Paper

A wideband smart antenna employing spatial signal processing

Monthippa Uthansakul and Marek E. Bialkowski

Abstract— A smart antenna with capability of beam steering in azimuth over a wide frequency band using only spatial signal processing is presented. Filters and tapped-delay networks employed in conventional wideband linear arrays are avoided by using a two-dimensional rectangular array structure. In this array, only constant real-valued weighting coefficients, realized with amplifiers or attenuators, are used to form a desired radiation pattern. In order to estimate direction of arrival of a wideband signal, the MUSIC algorithm in conjunction with an interpolated array technique is applied. In the interpolated array technique, a composite covariance matrix is generated, which is a simple addition of covariance matrices of narrowband virtual arrays, being stretched or compressed versions of a nominal array. A working prototype of this wideband array is presented. Its operation is assessed via full EM simulations and measurements.

Keywords— direction estimation, null steering, wideband beamforming, wideband smart antenna, wideband antenna.

1. Introduction

It has been recently envisaged that in order to meet high speed data transmission while avoiding undesired interference, future wireless systems have to utilize wideband smart antennas [1].

Until to date, most of the works on smart antennas have concerned narrow band communication systems that are characterized by a fractional bandwidth in the order of one to a few percents. The beamforming techniques used in these narrowband systems are unsuitable for wideband counterparts because they cause such adverse effects as main beam squinting and null shifting [2]. In turn, adapting wideband beam and null steering used in radar are unattractive to communication systems because of the use of a large number of filters or tapped delay networks [2, 3] which add to complexity and cost of the system. In order to overcome this hurdle fully spatial wideband beam forming techniques have been suggested for use in wideband smart antennas.

In the present paper, we design and develop a wideband smart antenna employing only spatial signal processing. The working principle of this array antenna has been introduced in [4]. It has been demonstrated by assuming a large number of antenna elements all in the form of point sources. The use of large number of elements was required for the proper functioning of beam steering algorithm, which involved an inverse district Fourier transform (IDFT) applied to an assumed radiation pattern. The effect of mutual coupling on performance of this wideband beamformer has not been considered. Another issue, which has not been addressed in [4], concerns direction of arrival (DOA) estimation of a wideband signal.

The present paper covers the above shortfalls and provides the following new contributions. It describes a refinement to the beam and null steering algorithm making it valid for small and large size wideband arrays. The modified algorithm neglects mutual coupling effects in the array, similarly as in the original algorithm [4]. However, its impact is assessed by performing full EM simulations and measurements on the array with real antenna elements. In order to overcome problems associated with increased sidelobe levels due to mutual coupling the following remedy is recommended. Using the modified beamforming algorithm (which neglects mutual coupling) an array with low sidelobes is initially designed. As a result of this strategy, when the mutual coupling is present the sidelobe level becomes acceptable. In order to obtain a DOA estimation algorithm which preserves spatial signal processing, an interpolated array technique initially demonstrated for a one-dimensional wideband linear array in [5] and [6], is adapted.

2. Design

2.1. Configuration

The configuration of a wideband smart antenna that employs a fully spatial signal processing for beam and null steering in the azimuth direction is shown in Fig. 1. It is constituted by $N_1 \times N_2$ wideband antennas arranged in a rectangular lattice. Amplifiers or attenuators connected to individual array elements produce weighting coefficients. The signal is combined by a summing network. In Fig. 1, d_1 and d_2 represent array spacing in two orthogonal directions and are usually chosen as half-wavelength at the highest frequency of a given frequency band of operation. Antenna elements are denoted by indices m_1 and m_2, where $-M_1 \leq m_1 \leq M_1$ and $-M_2 \leq m_2 \leq M_2$. The relation between N and M is $M_i = (N_i - 1)/2$. The purpose of using a 2D instead of 1D array is explained as follows. As the signal is assumed to arrive from the direction not perpendicular to the array's plane, the array's elements receive the signal's replicas with different phases. This results in a set of signals, which can be used for processing both in frequency and angular domains. If a 1D array is used, filters or tap-delay networks are necessary to process a wideband signal in these two domains.

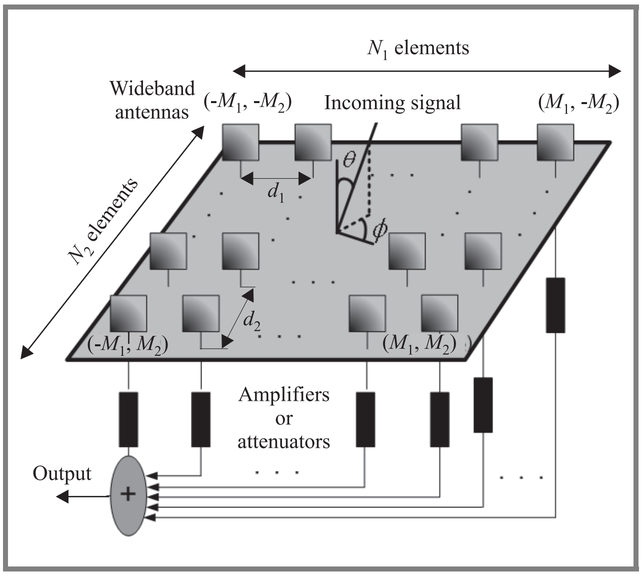

Fig. 1. Configuration of wideband beamformer constituted by wideband antenna elements arranged in a rectangular lattice followed by amplifiers/attenuators and a summing network.

2.2. Beamforming algorithm

Assuming that the signal arrives from azimuth ($\theta \sim 90°$), which is the usual case in mobile communication, the radiation pattern of the array being the function of angle and frequency is given by

$$H(f,\phi) = G_a(f,\phi) \sum_{m_1=-M_1}^{M_1} \sum_{m_2=-M_2}^{M_2} W_{m_1 m_2} e^{-j(\frac{2\pi f}{c})(d_1 m_1 \sin\phi + d_2 m_2 \cos\phi)}, \tag{1}$$

where f is the frequency variable, c is the speed of signal and $G_a(f,\phi)$ stands for the frequency-angle dependent gain of each antenna element.

This formula neglects mutual coupling effects in the array and assumes that individual elements feature identical radiation pattern. In order to determine the weighting coefficients $W_{m_1 m_2}$ in Eq. (1) a modified IDFT is applied to H as shown in formula (2):

$$W_{m_1 m_2} = \left(\frac{1}{N_{u_1} N_{u_2}}\right) \left\{ \frac{\sum\limits_{u_1=-0.5}^{0.5} \sum\limits_{u_2=-0.5}^{0.5} H(u_1,u_2)}{G(u_1,u_2) e^{-j2\pi u_1 m_1} e^{-j2\pi u_2 m_2}} \right\}, \tag{2}$$

where u_1 and u_2 are defined by $(fd_1/c)\sin\phi$ and $(fd_2/c)\cos\phi$, respectively. The modification concerns the numbers of sampling points N_{u_1} and N_{u_2} in $u_1 - u_2$ plane. In [4], $N_{u_1} = N_1$ and $N_{u_2} = N_2$ were used. The d_1 and d_2 are element spacing in two directions of the array.

According to our investigations, this assumption leads to insufficient constraints on the weights when a small number of elements form the array. As a result, a radiation pattern significantly deviating from the assumed one is produced. The use of $N_{u_1} = 2N_1$ and $Nu_2 = 2N_2$ eliminates this problem. This number of sampling points is used in our modified beamforming algorithm. Note that in calculating the weighting coefficients in formula (2), the given frequency bandwidth is slightly larger than the one over which the smart antenna has to operate [4]. The reason is to avoid the edge effect when implementing IDFT technique.

2.3. Null steering

The null steering task is similar to the beam forming task and is related to devising a radiation pattern featuring a main beam accompanied by nulls. For small size arrays, only the task of producing a single null is usually considered. In the proposed approach a simple step linear function, which assumes a main beam being directed to a desired signal direction and a null in interference direction, is used to generate the desired radiation pattern H in $u_1 - u_2$ plane. Having assumed H, the signal weighting coefficients are obtained using formula (2).

2.4. Direction of arrival estimation

Forming the beam towards the desired user and nulls towards interfering signals relies on the assumption that DOA of desired signal, as well as of interfering signals, is known to the system. This task is accomplished using a DOA estimation method. Unfortunately, many of them [7] are not suitable for the wideband array antenna, which is investigated here. This is because the DOA estimation method of wideband signals has to be accomplished using only spatial signal processing. Here, we adapt the solution in [5] and [6] that concerns one-dimensional (1D) wideband arrays. In this method, a wide frequency band is divided into multiple narrow bands with center frequencies $\{f_i\}$. In each band the signal is assumed to be received by a virtual array, which is "stretched" or "compressed" version of nominal array. The stretch/compress factor is such that all of the virtual arrays have the same radiation pattern (response). Because the virtual arrays have the same response (at their operating frequencies) it is possible to combine the covariance matrices for the different frequencies by simple addition. The outputs from virtual arrays are obtained via interpolation technique from the real one-dimensional array. The interpolation coefficients are selected so as to minimize the interpolation error for a signal arriving from a given sector (a range of bearing angles), at a particular frequency. The size of the sector has to be chosen to give good estimates of the virtual array outputs. The design of interpolator is done once and off-line. For the 2D array investigated here, the steering vector is defined in Eq. (3):

$$\mathbf{A}(\omega_i) = \left[\mathbf{a}(\omega_i,\phi_1)\, \mathbf{a}(\omega_i,\phi_2)\, \ldots\, \mathbf{a}(\omega_i,\phi_K)\right], \tag{3}$$

where

$$\mathbf{a}(\omega_i,\phi_k) = \left[\mathbf{a}_{-M_2}(\omega_i,\phi_k)\, \ldots\, \mathbf{a}_{M_2}(\omega_i,\phi_k)\right]^T \tag{4}$$

and

$$\mathbf{a}_{m_2}(\omega_i,\phi_k) = \left[e^{j\phi_k(-M_1,m_2)}\, \ldots\, e^{j\phi k(M_1,m_2)}\right]^T, \tag{5}$$

and k is the number of a radiating source in azimuth.

Having obtained (via the virtue of an interpolated array technique), the transformation of the problem of DOA estimation of a wideband signal into DOA estimation of narrowband signal, the application of narrowband MUSIC algorithm is straightforward. The detailed steps of this algorithm are shown in [7] and thus are not repeated here.

3. Results

3.1. Prototype development

In order to demonstrate the validity of the described concept of wideband array antenna, a 4×4-element prototype for operation in the band from 1.9 to 2.5 GHz is developed. The photograph of the full prototype is shown in Fig. 2. In order to meet wideband performance and tight array spacing, the array uses a square planar monopole as its element.

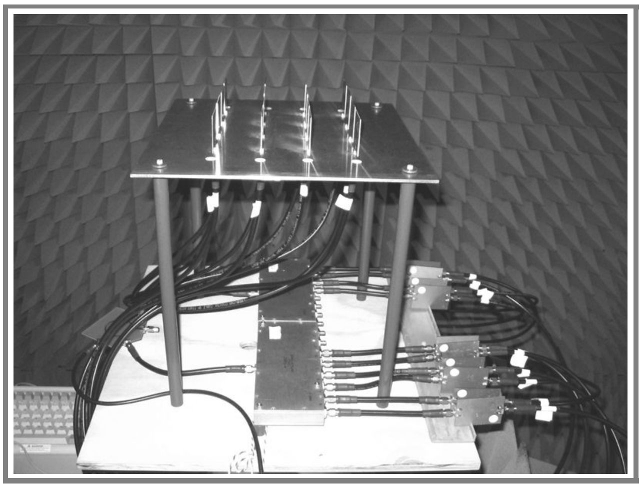

Fig. 2. Photograph of a 4×4 wideband array antenna with planar monopoles, two 1:8 dividers, attenuators and a rat-race hybrid.

This antenna element is designed using full EM analysis software based on the method of moments, FEKO® [8]. The 40-millimeter side size square planar monopole located 3 mm above the ground plane offers the 10 dB return loss bandwidth from 1.4 to 3.2 GHz, as proved by full EM simulations and measurements. The size of this antenna element meets half-wavelength spacing of $d_1 = d_2 = 5$ cm at 3 GHz. The feeding network uses two 1:8 power dividers, a rate race hybrid (to obtain signals of positive and negative amplitudes) and microstrip attenuators. The attenuators and the rat race hybrid are designed with the use of Agilent ADS® [9]. As the array is square in shape DOA and beam forming capabilities need to be checked only over the range of $-45°$ to $+45°$ angular sector. This is because the remaining ranges within the full sector of $-180°$ and $+180°$ are covered with respect to the other sides of the array.

3.2. Direction of arrival estimation performance

Figure 3 shows the results for DOA estimation using 2D (4×4) array compared with 1D (4×1) array [6] when the desired signal comes from the angles of 30° and $-20°$ off boresight direction. To obtain the presented results,

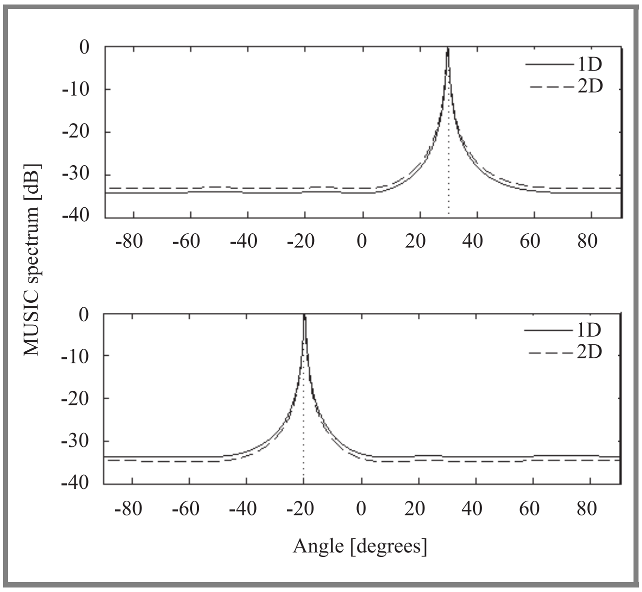

Fig. 3. MUSIC spectrum, $SNR = 10$ dB, fractional bandwidth = 27.3%, interpolated sector $[-45°/+45°]$, 7 frequency bins, 400 snapshots and 50 experiments.

400 snapshots (samples) in time domain, 7 frequency bins and 50 experiments over the angular sector of $-45°/+45°$ with the step size of 1° were used. The obtained results using the proposed DOA method for the 2D array are similar to those obtained with the DOA estimation algorithm devised for a 1D array [6].

3.3. Beamforming capabilities

Figure 4 shows the simulated (without and with mutual coupling) and measured results for the radiation pattern when the array points its main beam in the 45° off boresight direction. The assumed radiation characteristic H was chosen in the form of Chebyshev polynomial of 4th order to minimize side lobes, according to the strategy described in the paper. In order to achieve this radiation pattern, attenuators of 10, 5, 3 and 0.4 dB were used in the beamforming network.

It can be observed that the presence of mutual coupling increases sidelobe levels. However, as the assumed characteristic H (without mutual coupling) exhibits low side lobes, the actual radiation pattern (with mutual coupling taken into account) features an acceptable side lobes level. In all of the three cases presented in Fig. 4 the radiation patterns stay almost the same across the investigated frequency band between 1.9 and 2.5 GHz. In addition a good agreement between the measured and simulated (including mutual coupling) results is observed.

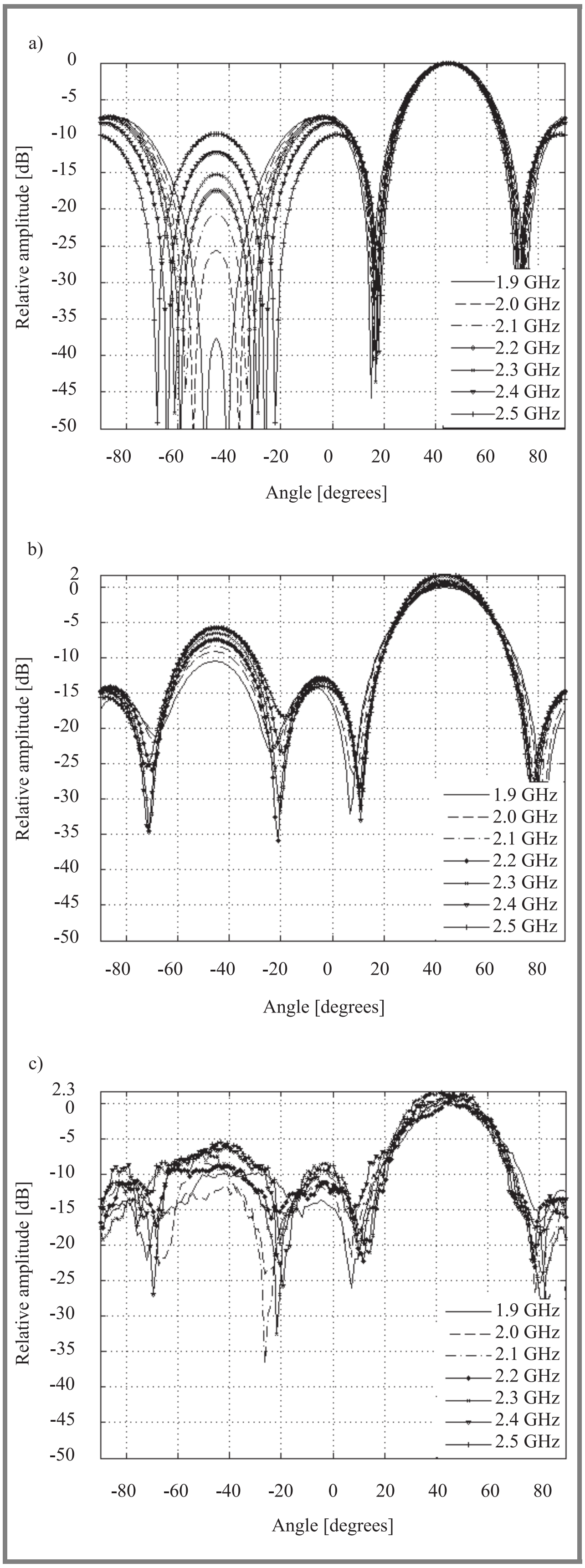

Fig. 4. Radiation pattern for the 4×4 array plotted for frequencies from 1.9 to 2.5 GHz when the desired direction is 45°: (a) simulated without mutual coupling; (b) simulated with mutual coupling; (c) measured.

Figure 5 presents the radiation pattern of the 5×5 array plotted from 1.9 to 2.5 GHz when the desired and undesired signal directions are 45° and 0°, respectively. The obtained result indicates that the array is perfectly able to direct its main beam to 45° while the null located at 0° varies from -29 to -21 dB throughout the designated band. The sidelobes are at a lower level than of the 4×4 array.

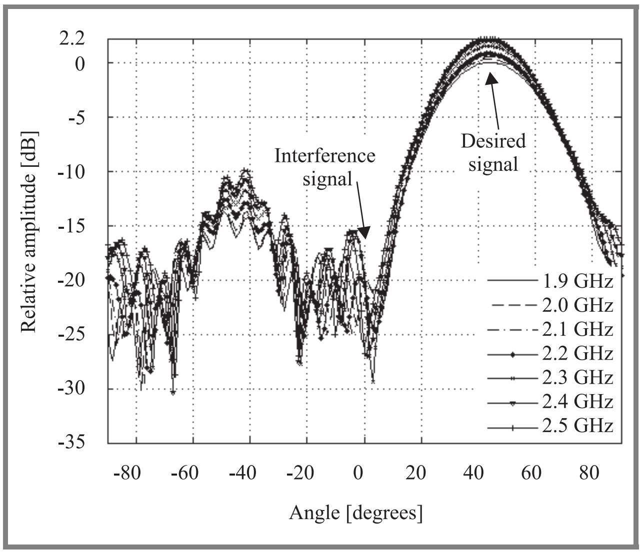

Fig. 5. Radiation pattern of a 5×5 element array plotted from 1.9 to 2.5 GHz when the desired beam direction is 45° and the desired null direction is 0°.

4. Conclusion

The design of a compact smart array antenna capable of beam formation in azimuth over a wide frequency band using only spatial beam forming technique has been described. The antenna requires only real weighting coefficients to form a radiation pattern which is approximately constant with frequency. This weighting scheme is of practical value, as it can be realized using only amplifiers and attenuators without resorting to filters, phase shifters or delay circuits that are employed in conventional wideband beamformers. In this array, DOA estimation of a wideband signal is based on the interpolated array technique and the MUSIC algorithm. It has been shown that the 2D spatial beamformer offers a similar quality DOA estimation of a wideband signal as its 1D counter part. The presented antenna system is of low cost and thus should be of interest to designers of wideband smart antennas for wireless communications supporting high data rate transmission.

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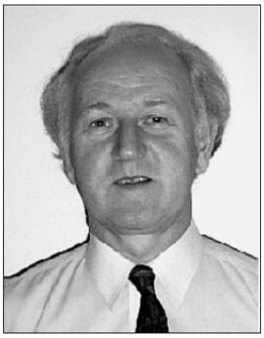

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