Antenna Array Beamforming Technology: Enabling Superior Aeronautical Communication Link Performance

Cheng Y. Lu*, Yimin Zhang, Jinsong Wu, Paul Cook*, Xin Li, and Moeness Amin

*Teletronics Technology Inc., Newtown, PA 18940
Center for Advanced Communications, Villanova University, Villanova, PA 19085

ABSTRACT

In this paper, we propose the exploitation of array beamforming technology in high-speed aeronautical communication applications, e.g., the integrated Network Enhanced Telemetry (iNET) system. By flexible steering of beams and nulls, an array can enhance desired signals whereas the undesired signals such as interference and jammers are suppressed. The proposed adaptive beamforming technology is DSP-based and network-aware, and is designed for the use at aerial vehicle platforms to increase transmission power efficiency, improve receiving signal sensitivity, mitigate interference/multipath effects, and extend the communication range.

KEY WORDS

Aeronautical communications, Beamforming, iNET, DSP, Multipath propagation.

1 INTRODUCTION

In this paper, we propose the exploitation of array beamforming technology in high-speed aeronautical communication applications where our goal is to maintain reliable and critical communications integrity and adequate signal-to-noise ratio (SNR) levels. The network architecture, e.g. iNET, adopts a network-based architecture for use in weapons and aircraft systems test and evaluation. Array beamforming techniques [1, 2, 4, 7] have been widely used in wireless communications, underwater acoustics, and radar systems for a variety of reasons. By flexible steering of beams and nulls, an array can enhance desired signals whereas the undesired signals such as interference and jammers are suppressed. The proposed adaptive beamforming technology is DSP-based and network-aware, and is designed for the use at aerial vehicle (AV) platforms to increase transmission power efficiency, improve receiving signal sensitivity, mitigate interference/multipath effects, and extend the communication range.

The aeronautical communications bring unique challenges as well as opportunities to the proposed beamforming technology. The major challenges lie in the hostile environment, highly dynamic mobility, and weak signal strength. In such situation, adaptive updates of the array weights require high complexity and sometimes may not be affordable for real-time optimization. Furthermore, determination of adaptive weights for transmit beamforming becomes even challenging since the channels cannot be reliably estimated. In the proposed beamforming method, therefore, we take different approaches to utilize the location information of the AV and the ground station (GS) as well as the flight dynamics known to the AV to support environment-aware beamforming. Array patterns are formed in real-time and are used for both transmission and receiving. The beamformer is designed to be robust against errors in the location and flight dynamics information.

The proposed beamforming technique achieves multifold advantages: (1) By using the proposed technique, the high-speed communication system can provide wireless communication links with improved link reliability and power efficiency. (2) The high-speed communication system can maintain a wireless link significantly beyond the maximum range supported by single-antenna based system. (3) The spatial selectivity allows the high-speed communication system to reduce sensitivity to multipath and jammers, and lower the transmit power in the direction of unintended directions to yield low probability of intercept (LPI).

2 BEAMFORMING

2.1 Concept

Beamforming may be performed through a weighted combination of radio signals from an array of antennas. Consider a receiving array, beamforming is to create the radiation pattern of the antenna array by adjusting the weights corresponding to each antenna output such that the phases of the signals in the direction of the desired sources/targets are added constructively, whereas they are nullified/mitigated in the directions of undesired/interfering sources/targets [1, 2].

In beamforming, both the amplitude and phase of each antenna element may be controlled [5, 6]. The combined relative amplitude and phase shift for each antenna is called a "complex weight." A beamformer for a radio transmitter applies the complex weight to the transmit signal (shifts the phase and sets the amplitude) for each element of the antenna array, then sums all of the signals into one that has the desired directional pattern [1, 2]. In digital beamforming, the operations of phase shifting and amplitude scaling for each antenna element, and summation for receiving, are done digitally. Either general-purpose DSP's or dedicated beamforming chips may be used. Digital processing requires that the signal from each antenna element is digitized using an A/D converter [3]. When radio frequencies are too high to be directly digitized at a reasonable cost, digital beamforming receivers use analog RF down converters to shift the signal frequency down before the A/D converters.

Once the antenna signals have been digitized, they are passed to digital down-converters that shift the radio channel's center frequency down and pass only the bandwidth required for one channel. The down-converters produce a "quadrature" baseband output at a low sample rate. The quadrature baseband I and Q components can be used to represent a radio signal as a complex vector (phasor) with real and imaginary parts. For beamforming, the complex baseband signals are multiplied by the complex weights to apply the phase shift and amplitude scaling required for each antenna element. A general-purpose DSP can implement the complex multiplication for each array element [3].

All RF and A/D converters share common oscillators so that they all produce identical phase shifts of the signal. Within the digital beamformer, all digital down-converters share a common clock, set for the same center frequency and bandwidth, and their digital local oscillators are in-phase so that all phase shifts are identical. In digital beamforming receiver, the complex weights for the antenna elements are carefully chosen to give the desired peaks and nulls in the radiation pattern of the antenna array. In beamforming for communications, the weights are chosen to give a radiation pattern that maximizes the quality of the received signal. Usually, a peak in the pattern is pointed to the signal source and nulls are created in the directions of interfering sources and signal reflections [2].

2.2 Beamforming Algorithms

Consider that a receiver antenna array, consisting of N elements, is located at the AV. The coordinate of the *l*-th array element is denoted as (x_l, y_l, z_l). For convenience, we also define the AV reference point as the origin of the coordinates.

While the AV reference point can be arbitrarily chosen within or around the array, we assume it is located at the gravity center of the array.

The distance between the ground station (GS), located at (x_g, y_g, z_g), and the AV reference point, located at $(0,0,0)$, is expressed as

$$r = \sqrt{(x_g)^2 + (y_g)^2 + (z_g)^2} \tag{1}$$

Denote $\mathbf{u}$ as the unit vector in the direction from the GS transmitter to the AV reference point. It can be expressed as

$$\mathbf{u}_g = \left[\frac{x_g}{r}, \frac{y_g}{r}, \frac{z_g}{r}\right] = \left[\sin q_g \cos f_g, \sin q_g \sin f_g \cos q_g\right] \tag{2}$$

where q_g and f_g is the elevation and azimuth angle of the GS with respect to the AV reference point.

Let $\tilde{s}(t)$ be the RF signal waveform transmitted from the GS. The signal received at the *l*-th antenna element of an array at the AV is expressed as

$$\tilde{r}_l(t) = a_l \tilde{s}(t - t_l) g_l(\mathbf{u}_g)) + \tilde{n}_l(t) \tag{3}$$

where a_l and t_l, respectively, denote the channel gain and time delay of the channel between the GS transmitter and AV receiver, and $g_l(\mathbf{u}_g)$ denotes the antenna gain of the *l*-th array element in the direction of the GS. In addition, $\tilde{n}_l(t)$ is the additive noise.

Note that, while a_l, t_l, and $\mathbf{u}_g$ gradually changes over time as the AV maneuvers, we omit (t) for notation simplicity.

Without loss of generality, we can conveniently define that t is referenced to the waveform received at the AV reference point (that is, the time delay evaluated at the AV reference point is zero). In addition, the GS is typically considered to be at the far field of the AV, that is, the distance between them is much larger than the dimension of the GS antenna and AV antenna array. In this case where multipath fading is not considered, the channel gain can be assumed to satisfy $a_1 = ... = a_N = a$, and the time delay at the l-th array element can be expressed as

$$t_l = \frac{\mathbf{r}_l^T \mathbf{u}_g}{c} \tag{4}$$

where $\mathbf{r}_l = [x_l, y_l, z_l]$, $c \approx 3\times10^8$ m/s is the speed of wave propagation, and ^{T} denotes transpose.

When $\tilde{s}(t)$ is a narrowband signal (i.e. the signal bandwidth BW is much smaller than the carrier frequency), the following (approximate) relationship holds (with the use of proper complex signal notation),

$$\tilde{s}(t - t_l) \approx \tilde{s}(t) e^{-jw_c t_l}, \tag{5}$$

where $w_c = 2p f_c$ is the radian carrier frequency. In this case, we can define $s(t)$ as the equivalent baseband waveform of $\tilde{s}(t)$, and the equivalent baseband signal received at the l-th array element becomes

$$r_l(t) = a_l s(t) e^{-jw_c t_l} g_l\left(\mathbf{u}_g\right) + n_l(t) \tag{6}$$

Therefore, the equivalent baseband signals received at the N elements can be expressed in the following vector format

$$\mathbf{r}(t) = \begin{bmatrix} r_1(t) \\ \vdots \\ r_N(t) \end{bmatrix} = s(t) \begin{bmatrix} a e^{-jw_c t_1} g_1(\mathbf{u}_g) \\ \vdots \\ a e^{-jw_c t_N} g_N(\mathbf{u}_g) \end{bmatrix} + \begin{bmatrix} n_1(t) \\ \vdots \\ n_N(t) \end{bmatrix} = s(t)\mathbf{h}(\mathbf{u}_g) + \mathbf{n}(t). \tag{7}$$

Beamforming is achieved by the weighted summation of the signal received at the N elements. Let $\mathbf{w} = [w_1, ..., w_N]$ be the weights corresponding to the N array elements, the array output is obtained as

$$y(t) = \mathbf{w}^T \mathbf{r}(t) \mathbf{h}(\mathbf{u}_g) + \mathbf{w}^T \mathbf{n}(t) \tag{8}$$

When the elements of $\mathbf{n}(t)$ are independent and identically distributed (i.i.d.) zero-mean complex Gaussian processes with covariance matrix $s_n^2 \mathbf{I}_N$, where $\mathbf{I}_N$ denotes the $N \times N$ identity matrix, the maximum signal-to-noise ratio (SNR) is achieved when

$$\mathbf{w} = a \mathbf{h}^*(\mathbf{u}_g) \tag{9}$$

where a is a constant, and $(\cdot)^*$ denotes complex conjugate operation. Note that a does not affect the SNR because it is multiplied to both the signal and noise. However, it is often convenient to assume a unit norm of $\mathbf{w}$, that is, $a = \left|\mathbf{h}^*(\mathbf{u}_g)\right|^{-1}$, where | . | denotes the norm of a vector. That is,

$$\mathbf{w} = \frac{\mathbf{h}^*(\mathbf{u}_g)}{\left|\mathbf{h}^*(\mathbf{u}_g)\right|}. \tag{10}$$

In this case, the output noise power remains the same as that of a single antenna case, that is,

$$P_n = E\left[\left|\mathbf{w}^T\right|^2\right] = s_n^2 \tag{11}$$

with $E[.]$ representing statistical expectation. The output signal power becomes

$$P_s = \sum_{l=1}^{N}\left|a g_l\left(\mathbf{u}_g\right)\right|^2 s_s^2, \tag{12}$$

where $s_s^2 = E\left[\left|s(t)\right|^2\right]$.

In particular, when all the antennas are isotropic, $g_1\left(\mathbf{u}_g\right) = ... = g_N\left(\mathbf{u}_g\right) = g\left(\mathbf{u}_g\right)$, we have

$$P_s = N\left|a g\left(\mathbf{u}_g\right)\right|^2 s_s^2, \tag{13}$$

which is N times higher than the received signal power obtained from a single receive antenna. That is, an array gain of N (that is, $10\log_{10}(N)$ dB) is achieved.

2.3 Array Factor

To signify the array process gain in different directions, it is often useful to analyze the array factor, which does not include array element patterns. The array signature corresponding to the direction $\mathbf{u}_g$ is characterized by

$$\mathbf{h}(\mathbf{u}_g) = \begin{bmatrix} e^{-jw_c t_1} \\ \vdots \\ e^{-jw_c t_N} \end{bmatrix} = \begin{bmatrix} e^{-jk_c \mathbf{r}_1^T \mathbf{u}_g} \\ \vdots \\ e^{-jk_c \mathbf{r}_N^T \mathbf{u}_g} \end{bmatrix} \tag{14}$$

where $k_c = w_c / c = 2p / l$. Note that the norm of $\mathbf{h}(\mathbf{u}_g)$ in this case is $\sqrt{N}$.

Similarly, the array signature corresponding to an arbitrary direction $\mathbf{u}$=[sin$q$ cos$f$, sin$q$ sin$f$, cos$q$] can be expressed as

$$\mathbf{h}(\mathbf{u}) = \begin{bmatrix} e^{-jk_c \mathbf{r}_1^T \mathbf{u}} \\ \vdots \\ e^{-jk_c \mathbf{r}_N^T \mathbf{u}} \end{bmatrix} \tag{15}$$

The array factor is typically expressed as a function of q and f. When $\mathbf{w} = \frac{1}{\sqrt{N}}\mathbf{h}^*(\mathbf{u}_g)$, the array factor is obtained as

$$P(q,f) = \frac{1}{\sqrt{N}}\mathbf{h}^H(\mathbf{u}_g)\mathbf{h}(\mathbf{u}) = \frac{1}{\sqrt{N}}\sum_{l=1}^{N} e^{jk_c\left(\mathbf{r}_l^T\mathbf{u}_g - \mathbf{r}_l^T\mathbf{u}\right)}, \tag{16}$$

where $(\cdot)^H$ denotes conjugate transpose.

To evaluate the array factor at a different frequency, say, $f_c = w/(2p)$, where the weight vector remains to be $\mathbf{w} = \frac{1}{\sqrt{N}}\mathbf{h}^*(\mathbf{u}_g)$, obtained in f_c, we modify equation (15) as

$$\mathbf{h}(\mathbf{u},w) = \begin{bmatrix} e^{-jk_c\mathbf{r}_l^T\mathbf{u}} \\ \vdots \\ e^{-jk_c\mathbf{r}_l^T\mathbf{u}} \end{bmatrix}. \tag{17}$$

Thus, the array factor becomes

$$P(q,f,w) = \frac{1}{\sqrt{N}}\mathbf{h}^H(\mathbf{u}_g)\mathbf{h}(\mathbf{u},w) = \frac{1}{\sqrt{N}}\sum_{l=1}^{N} e^{j\left(k_c\mathbf{r}_l^T\mathbf{u}_g - k\mathbf{r}_l^T\mathbf{u}\right)}, \tag{18}$$

where $k = w/c$.

While we have used the array for receive beamforming, the array performance, in terms of the array gain and array patterns, remains unchanged when the array is used for transmit beamforming.

3 NUMERICAL RESULTS

To demonstrate the effectiveness of the beamforming techniques, numerical calculations are performed. A rectangular array of four elements is considered (Fig. 1). The distance of any of two adjacent array elements is denoted as $D_a = 0.5l$. It can be shown that, a 25 MHz bandwidth has a negligible effect in the array gain when the system is operated in 2.3 GHz, so we focus only on narrowband array processing hereafter.

3.1 Beamforming Performance with Identical Antenna Patterns

For simplicity, monopoles with a proper ground plane are used to demonstrate the beamforming performance with the use of directional antennas. For a vertically polarized monopole, located in the *z*-direction, the normalized half-wavelength antenna pattern is expressed as

$$P(q,f) = \frac{\cos\left(\frac{p}{2}\cos q\right)}{\sin q}, \tag{19}$$

where $0 \le q \le 90^{o}$ for the array located on the top of the aircraft, or $90^{o} \le q \le 180^{o}$ for the array located in the bottom. The normalized antenna pattern is used so that we can focus on the contribution of the array. The antenna patterns, shown in the following discussion, are the joint patterns of the top and bottom monopole based arrays, which are equivalent to the respective dipole based array patterns.

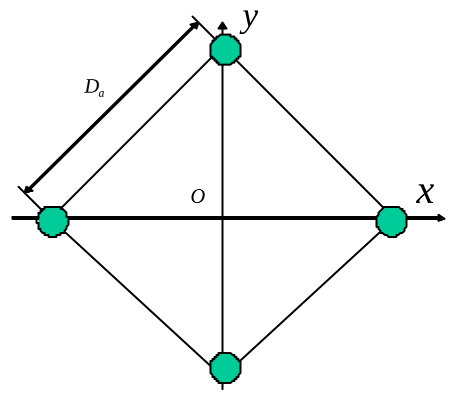
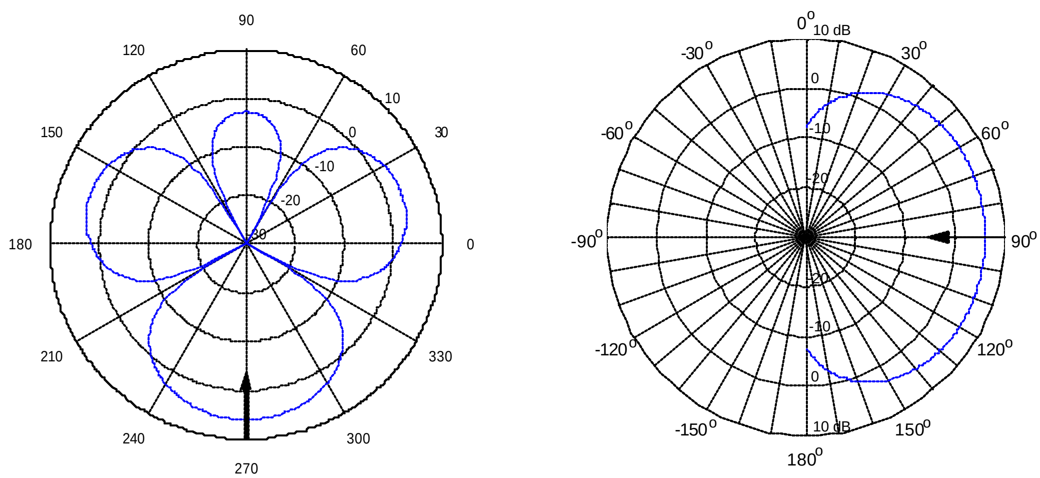


Figure 1 Rectangular array configuration.

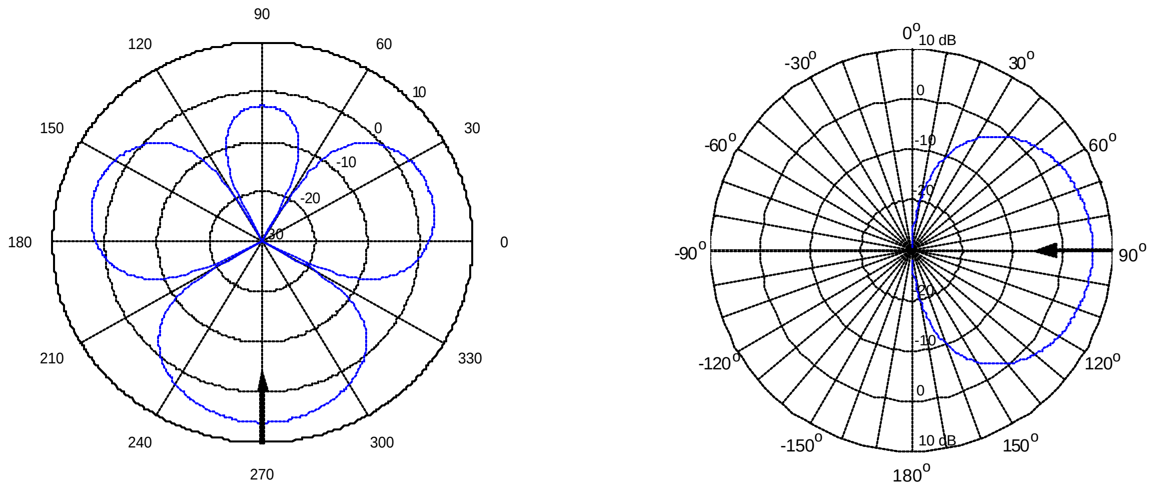


(a) f -plane array pattern (b) q -plane array pattern

Figure 2 f -plane and q -plane array patterns (DOA: $q_g = 90^{o}$ and $f_g = 270^{o}$).

Fig. 2 provides an example of the f-plane and q-plane array pattern. The array factor in the maximum direction is $10\log_{10}(4) \approx 6.02$ dB. The same array factor can be achieved by using other array configurations with identical antenna elements, but the beamwidth and sidelobe levels may vary for different array configurations.

The overall radiation of the array in this case is the product of the array factor and the antenna pattern of each monopole. Figure 3 shows the f -plane and q -plane array patterns. The DOA of the GS is assumed to be $q_g = 90^{o}$ and $f_g = 270^{o}$. In this case, an array gain of 6.02 dB is achieved. the f -plane pattern is the same as the array factor as the dipole is omni-directional in the f -plane, whereas the q -plane pattern is attenuated by the antenna pattern for $q \neq 90^{o}$. In particular, a null in $q = 0^{o}$ or $q = 90^{o}$ is observed as determined by the monopole antenna pattern.



(a) f -plane array pattern (b) q -plane array pattern

Figure 3 f -plane and q -plane array patterns (DOA: $q_g = 90^{o}$ and $f_g = 270^{o}$).

3.2 Beamforming Patterns with Different Antenna Patterns

The array can be flexibly designed to use antenna with different antenna patterns, such as different types of antennas or identical antennas but oriented differently. In this case, the array gain may be reduced from $10\log_{10}(N)$ dB, however, the array may achieve some desirable beamforming characteristics and/or antenna conformability.

In the following, we show an illustrative scenario, where an array consists of four monopoles in different directions, is used to remedy the null direction problem, as we discussed in the previous subsection. In this array, we use the same array geometry as in Figure 1. However, in this case, three monopoles are vertically polarized, whereas the other one is horizontally oriented. This may happen, for example, that the antennas are located in different sides of an aircraft. For simplicity, the edge effect of the ground plane is ignored.

Figure 4 shows the f -plane and q -plane array patterns when the DOA of the signal wave from GS is located at $q_g = 90^{o}$ and $f_g = 270^{o}$ and three antennas are effective, and an array gain of $10\log_{10}(3) \approx 4.77$ dB is achieved. That is, there is a loss of about 1.25 dB in the array gain. On the other hand, when the DOA becomes $q = 0^{o}$, one antenna is effective. As shown in Figure 5, the array gain is $10\log_{10}(1) = 0$ dB, which is compared to 0 (i.e., $-\infty$ dB) in the array using four identical monopoles.

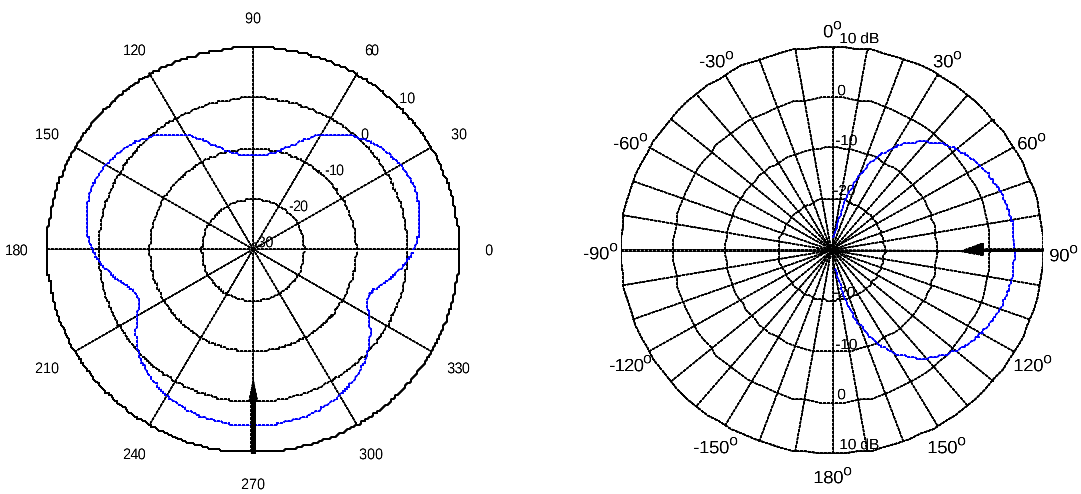
3.3 Important Remarks

Due to space limitations, the following issues are briefly summarized.

1. We have examined different array configurations. As we discussed in Section 3.1, all the arrays achieve the same array gain, provided that the antenna orientations are the same for the different arrays, and the mutual coupling effect is ignored. However, the beamwidth and sidelobe levels vary for different array configurations. Thus, the actual

array configuration is designed to afford optimized performance to meet the beamwidth and sidelobe requirements and to minimize the performance degradation in the presence of various parameter errors, which may occur, for example, due to the measurement inaccuracy and/or limited update rate of the measurement parameters.

2. The array weights can be dynamically updated to change the directivity suited in different scenarios. For example, when the aircraft is closer to the GS and has a low altitude, the heading information may vary much more rapidly. In this case, a broader beam may be formed in the vicinity of the GS so as to provide robust beamforming even when the heading information is updated in a low rate.
3. It is evident from the above discussions and simulation examples that the maximum array factor is achieved in the direction associated with the GS. In comparison, the array factor in other directions is relatively attenuated. Therefore, compared to single antenna system, the effect of multipath interference and/or jamming can be mitigated even without their a priori information or sophisticated interference mitigation.
4. Because the array beams have reasonable beamwidth, the array gain is not very sensitive to small variations in the position or orientation of the AV, or the estimation errors of such information. For example, the array gain corresponding to a few degrees of change in the orientation is insignificant. Thus, the requirement for the speed of updating array weights, even for a highly maneuvering AV, is not demanding.
5. Array configuration is usually optimized in terms the wavelength at the operating frequency. Thus, an array configuration optimized for one frequency band may not be necessarily optimal for another frequency band. An array configuration designed for a higher frequency band, however, often can provide satisfactory performance for a lower frequency band with several percent frequency differences.



(a) f -plane array pattern (b) q -plane array pattern

Figure 4 f -plane and q -plane array patterns (DOA: $q_g = 90^{o}$ and $f_g = 270^{o}$)

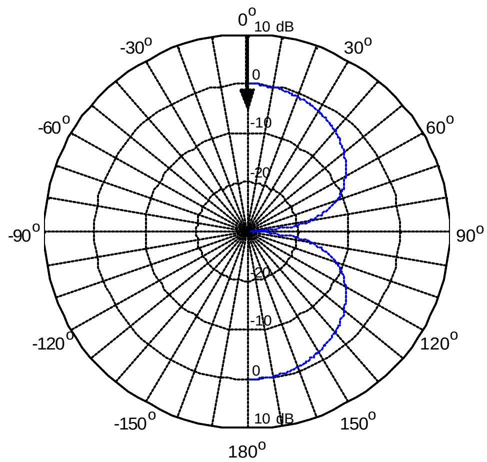


Figure 5 θ -plane array patterns (DOA: $\theta_g = 0^\circ$)

4. CONCLUSION

In this paper, we have proposed the use of array beamforming technology in high-speed aeronautical communication applications, such as the integrated Network Enhanced Telemetry (iNET) system. The theoretical background as well as simulation results were provided. It has been shown that array beamforming can be flexibly designed to maximize the array gain up to the number of array antennas as well as to provide robust wireless links in varying operating scenarios. The array gain achieved in the proposed techniques permits reduced transmit power and/or extended communication range. The spatial selectivity offered by the array pattern provides additional protection against multipath interference and/or jamming.

REFERENCES

[1] R. A. Monzingo and T. W. Miller, *Introduction to Adaptive Arrays*, John Wiley, 1980.

[2] B. D. V. Veen and K. M. Buckley, "Beamforming: A versatile approach to spatial filtering," *IEEE ASSP Magazine*, April 1988.

[3] J. G. Proakis and D. K. Manolakis, *Digital Signal Processing: Principles, Algorithms and Applications*, Prentice Hall, Oct. 1995.

[4] D. H. Johnson and D. E. Dudgeon, *Array Signal Processing*, Prentice Hall, 1993.

[5] Y. Zhang, X. Li, and M. G. Amin, "Multi-channel smart antennas in wireless networks," *Asilomar Conf. Signals, Systems, and Computers*, Pacific Grove, CA, Oct.-Nov. 2006.

[6] X. Li, Y. Zhang, and M. G. Amin, "Performance of wireless networks exploiting multi-channel smart antennas in multipath environments," *Int. Waveform Diversity and Design Conf.*, Pisa, Italy, June 2007.

[7] A. Singh, P. Ramanathan, and B. V. Veen, "Spatial reuse through adaptive interference cancellation in multi-antenna wireless networks," *IEEE GlobeCom*, vol. 5, pp. 3092-3096, Nov. 2005.