FIR-filter based RF-beamforming network for wideband and ultra wideband antenna arrays

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Abstract

In this paper, a wideband electronic beamforming concept applicable for UWB short pulse systems is derived and experimentally demonstrated. In particular, a Finite Impulse Response (FIR)-filter based beamforming network is configured such that an antenna array yields a desired radiation pattern within a predefined frequency range. An experimental verification of a wideband transmission system is presented for the frequency range from 1.5 GHz to 2.0 GHz. The transmit system consists of a wideband power divider feeding a matrix of analog FIR-filters. These FIR-filters control a circular antenna array of four elements. The receiver consists of a single antenna element. Measurements are performed in frequency domain using a vector network analyzer. The results show an almost frequency independent radiation pattern with a half power beamwidth of 72° and a maximum sidelobe level of -10 dB.

Index Terms

antenna array, UWB, beamforming, FIR-filter, UWB-antenna, mutual coupling, equalization

I. INTRODUCTION

Recently, the so-called ultra-wideband (UWB) technology has become very popular, since the American Federal Communications Commission (FCC) defined in 2002 a standard, ruling the commercial use of this technology [1]. Core of the regulation is a frequency mask with very low spectral power density over an enormous bandwidth of about 7.5 GHz. To use the enormous UWB bandwidth two different concepts are essentially discussed: the carrier based multiband UWB-OFDM (Orthogonal Frequency Division Multiplex) [2] and the carrier-free impulse-UWB, which transmits data by short pulses. Both approaches provide a number of bandwidth-demanding low power applications in fields of wireless communications, networking, radar imaging, and localization systems [3]. However, the low spectral power density close to the noise floor also demands a sophisticated optimization in terms of system sensitivity, efficiency and range increase. One optimization step could be to focus system power towards a desired direction; other steps could be to keep interference levels due to sidelobes low or to use frequency-independent

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patterns to counter pulse distortion. In conventional wireless systems, spatial processing has been known for decades and has experienced a renaissance in these days under the key word “smart antennas” [4].

Spatial processing for ultra wideband signals is a new challenge in research and is not well investigated up to now. Conventional beamforming concepts due to the enormous bandwidth are not sufficient. Mathematical frameworks for frequency-independent beamforming are presented in [5, 6]. For impulse-UWB applications, beamforming networks with true time delays (TTD) have been investigated [7, 8]. Only very few publications engage in the context of time-domain beamforming on other geometries than linear antenna arrays, e.g. planar or circular antenna arrays [9, 10]. In addition, many of the contributions on ultra wideband beamforming consider antenna arrays with isotropic antenna elements without any mutual coupling effects, which is not sufficient in realistic scenarios.

The aim of this paper is to frame an appropriate approach to wideband electronic beamforming applicable for UWB signals and to present first experimental results in the microwave frequency range. We focus on impulse-UWB applications because this approach provides new challenges on beamforming concepts, which cannot be satisfied by conventional concepts. Nevertheless, the results can be transferred easily to multiband UWB-OFDM or other systems with very wide instantaneous bandwidth.

Our paper is organized as follows: In Section II the mathematical framework describing antennas in the time-domain is presented. The model is exploited for antenna arrays, considering mutual coupling effects. This delivers the impulse response of the embedded antenna element and is part of the FIR-filter based beamforming concept described in Section III. Theory is explained for a linear antenna array and can be easily adapted for circular antenna arrays. On the basis of this mathematical framework, a real-time demonstrator, consisting of a four-element circular antenna array equipped with analog FIR-filters and a wideband power divider as a beamforming network has been realized, Section IV. A Proof-of-concept experiment was designed for a 28% relative bandwidth frequency range from 1.5 GHz to 2.0 GHz, compatible with conventional printed circuit board (PCB)-technology and the use of surface mount devices (SMD). Finally, in Section V, a concept and measurement results for an electronically adjustable FIR-filter for a fractional bandwidth of more than 3.2 at microwave frequencies is presented.

II. Time-Domain Antenna Modeling

A single antenna can be characterized in time-domain by its effective height $h_t(\hat{r}, t)$, where $\hat{r}$ is the unity vector of the observation point. Alternatively, this unity vector can be described by the pair of azimuth and elevation angle $(\varphi, \theta)$. In transmitting mode, this quantity connects the relation between the incident current $I^+(t)$ at the feeding port of the antenna and the radiated farfield as [11]

$$E_{\text{rad}}(r, t) = -\frac{\mu}{4\pi r} \frac{\partial}{\partial t} h_t(\hat{r}, t) * I^+(t) * \delta(t - r/c_0)$$

(1)

and serves as the impulse response of the transmitting antenna $h_t(\hat{r}, t)$. The notation (*) indicates a convolution operator, $r$ is the observation point vector on the distance $r$, $c_0$ is the speed of light in vacuum and $\mu$ is the permeability. Note that the impulse response of the antenna is defined with respect to the incident current $I^+(t)$, not with respect to the total current $I(t)$. Hence, possible reflections by antenna mismatch are contained in the impulse response.
For the reception mode, the impulse response $h_r(\hat{r}, t)$ connects the traveling voltage wave $U^-(t)$ at the antenna’s output port and the incident electric field:

$$U^-(t) = -\frac{1}{2} E_i(\hat{r}, t) * h_r(\hat{r}, t). \quad (2)$$

Note that the impulse responses in (1) and (2) are vector quantities describing the two cases of horizontal and vertical polarization of the antenna. They can be written as scalar quantities, if no cross-polarization is considered.

The antenna impulse response can be determined by simulations or by measurements. The former is done numerically by appropriate analysis of the current distribution on the antenna surface [11]. Measurements can be performed in frequency-domain using a vector network analyzer and transformed to time-domain by inverse discrete Fourier transformation [12]. Unfortunately, the definitions for the transmitting antenna impulse response $h_t(\hat{r}, t)$ and the receiving antenna impulse response $h_r(\hat{r}, t)$ is not unique and there are various proposals in literature [11–14]. In (1) and (2) we used our own definition related to traveling wave quantities. If the transmitting and the receiving antenna are identical, the relationship $h_t(\hat{r}, t) = h_r(\hat{r}, t) = h(\hat{r}, t)$ can be derived.

A. Impulse response of an embedded antenna element

If an antenna element is embedded in an antenna array, the total radiated electric field becomes a function of the individual antenna impulse responses and feeding currents. With reference to (1) the following expression for the total radiated electric field can be found:

$$E_{rad}(r, t) = -\mu \frac{1}{4\pi r} \frac{\partial}{\partial t} \{ h_1(\hat{r}, t) * I_1^+(t) + h_2(\hat{r}, t) * I_2^+(t) + \ldots + h_N(\hat{r}, t) * I_N^+(t) \} * \delta(t - r/c_0) \quad (3)$$

where $h_n(\hat{r}, t)$ is the impulse response of the $n$th antenna element embedded in an array of $N$ antennas. Mutual coupling is considered by expanding the impulse response of the embedded antenna elements as:

$$h_n(\hat{r}, t) = h_{in}(\hat{r}, t), \quad (4)$$

where $h_{in}(\hat{r}, t)$ describes the impulse response of the $i$th antenna element in the array connected to the $n$th feeding current, see Fig. 1. The impulse response of the $n$th embedded antenna element can be simulated or measured with respect to its position in the antenna array and the antenna array geometry [15].

III. TIME-DOMAIN BEAMFORMING

A. Principle of the FIR-filter concept

In conventional phased array or digital beamforming concepts [16], we multiply the signal at each antenna element by a complex weighting factor before radiation. Since these complex weighting factors are frequency independent, it is not sufficient for ultra wideband signals. Also at the use of true-time delays (TTD) replacing phase shifters, where the phase varies linearly with frequency, the radiation pattern depends on frequency. Rather, a weighting
element with a general frequency dependent characteristic respecting amplitude and phase is required. Technically this can be realized by Finite Impulse Response (FIR) filters [17], common devices in the field of digital signal processing. At a first glance, digital FIR-filters could be used in digital beamforming for UWB-signals. However, the extremely high bandwidth places great demands on A/D converters and is currently not feasible. Alternatively, FIR-filters can be realized as analog circuits, particularly in the microwave frequency range [18].

The principle of electronic beamforming in transmitting mode with FIR-filters is depicted in Fig. 2. A train of four short pulses is generated by a source and divided into \( N \) branches by a power splitter. Each signal passes a weighting function \( w_n(t) \) and feeds an antenna elements embedded in a linear array. The antenna elements are denoted by the embedded impulse response \( h_n(\hat{r}, t) \). The radiated electric farfield \( E_n(r, t) \) due to the \( n \)th antenna element is related to (1) and Fig. 2a

\[
E_n(r, t) = -\frac{\mu}{4\pi r} \frac{\partial}{\partial t} h_n(\hat{r}, t) * w_n(t) * I_n^+(t - t_{0,n}) \tag{5}
\]

where \( I_n^+(t) \) describes the feeding current at the \( n \)th antenna element. Assuming an ideal lossless power splitter the current through the \( n \)th branch is

\[
I_n^+(t) = \frac{1}{\sqrt{N}} I^+(t). \tag{6}
\]

The term \( t_{0,n} \) is the spatial time-delay of the \( n \)th antenna element relative to the first antenna element in the farfield observation point. Assuming a linear antenna array (Fig. 2a) the spatial time-delay is

\[
t_{0,n} = (n - 1) \frac{d}{c_0} \sin \varphi, \quad 0 \leq \varphi \leq \pi. \tag{7}
\]

Thus, the total radiated electric farfield at the observation point is

\[
E_{\text{rad}}(r, t) = -\frac{1}{\sqrt{N}} \frac{\mu}{4\pi r} \frac{\partial}{\partial t} \left\{ \sum_{n=1}^{N} h_n(\hat{r}, t) * w_n(t) * \delta(t - t_{0,n}) \right\} * I^+(t). \tag{8}
\]

The time-domain radiation signal is shown as a function of the angle-of-departure \( \varphi \) for a simple example in Fig. 2b. At \( \varphi_0 = 30^\circ \), a train of four short pulses can be clearly distinguished. For other angles the signal amplitudes decay and do not exceed a given sidelobe level (SLL); here, although not explicitly shown on the graph, \( \text{SLL} \leq -20 \text{ dB} \).

The weighting functions \( w_n(t) \) comply with the impulse response of appropriately configured FIR-filters. Particularly they are configured such that the transmitted signals are constructively superposed in a desired direction (main beam). For all other directions, the signals superpose destructively, respecting a prescribed radiation pattern. From this, it could be concluded that the weighting functions \( w_n(t) \) are just true-time delays (TTD). In fact, we do have to compensate the spatial time difference between the antenna elements in the farfield observation point: Each weighting function can be represented as a zero-degree weighting function \( w_{0,n}(t) \), which would be required for beamforming at broadside (no scan), convoluted with the spatial time-delay compensation \( \tau_{\text{feed},n} \) required for scanning of the beam, that is

\[
w_n(t) = w_{0,n}(t) * \delta(t - \tau_{\text{feed},n}). \tag{9}
\]
Assuming a linear antenna array, as in Fig. 2a, the spatial time-delay compensation is
\[
\tau_{\text{feed},n} = (N - n) \frac{d}{c_0} \sin \varphi, \quad 0 \leq \varphi \leq \pi; \quad n = 1 \ldots N.
\] (10)

Unlike for a typical time-delay steering, the zero-degree weighting function \( w_{0,n}(t) \) in (9) is not a constant but a general time-dependent function, which has to be realized by the FIR-filters. This degree-of-freedom is necessary in order to get full control over destructive and constructive signal superposition over all angle and frequency range. This in turn allows the optimization of radiation patterns w.r.t. system requirements, e.g., a frequency independent radiation pattern regarding main beam direction, beamwidth or sidelobe level (SLL) or null positions or all at the same time, which presents the most demanding pattern requirement.

Another (optional) function of the weighting functions is the equalization of antenna and channel effects [19]. Both antenna and channel characteristics are functions of frequency and they operate like filters in UWB-systems. This filtering may change the shape of the pulse being transmitted. If these frequency responses are moderately flat, the set of weighting functions can compensate them.

In Fig. 3, the structure of the FIR-filter based beamforming network is depicted. Each branch is equipped with a FIR-filter, which forms the signal before it is radiated by an antenna. A FIR-filter combines time delay sections \( \tau \) with amplitude control elements \( a_{nm} \). These FIR-Filter coefficients are real and biphase, therefore
\[
-1 \leq a_{nm} \leq 1 \quad a_{nm} \in \mathbb{R}.
\] (11)

The incremental time delay \( \tau \) is due to the Nyquist-criteria
\[
\tau \leq \frac{1}{2f_h}
\] (12)
where \( f_h \) is the upper limit of the frequency band. To avoid grating lobes for any frequency the interelement distance \( d \) is chosen to be maximum \( \lambda/2 \) for the highest frequency in the band, thus
\[
d \leq \frac{c_0}{2f_h}.
\] (13)

The impulse response of the \( n \)th FIR-filter with \( M \) stages can be obtained related to Fig. 3
\[
w_n(t) = \sum_{m=1}^{M} a_{nm} \delta(t - (m - 1)\tau).
\] (14)

From (8) and (14) the impulse response of the total antenna array an be obtained as
\[
h_{\text{arr}}(\hat{r}, t) = \sum_{n=1}^{N} h_n(\hat{r}, t) * \delta(t - t_{0,n})
\]
\[
* \sum_{m=1}^{M} a_{nm} \delta(t - (m - 1)\tau)
\] (15)

Applying the Fourier transformation on (15) yields the transfer function of the antenna array:
\[
H_{\text{arr}}(\hat{r}, \omega) = \sum_{n=1}^{N} H_n(\hat{r}, \omega)
\]
\[
* \sum_{m=1}^{M} a_{nm} e^{-j\omega(t_{0,n} + (m - 1)\tau)}
\] (16)
The challenge is now to determine the matrix of FIR-filter coefficients related to a desired radiation pattern and frequency range. Therefore, an appropriate steering algorithm is required.

**B. Steering algorithm**

The steering algorithm to set up the FIR-filter coefficients starts with a desired radiation pattern \( H_{\text{ref}}(\hat{r},\omega) \), which covers the main characteristics like steering direction, beamwidth, sidelobe level, nulls, etc. In the case of a frequency independent desired pattern, this is the reference for the pattern between the lower and upper frequency limit. Depending on system requirements, the reference pattern could also be chosen differently, e.g., relaxing the sidelobe level and the null directions. However, in this section and in section IV, we chose the most demanding specification of a fully frequency-independent pattern in order to demonstrate the method. The upper frequency limit automatically defines the incremental time delay of the FIR-filters and the interelement distance, see (12) and (13). After choosing preliminary orders \( N \) and \( M \), the FIR-filter coefficients can be obtained numerically, e.g. by mathematical optimization like the least squares method or convex optimization,

\[
\text{minimize } \| H_{\text{arr}}(\hat{r},\omega) - H_{\text{ref}}(\hat{r},\omega) \|. \quad (17)
\]

Here, \( \| \cdot \| \) indicates the norm of the expression and \( H_{\text{ref}}(\hat{r},\omega) \) the transfer function of the reference radiation pattern, while the transfer function of the antenna array is calculated based on the array parameters, the embedded elements transfer functions \( H_n(\hat{r},\omega) \) and the filter coefficients assumed in each optimization step. If the resulting radiation pattern agrees appropriately with the reference radiation pattern for all frequencies, the algorithm stops and delivers the final coefficient matrix for the FIR-filter beamforming network. In our investigations, we used the SeDuMi software package (GNU/GPL open source license) in Matlab for the optimization and used simulated antenna element impulse responses \( H_n(\hat{r},\omega) \) calculated by electro-magnetic (EM) simulation (CST Microwave Studio). The quality of the agreement of the patterns can be measured by pattern correlation and expressed by a correlation coefficient, which is compared to a target correlation coefficient [17]. If the current correlation coefficient is less than the target correlation coefficient, the number of elements \( N \) and \( M \) has to be increased. For the case that the current correlation coefficient is grater than the target correlation coefficient, \( N \) and \( M \) can be decreased to ensure that no elements are wasted. It is seen that the filter coefficients depend on the transfer function of the embedded elements; in a real antenna system, the transfer functions may differ among the elements depending on the chosen array geometry and may vary if the environment close to the array changes. In any case, precalculated or measured transfer functions may be stored and recalled by the steering algorithm; or precalculated sets of filter coefficients for all required beam patterns may be stored and recalled by a beam steering processor. A calibration system may be required to check the stability of the antenna array and the electronic FIR-filter and to allow adaptation to changing close-in environment.

Using this algorithm, it is also possible to compensate additional frequency dependent factors of the UWB transmission chain. Thus, the optimization problem in (17) can be expanded as follows:

\[
\text{minimize } \| H_{\text{comp}}(\hat{r},\omega) H_{\text{arr}}(\hat{r},\omega) - H_{\text{ref}}(\hat{r},\omega) \|. \quad (18)
\]
where the vector $H_{\text{comp}}(\tilde{r}, \omega)$ includes all additional angular-frequency dependent factors of the transmission chain, e.g. the time-derivative in (8) and the impulse response of the receiving antenna.

C. Circular antenna array

So far, derivations were made by assuming a linear antenna array. In the experimental part of this paper, a circular antenna array is used, where $N$ antenna elements are uniformly located along a ring with the radius $a$. The mathematical framework to configure the beamforming network for an UWB circular antenna array is identical to the linear one and the equations can be adopted. Only in (15) and (16) the spatial time-delay of a circular antenna array has to be considered, which is [20]

$$t_{0,n} = \frac{a}{c_0} \sin \theta \cos(\varphi - \varphi_n)$$

with

$$\varphi_n = 2\pi \left( \frac{n}{N} \right)$$

and

$$a \leq \frac{c_0}{4\pi} \frac{N}{f_h}.$$  

In (19) it is assumed that the elements are located on the azimuth plane ($xy$-plane).

IV. Experimental verification

For the experimental verification of the FIR-filter based beamforming network, a complete wideband transmission system is configured. It consists of a circular antenna array in transmission mode and a single receiving antenna, see Fig. 4. The antenna elements are modified Monocone-antennas, presented in [21]. The antenna array is controlled by appropriately configured analog FIR-filters. A broadband Wilkinson power divider uniformly divides the input signal and feeds the FIR-filters. Measurements are performed in frequency domain by a vector network analyzer (VNA) HP 8510. From the measured transmission between the antenna array and the receiving antenna, the transfer function $H_{\text{arr}}(\varphi, f)$ of the complete antenna array (antenna elements + beamforming network) can be extracted and related to (16). Measurements are performed in the azimuth plane ($xy$-plane) only, between $-180^\circ \leq \varphi \leq 180^\circ$; cross-polarization is omitted in this task.

The task of this experiment is now to realize a desired radiation pattern within a given wide frequency range. Therefore, FIR-filters have to be realized and configured appropriately, as described in section III of this paper. However, the choice of the upper frequency limit in the demonstrator system was dictated by the technology available to us for the realization of the FIR filter circuits. The desired radiation pattern requirements concerning beamwidth, SLL and bandwidth had to be chosen such that the filter order was kept in a manageable low range. In this context, it becomes clear that the analog FIR-filters are the critical components in the wideband transmission chain which at present do not allow us the realization of a transmission system covering the UWB regulated frequency range between 3 and 10 GHz.
A. Analog FIR-Filter concept

The main challenging task to demonstrate RF beamforming on wideband-signals is to create a suitable concept for the analog FIR-filters. It consists mainly of bi-phase adjustable attenuators or amplifiers and delay lines. Fig. 5 shows the structure, which can be realized as microstrip circuit on a dielectric substrate (MIC-technology). The input signal is fed to an input transmission line and a traveling wave propagates towards the matched end termination. The transmission line is segmented with interconnecting extra transmission line loops representing the time delay units. If created in monolithic integrated circuit technology, they may be realized off-chip in order to keep the chip size low or they may also be realized on-chip as lumped LC networks, a similar approach as used in a microwave equalizer circuit, recently published in [22]. Within each segment of the transmission line, there is one bi-phase variable attenuator connected to the input signal transmission. The input impedance of the attenuator has to be high in order not to disturb the forward traveling wave on the input-line. The attenuated signal is coupled to the output transmission line on the other side. There, we also terminate by a matched load at one end while the other port supplies the output signal to a matched load. Signal injection from each segment has to sum up on the output transmission line, which can be achieved by the creation of a traveling wave on the output transmission line and assuming superposition of contributions from each segment.

In order to keep well in the limits of our presently available technology of conventional hybrid MIC or PCB-technology and the use of surface mount devices (SMD), the FIR-filter circuits for our demonstrator were designed for an upper frequency of 2.0 GHz and a fractional bandwidth of 1.33; as seen later in Section V, this technology limits our maximum frequency to approximately 3 GHz such that a demonstration circuit covering the regulated UWB band up to 10 GHz is unattainable presently. Further, in this experimental verification the FIR-filter circuits employed fixed attenuators as weighting stages. In Section V of this contribution, we show the concept of an adjustable analog FIR-filter in microstrip-line technique, which can be digitally controlled and which operates over a fractional bandwidth of 3.2.

B. Configuration of the experimental setup

From evaluation of the array transfer function, it is found that the attainable relative bandwidth of the array is limited mainly by the chosen order of the FIR filters, which can be seen in Table I. Here, the required filter order $M$ has been calculated as a function of the ratio of upper frequency limit to lower frequency limit $f_h/f_l$ (fractional bandwidth) and the number of antenna elements $N$. The latter also fixes the half power beamwidth (HPBW) of the radiation pattern. Assuming the lowest number for $N$ and $M$ (both four), we get a fractional bandwidth of 1.33 and HPBW of 72°. This makes a lower frequency limit of 1.5 GHz and an absolute bandwidth of 500 MHz. It is also seen from the table that a four-element array designed for the full UWB fractional bandwidth of 3.2 would require at least a filter order of $M=8$ which was judged an unnecessary high effort for the present proof of principle experiment.

For the target frequency range from 1.5 GHz to 2 GHz, we have chosen the required transfer function of the antenna array to be frequency independent such that its azimuth pattern characteristics should agree with a reference
TABLE I: Required number of antenna elements $N$ and filter order $M$ of a circular array related to the half power beamwidth (HPBW) and relative bandwidth, assuming $\text{SLL}\leq-10\text{ dB}$; spacing of elements constant and radius of the array increasing with the number of elements $N$.

<table>
<thead>
<tr>
<th>HPBW ($^\circ$)</th>
<th>elements $N$</th>
<th>bandwidth $f_h/f_l$</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>1.33</td>
<td>2</td>
</tr>
<tr>
<td>72$^\circ$</td>
<td>4</td>
<td>4</td>
</tr>
<tr>
<td>40$^\circ$</td>
<td>8</td>
<td>4</td>
</tr>
<tr>
<td>16$^\circ$</td>
<td>16</td>
<td>4</td>
</tr>
</tbody>
</table>

Radiation pattern for all frequencies across the band. Considering the transfer function of the coupled Monocone-antenna elements and applying (17), yields the following normalized matrix of coefficients:

$$A_{\text{arr}} = \begin{bmatrix}
-0.23 & -0.62 & 0.37 & -0.62 \\
-0.95 & 1 & -0.19 & 1 \\
0.83 & -0.67 & 0.18 & -0.67 \\
-0.74 & 0.34 & 0.02 & 0.35
\end{bmatrix} \quad (20)$$

Note that the coefficients for the 2nd and 4th FIR-filter are identical due to the rotational symmetry in the azimuth plane.

C. Measurement results

In Fig. 6, a picture of the four-antenna circular array can be seen. The transmitting antenna array is mounted on a rotatable table, which enables radiation pattern measurements. The transmitting antennas’ corresponding FIR filters and power divider is mounted beneath the antenna array table. The entire project setup is shown in an anechoic chamber.

In Fig. 7a, the measured transfer function $H_{\text{arr}}(\varphi,f)$ is depicted as a function of frequency and the azimuth angle. It can be seen that the transfer function represents an almost frequency independent radiation pattern of the antenna array. The comparison to the reference radiation pattern can be seen in Fig. 7b. The curves indicate a good correlation to the reference pattern. The sidelobe level is mainly below -10 dB, nulls are approximately at a constant angle and below -20 dB relative to the main beam direction.

V. CONCEPT OF ADJUSTABLE ANALOG FIR-FILTERS

The analog FIR-filters for the experimental verification of the beamforming concept were not electronically adjustable. This kept the circuits simple, but very inflexible. In order to enable electronic beamforming, electronically adjustable FIR-filters are required. Hence, in this section the concept and experimental results of a single adjustable FIR-filter is presented.
The general concept and structure of an adjustable FIR-filter coincides with the one, which was already described in Section IV-A, depicted in Fig. 5. The challenge is the realization of an adjustable weighting stage for a wideband frequency range, see Fig. 8a. At the input, a so-called phase splitter divides the input signal into two output signals, which are phase shifted by 180°. The phase splitter mainly consists of a single field effect transistor (FET), where the positive and negative signal paths follow the source and drain terminals, respectively.

Next, the signal amplitude is controlled by adjustable amplifiers or attenuators, where a digital control unit translates the weighting function of the corresponding FIR-filter into analogue steering voltages. If the digital control unit indicates a positive coefficient for the corresponding weighting stage, only the positive signal path is active whereas the negative signal path is closed and vice versa. At the output, both signal paths are combined by a resistive power combiner. A common gate circuit yields a high output impedance of the weighting stage.

The technical requirements for the weighting stages are high, because within the desired frequency range the following specifications should be met:

- frequency independent weighting amplitude
- constant phase difference of 180° between the positive and negative signal path
- high input- and output impedance.

An experimental layout of a fourth-order FIR-Filter in microstrip-line technique is presented in Figure 8b, where the top metalization layer can be seen. The circuit is designed on RT/Duroid 5870 substrate with thickness of 0.787 mm and dielectric constant of 2.33. In the upper part of the circuit, three transmission line loops can be recognized, which represent the incremental time delays \( \tau \). The four weighting stages are arranged perpendicularly to the input transmission line. In vicinity to each phase splitter, a finger structure can be recognized, representing a small capacitance between source terminal and ground. This capacitance improves the 180°-phase difference between the two signal paths for higher frequencies. The other passive elements are mainly SMD 0805; the large cases at the bottom part of the circuit contain RF-chokes. The active elements are Mitsubishi Electric MGF 1302 field effect transistors. The FIR-filter circuits have been simulated and optimized using Agilent Advanced Design System (ADS).

For RF measurements, the circuit is connected to a digital control unit, which translates the digital weighting functions into analog steering voltages using 8-bit digital potentiometers. The measurement results of the transmission amplitude and phase are shown in Fig. 9. Here, a single weighting stage is opened to its maximum, whereas the other weighting stages are closed. Fig. 9a shows the amplitude of the positive and negative signal path of the investigated weighting stage in comparison to an ideal reference value. The flatness of the curves is acceptable within the frequency range from 0.76 GHz to 2.6 GHz, a fractional bandwidth of 3.6. The phase of both paths and the phase difference is depicted in Fig. 9b. It also shows acceptable result for the investigated frequency range.

A summary of all measured results is given in Table II. The experimental results prove that an electronic analog FIR filter circuit for an ultrawide fractional bandwidth larger than 3.2 can be successfully realized. However, due to the employed circuit technology, the operating frequencies are limited to a "scaled-down" range of the regulated UWB band. Presently, an extension of our MIC technology by using wideband drop-in amplifiers is
under investigation which would allow the frequency range to be extended. However, as an ultimate solution for the practical application of the FIR-filter beam forming approach, it seems necessary to realize FIR filter circuits in monolithic IC technology, leading to small-size and low-cost components.

<table>
<thead>
<tr>
<th>parameter</th>
<th>property</th>
</tr>
</thead>
<tbody>
<tr>
<td>frequency range</td>
<td>$f = 0,76 \ldots 2,6$ GHz</td>
</tr>
<tr>
<td>return loss (in- and output)</td>
<td>$&lt;-10$ dB</td>
</tr>
<tr>
<td>max. weighting amplitude</td>
<td>$\pm 0,34$ (steerable in 256 steps)</td>
</tr>
<tr>
<td>switching time</td>
<td>$18 \mu$s</td>
</tr>
<tr>
<td>variation amplitude vs. freq.</td>
<td>$&lt; 20 %$</td>
</tr>
<tr>
<td>variation phase difference vs. freq.</td>
<td>$&lt; 16 %$</td>
</tr>
<tr>
<td>power consumption</td>
<td>$&lt; 100$ mW</td>
</tr>
</tbody>
</table>

VI. CONCLUSION

In this paper, a method for wideband beamforming has been presented and experimentally verified. A FIR-filter based beamforming concept has been exploited for the practical realization in the Gigahertz frequency range. A mathematical framework for linear and circular antenna arrays was derived, considering all relevant influences like antenna characteristics and mutual coupling effects. Analog FIR-filters with fixed weighting stages have been realized in conventional MIC-technology and were used in a beamforming network for a proof-of-concept four-element circular antenna array. The experimental system showed an almost frequency independent radiation pattern within the frequency range from 1.5 GHz to 2.0 GHz. Finally, a fourth order electronically adjustable FIR-filter has been designed and experimentally characterized for a frequency range from 0.76 GHz to 2.6 GHz using MIC-technology.

This contribution also discussed the application of the FIR-filter based wideband beamforming to UWB systems, considering all relevant effects in the mathematical framework. However, practical implementations would require small-size and low-cost FIR filter circuits covering the frequency range up to 10 GHz. This requires monolithic integration technologies instead of the MIC technology used in the proof-of-concept experiment. Yet, the required capabilities are already available in present CMOS and SiGe IC technologies as was demonstrated by an UWB equalizer circuit realized in CMOS technology [22]. Never-the-less, a monolithic microwave analog FIR filter circuit presently has no example and a design would predictably require a research project by its own to cover the development of extensions of available circuit functionalities and design rules and the evaluation of potential circuit concepts.

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concept considering antenna and channel effects,” in *European Conference on Antennas and Propagation (EuCAP)*, Nice, France, Nov. 2006.


![Fig. 1: Composition of the impulse response of an antenna embedded in an antenna array with \( N = 2 \) elements.](image)
Fig. 2: (a) Principle of time-domain beamforming using a FIR-filter based beamforming network and (b) its radiated electric farfield as a function of the angle-of-departure $\phi$. Here we consider that a train of four short pulses has been generated and the beamforming network is configured such, that the main beam is pointed to $30^\circ$ and the sidelobe level is below -20 dB.
Fig. 3: FIR-filter structure with $N$ antenna elements and $M$th filter order.

Fig. 4: Principle setup for the experimental verification.

Fig. 5: Concept for an analog FIR-filter in microstrip-line technique.
Fig. 6: Photography of the setup for measuring the transfer function of the FIR-filter controlled antenna array.

Fig. 7: (a) Measured transfer function of the circular antenna array as a function of frequency and azimuth angle, (b) Comparison with reference radiation pattern at three frequency points.
Fig. 8: (a) Components of a single weighting stage, (b) photography of a FIR-filter in microstrip-line technology with four digitally adjustable weighting stages.
Fig. 9: Measurement results of the transmission behavior of a single weighting stage, embedded in the FIR-filter. (a) Maximum amplitude of the positive and negative signal path, (b) phase and phase difference vs. frequency.