fig. 13 Space-time-polarisation filter structure for pattern synthesis of conformal array [4]

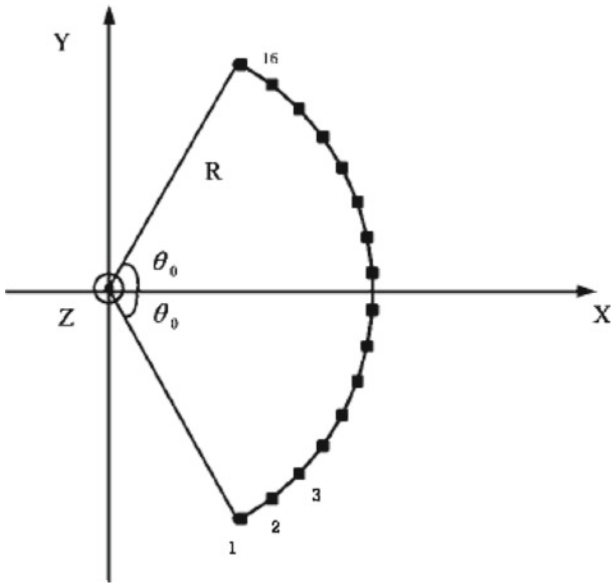


Fig. 14 Top view of the cylindrical array of 16 elements [4]

effect of the conformal cylindrical array. Despite of its superior performance, the presents filter structure leads to higher system complexity.

Consider an array of 16-element ring array that conforms to a cylindrical surface, as shown in Fig. 14. The element of the ring array is indexed by *m* counter-clockwise.

In simulation, the distance between the elements is set as $0.5\lambda_0$ and $\theta_0 = 60°$. The target pattern is defined by the 3 dB main lobe beam width (15°), side lobe level (−30 dB) and

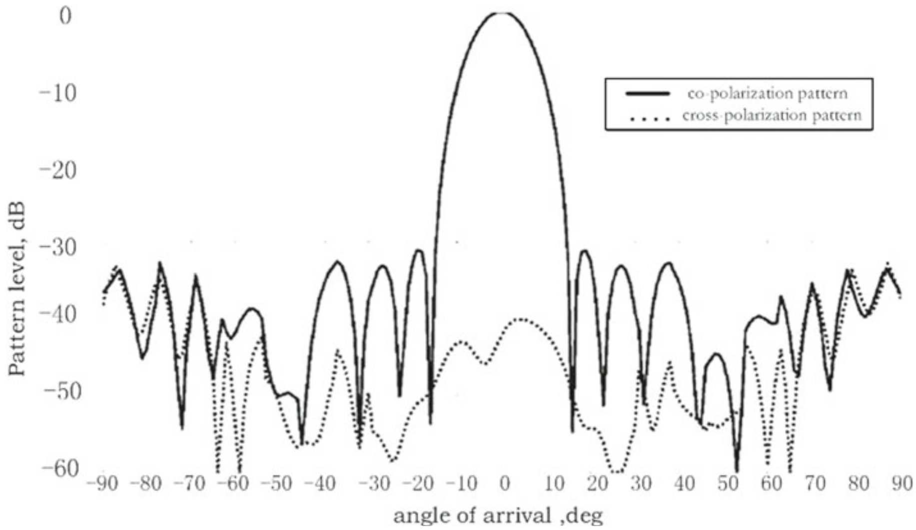


Fig. 15 Array pattern at $0.8f_0$ [4]

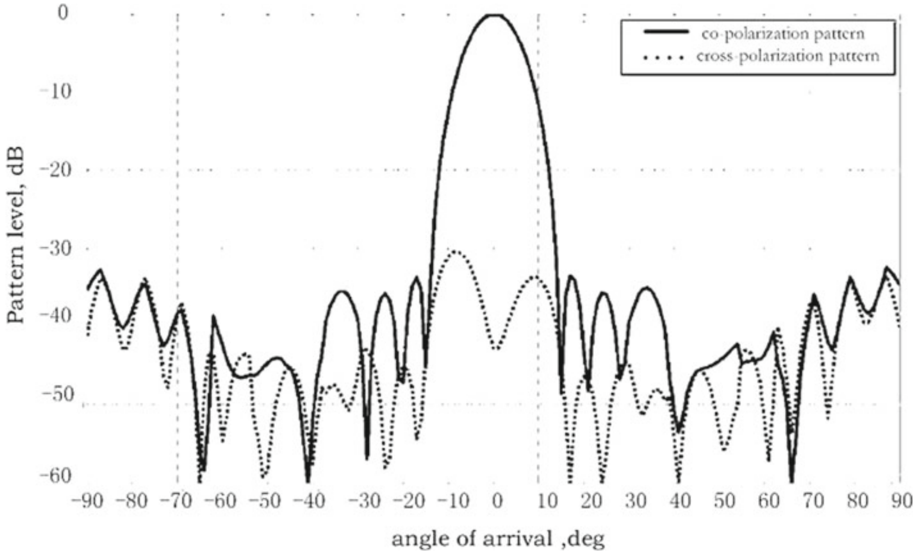


Fig. 16 Array pattern at $0.9f_0$ [4]

cross-polarisation level (−30 dB). The array pattern is optimized for frequency range of 0.8 f_0–f_0. The FIR filter with $M = 5$ is used for each element and the sample rate is set to be f_0. After about 20 iterations, the resulting patterns are shown in Figs. 15, 16, 17, and 18. Figures 15, 16, and 17 show the pattern obtained at $0.8f_0$, $0.9f_0$ and f_0, respectively. Figure 18 shows the co-polarisation pattern for the bandwidth $[0.8f_0, f_0]$.

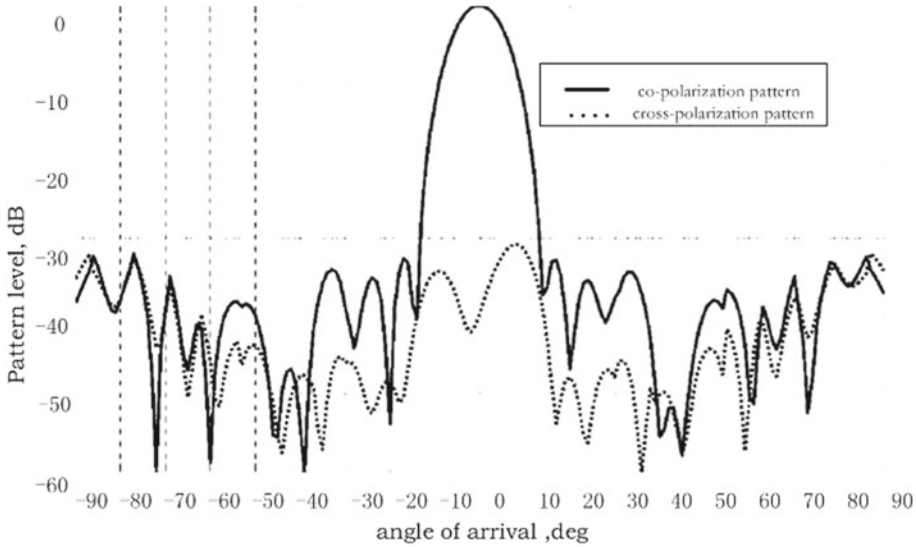


Fig. 17 Array pattern at f_0 [4]

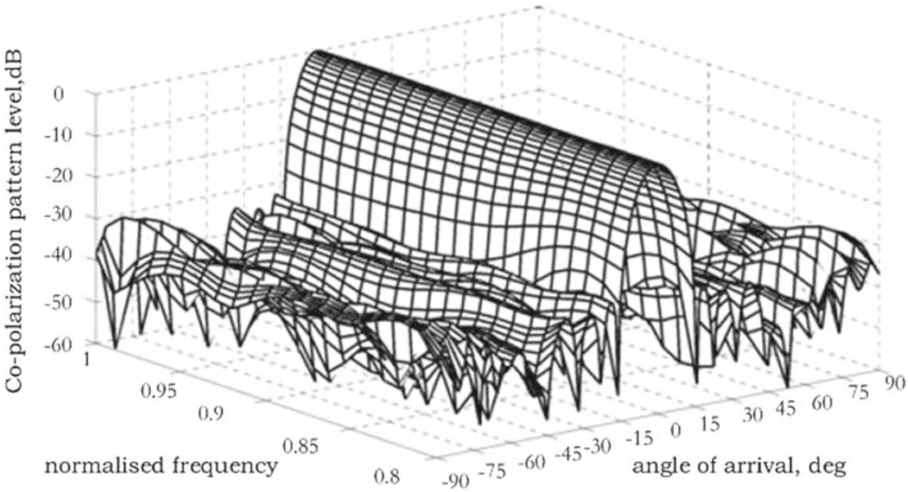


Fig. 18 Cross-polarisation array pattern for bandwidth $[0.8f_0, f_0]$ [4]

4 Conclusions

Based on the array structure, this paper grouped the methods of FIB into two kinds: one-dimensional array and multi-dimensional array, where the details were discussed. Generally, all of these methods are proposed in order to achieve FIB, but most run on linear array due to its easy operation while it is difficult for circular array at the lower frequency. Conformal array antennas have advantages of reducing aerodynamic drag, broad angle coverage and space-saving, which can be applied in a variety of fields such as space-borne, airborne, ship-borne, missile-borne radar, space vehicles and wireless communication.

Acknowledgments This work was supported by the National Natural Science Foundations of China (No.60773203), and the open research fund of National Mobile Communications Research Laboratory, Southeast University (No. W200815).

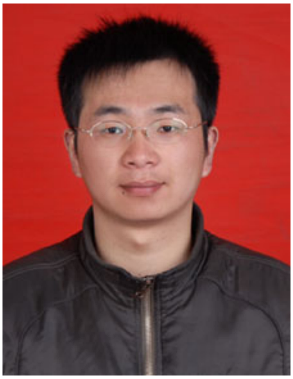
References

1. Widrow, B., Mantey, P. E., Griffiths, L. J., & Goode, B. B.(1967). Adaptive antenna systems. *Proceedings of the IEEE, 55*(12), 2143.
2. Frost, O. L.(1972). An algorithm for linearly constrained adaptive array processing. *Proceedings of the IEEE, 60*(8), 926–935.
3. Yunhong, L., Ho, K. C., & Kwan, C. (2006). Array pattern synthesis with frequency invariant property for concentric ring array. *IEEE Transactions on Signal Processing, 54*, 780–784.
4. Wang, B. H., Guo, Y., Wang, Y. L., & Lin, Y. Z. (2008). Frequency-invariant pattern synthesis of conformal array antenna with low cross-polarisation. *Microwaves, Antennas & Propagation, IET, 2*, 442–450.
5. Do-Hong, T., & Russer, P.(2004). Signal processing for wideband smart antenna array applications. In *IEEE microwave magazine* (pp. 57–67).
6. Ward, D. B., Kennedy, R. A., & Williamson, R. C. (1996). FIR filter design for frequency invariant beamformers. *Signal Processing Letters, IEEE, 3*, 69–71.
7. Liu, W., McLernon, D. C., & Ghogho, M.(2009). Design of frequency invariant beamformer without temporal filtering. *IEEE Transactions on Signal Processing, 57*(2), 798–802.
8. Wei, L., & Weiss, S. (2008). Design of frequency invariant beamformers for broadband arrays. *IEEE Transactions on Signal Processing, 56*, 855–860.
9. Dotlic, I. D. (2004). Minimax frequency invariant beamforming. *Electronics Letters, 40*, 1230–1231.
10. Trucco, A., Crocco, M., & Repetto, S.(2006). A stochastic approach to the synthesis of a robust frequency-invariant filter-and-sum beamformer. *IEEE Transactions on Instrumentation and Measurement, 55*(4).
11. Repetto, S., & Trucco, A.(2006). Designing superdirective microphone arrays with a frequency-invariant beam pattern. *IEEE Sensors Journal, 6*(3).
12. Chan, S. C., & Chen, H. H. (2007). Uniform concentric circular arrays with frequency-invariant characteristics—Theory, design, adaptive beamforming and DOA estimation. *IEEE Transactions on Signal Processing, 55*, 165–177.
13. Chen, H. H., Chan, S. C., & Ho, K. L. (2007). Adaptive beamforming using frequency invariant uniform concentric circular arrays. *IEEE Transactions on Circuits and Systems I: Regular Papers, 54*, 1938–1949.
14. Dotlic′, I. D., & Zejak, A. J. (2001). Arbitrary antenna array pattern synthesis using minimax algorithm. *Electronics Letters, 37*(4), 206–208.
15. Zrnic′, B. M. (1998). *Range sidelobe suppression for pulse compression radars utilizing modified RLS algorithm.* In *Proceedings of the IEEE ISSSTA'98, IEEE 5th International Symposium on Spread Spectrum Techniques and Applications* (pp. 1008–1011). Sun City, South Africa.

Author Biographies

Ning Xie received the B.E., and Ph.D. degrees in communications and information system from Sun Yat-Sen University, China, in 2002, and 2007, respectively. He joined the College of Information Engineering, Shenzhen University, Guangdong, China, in July 2007. His research interests include adaptive array processing, array pattern synthesis, communications circuits and wireless communications.

Hui Wang received his B.S., M.S. and Ph.D. degrees from Xi'an Jiaotong University, in 1990, 1993, and 1996, respectively, all in telecommunication. He is now a professor and dean of the College of Information Engineering, Shenzhen University. His research interests include wireless communication, signal processing, and distributed computing systems, in which, he is the author or co-author of more than 50 international leading journals, conferences and book chapters. He is a member of IEEE.



Hongwei Liu received the B.S. degree in communications engineering from Hunan University of Arts and Science, China, in 2008. He is pursuing Master's degree in Shenzhen University, Guangdong, China; his research interests include adaptive array processing, array pattern synthesis and wireless communications.