Concept of Microwave Electronic Steered Array using Analogue FIR-Filter

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Abstract — A concept for wideband electronic steered array using analogue FIR-filters has been formulated. This paper focuses on microwave applications, e.g., wideband antenna radiation pattern, where we explain the design of a desired radiation pattern for bandwidth 2:1. Therefore we consider a linear narrowband radiation pattern and expand it to a wideband array synthesis. The requirements for higher bandwidth, e.g. 3:1 in ultra wideband (UWB) communication systems, w.r.t. different design parameters are obtained. Finally, we propose the requirements for practical realization at microwave signal level.

I. INTRODUCTION

Recently, interest for very wideband antennas has grown, e.g. in applications for Pulse Radar [1], Synthetic Aperture Radar and for UWB-pulse communication systems [2]. Requirements for antenna pattern characteristics have been established going beyond those for conventional array antennas and active phased array antennas designed for radar and communication systems: Instantaneous bandwidth of e.g. 3:1 is required with wide angle beam scan, side-lobe control, control of nulls and with constant beamwidth and constant beam pointing and null-direction over frequency [3]. Such goals currently drive many workers in the field of signal processing and smart antennas, where it is assumed that the signals from the elements of an array can be processed digitally in a computer (digital signal processing) and beams are formed digitally [4]. However, digital beamforming seems to be improper for microwave applications in the GHz-range due to extremely high sampling rate.

Looking at the beamforming concepts being employed, it is seen that frequency dependent weighting coefficients must be used in order to make antenna patterns frequency independent and that some concepts use the Finite Impulse Response (FIR) filter structure to realize such responses [5]. The aim of this paper is to propose the use of analogue FIR-filter structures (at microwave signal level) for the design of wideband electronic steered array antennas.

II. MICROWAVE CONTROL CONCEPTS

In conventional microwave electronic steered arrays (ESA) we make use of phase shifters (PS), amplitude shifters (AS) and true time delay-elements (TTD) in order to control the weighting factor of elemental signals before summation [6], Fig. 1. Typical characteristics of the phase shifters are the frequency independent phase shift, Fig. 2a, which results in a squint of the main beam with frequency. In broadband (instantaneous bandwidth) array antennas, we use TTD-elements [7] with linear phase variation versus frequency, Fig. 2b, in order to keep the beam constant with frequency. However, both techniques result in a shift of nulls and in a widening of the beam with frequency. Amplitude control is employed usually to set the side-lobes at low levels but can in principle be employed also to reduce the active length of the array, resulting in wider beams; phase spoiling has also been used for that purpose. However, both techniques would require either frequency dependent amplitude control or phase control, both non-linear with frequency in a rather complex pattern, e.g. as shown in Fig. 2c, which can be realized, e.g. by a FIR-filter.

A. FIR-Control Elements

In principle, the frequency dependence of phase and amplitude at each antenna element, as required from array theory, can be realized by a number of filter types. One particularly appropriate type is the FIR-filter, because it can be thought of as an extension of the TTD-control element with integrated amplitude control and the additional freedom of inserting several different delays at the same time, see Fig. 3. On each antenna element, the FIR-filter causes a phase response that varies with frequency. So, spatial phase shifts due to different frequencies are temporally equalized by the FIR-filters.
III. DESIGN PROCEDURE OF A WIDEBAND ANTENNA RADIATION PATTERN

The design procedure for the wideband radiation pattern begins with the design of a narrowband radiation pattern. Here, all known synthesis techniques can be applied, e.g. uniform weighting or windowing methods [8], [9] in order to reduce sidelobe level. This design step delivers the required number of narrowband antenna elements \( I \) and (complex) weighting coefficients \( w_i \).

Using these results, a frequency independent spatio-temporal response function is obtained [10], [11], from which the FIR-filter coefficients can be determined. In order to demonstrate the viability of our concept, we consider the narrowband radiation pattern of \( I = 5 \) element antenna array with equal spacing of \( \lambda/2 \) between the elements and uniform antenna weighting \( |w_i| = 1 \). The main beam direction or steering angle should be \( \theta_0 = 20^\circ \); the resulting radiation pattern is shown in Fig. 4.

We design this radiation pattern to be constant over a bandwidth 2:1 using the array-FIR structure shown in Fig. 5. A linear array is assumed to have \( N \) antenna elements; the spatial distance \( d \) between the antenna elements is \( \lambda/2 \) for the highest frequency \( f_h \). Each element is connected to an \( M \)th order FIR-filter with the real weighting coefficients \( a_{nm} \) and delay lines of time delay \( \tau \). The input signal \( x(t) \) is obliquely incident under an angle \( \theta \). The resulting output signal \( y(t) \) can be obtained in time-domain as

\[
y(t) = \sum_{n=1}^{N} \sum_{m=1}^{M} a_{nm} x(t-(n-1)T_0-(m-1)\tau)
\]

(1)

A delay time \( T_0 \) exists between the antenna elements and is obtained by

\[
T_0 = \frac{d}{c_0} \sin \theta,
\]

(2)

where \( c_0 \) is the velocity of light. Eqn. (1) can be transformed to frequency domain, yielding the frequency response of the system as a function of frequency \( f \) and incident angle \( \theta \) as follows

\[
H(f, \theta) = \frac{Y(f, \theta)}{X(f)} = \sum_{n=1}^{N} \sum_{m=1}^{M} a_{nm} e^{-j2\pi f \left((n-1)\frac{d}{c_0} \sin \theta + (m-1)\tau\right)}
\]

(3)

The real filter coefficients \( a_{nm} \) can be obtained by performing the 2D inverse discrete Fourier transformation (IDFT) on the frequency response \( H(f, \theta) \) [8]. The resulting wideband antenna radiation pattern for three different frequencies in the band 2:1 is shown in Fig. 6. For the simulation, the delay time \( \tau \) is assumed to be \( d/c_0 \). It can be seen that the wideband radiation pattern does not match exactly with the desired radiation pattern. The latter is an envelope, containing the main beam and sidelobe level of the wideband pattern. The smaller mainbeamwidth of the wideband pattern compared to the desired one, is due to the higher number of antenna elements \( N > I \). However, in main beam direction, the wideband patterns coincide in the investigated frequency band between 0.45\( f_h \) and 0.9\( f_h \), where \( f_h \) is the highest frequency and serves as a reference for the interelement spacing \( d \). It should be mentioned that the frequency \( f_h \) is chosen to be greater than what is required for the frequency band. Since we use a limited...
number of weighting elements, the frequency response $H(f,	heta)$ strongly varies at the frequency band boundaries. However, in the chosen frequency band 2:1, also the nulls are constant (up to a signal level of -32 dB) and the sidelobe levels are below the envelope.

A. Required number of antenna elements and filter-order

The required number of antenna elements $N$ and the FIR-filter order $M$ depend on various design parameters: The desired beam pattern, bandwidth and main beam direction $\theta_0$. Theoretically, an infinite number of antenna elements and infinite filter order would result in an ideal frequency independent beam pattern. But in practice, the maximum viable array element number and filter order are limited. Ref. [10] concludes that the number of antenna elements and the filter order should be more than three times the numbers of the desired narrowband array. But this assumption neglects the desired bandwidth and main beam direction $\theta_0$. In addition, ref. [11] presents an expression taking into account the main beam direction $\theta_0$ in the determination of the array size. Nevertheless, we can check the congruence of our wideband radiation patterns by calculating their correlation; ideally, the correlation should be equal to one. For a given finite number of array and filter-elements, the correlation is lower than one. The worst congruence or the minimum correlation is between the radiation pattern of the lowest and highest frequency in the desired array. In our design example, Fig. 6, the minimum correlation is $x_{cor} = 0.97$, which indicates good congruence. We arbitrarily assume that the congruence in the frequency band is suitable, if the minimum correlation is greater than $x_{cor} = 0.97$.

Fig. 7 shows a plot of the minimum correlation as a function of the array element number $N$ and the filter order $M$ for the desired wideband radiation pattern. We can conclude from this simulation that a minimum antenna element number of $N = 20$ and a filter order of $M = 11$ should be chosen. It can be shown that the optimal ratio of array element number and filter order varies as a function of the main beam direction $\theta_0$. Angles close to $\theta_0 = 0^\circ$ require lower filter order and greater array size. With increasing angle $\theta_0$ this ratio changes, up to $\theta_0 = 90^\circ$, where the maximum filter order and the minimum array element number is required.

Fig. 8 shows a plot of the normalized weighting coefficients $a_{nm}$, resulting from our design example. The major weights are located on a diagonal line, whose angle is a function of the steering angle $\theta_0$. This can be explained by considering a simple TTD-beamformer: The signal $x_1(t)$, reaching the first antenna element, has to be delayed maximum in order to equalize the spatial phase-shift. The required delay decreases from antenna element to element. The higher the steering angle, the higher is the required delay. So, the line of major weights rotates as a function of the steering angle.

B. Requirements for different bandwidths

It is clear that the requirements for the array element number $N$ and the filter order $M$ varies w.r.t. the desired bandwidth and the narrowband synthesis target, e.g. the required beamwidth, number of nulls, etc. Table I summarizes the results for various bandwidths and steering angles, using the above correlation method. The angle $\theta_0 = 0^\circ$ requires the largest array size, whereas the angle $\theta_0 = 90^\circ$ requires the highest filter order. In addition to the five-element uniform narrowband weighting, a Tschebyscheff-windowing was used with 30 dB sidelobe attenuation.

It can be seen that both the antenna element number and the filter order increase with the bandwidth. Furthermore, the Tschebyscheff-windowing yields a relief of the requirements.
TABLE 1

<table>
<thead>
<tr>
<th>bandw.</th>
<th>narrowband pattern</th>
<th>$\theta_0 = 0^\circ$</th>
<th>$\theta_0 = 90^\circ$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.5:1</td>
<td>$I = 5$ (uniform weighting)</td>
<td>N = 15</td>
<td>N = 14</td>
</tr>
<tr>
<td></td>
<td>$I = 5$ (Tscheb. windowing)</td>
<td>M = 5</td>
<td>M = 9</td>
</tr>
<tr>
<td>2:1</td>
<td>$I = 5$ (uniform weighting)</td>
<td>N = 20</td>
<td>N = 16</td>
</tr>
<tr>
<td></td>
<td>$I = 5$ (Tscheb. windowing)</td>
<td>M = 7</td>
<td>M = 13</td>
</tr>
<tr>
<td>3:1</td>
<td>$I = 5$ (uniform weighting)</td>
<td>N = 20</td>
<td>N = 20</td>
</tr>
<tr>
<td></td>
<td>$I = 5$ (Tscheb. windowing)</td>
<td>M = 7</td>
<td>M = 15</td>
</tr>
</tbody>
</table>

This is because the Tschebyscheff-windowing increases the narrowband beamwidth. Thus, the requirements for the wideband radiation pattern are reduced, resulting in a lower number of required antenna elements and lower filter order.

So, for the 2:1 bandwidth, a minimum array element number of $N = 16$ elements and a FIR-filter of $10 (M - 1)$ incremental steps is required in order to steer all angles $\theta_0$, assuming $I = 5$ narrowband elements with Tschebyscheff-windowing. For uniform element spacing of $\lambda/2$ (at highest frequency), the total array length becomes $(N - 1) \cdot \lambda/2 = 15/2$ and main beamwidth of $12.6^\circ$.

IV. CONCEPT FOR PRACTICAL REALIZATION

The amplitude weighting factors $\alpha_{nm}$ of the FIR-filter requires bi-phase variable attenuators ($-1 \leq \alpha_{nm} \leq 1$), which can be realized broadband and as integrated circuits. Practical problems of realization of microwave FIR-filter structures will therefore arise mainly from the spacious delay lines which have to fit in a predictably limited volume and which have to be connected to the attenuator (integrated) circuit.

The variability of the filter response in practical realizations must be limited due to the limitations on the total length of the delay possible and the size of increments $\tau$. In general, time delay increments can be realized at microwave frequencies by using sections of transmission line, either on-chip as lumped LC-networks [12] or off-chip.

Nevertheless, the principle setup of the single FIR-elements, Fig. 9, reminds of a traveling wave amplifier [13]. The weightings $\alpha_{nm}$ are realized by FET-amplifiers, which are distributed along microstrip-lines, realizing the time-delay $\tau$. By biasing the FET-gates, the amplification of each transistor can be varied according to the required weighting coefficients. This simple concept neglects the need for both positive and negative weighting coefficients. In [14] we present a modified concept, providing in-phase and out-phase amplification.

Using the traveling wave concept, our design example in practice, for a frequency band from 2 GHz to 4 GHz, would require a total array width of 64 cm ($f_0 = 4.45$ GHz, $d = \lambda/2$)$. Realizing the time delays $\tau = \varepsilon_{eff} \lambda/2$ of each constant FIR-filter in microstrip-line techniques with effective dielectric constant $\varepsilon_{eff} = 2.3$, the total length of the transmission lines shrinks to 22 cm. This length should be reduced dramatically, if the transmission lines are coiled, as indicated in Fig. 9. Such array-FIR structure size should be acceptable, e.g. in access point applications.

V. CONCLUSION

The design procedure of a FIR-filter controlled antenna array for wideband radiation pattern and some trade-off considerations have been presented in this paper. On a concrete design example, we proposed that the required number of antenna elements is between 2 and 6 times that of the narrowband element number, the filter order is between 5 and 15, both depending on the desired bandwidth and steering angle. Our concept for practical realization at microwave signal level is promising due to the use of standard RF-components.

Our future research will focus on the robustness of the wideband radiation pattern design. We will investigate the results in view of systematical and statistical variation of the design parameters due to practical tolerance problems. This could be the non-ideal variation of interelement spacing, time-delay, weighting coefficients, non-uniform gain of the antenna elements, influence of mutual coupling between antenna elements, etc.

REFERENCES