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Practical Beamforming Technologies for Wideband Digital Array Radar

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Abstract—Wideband digital beamforming (WDBF) technology is a key for the rapidly developing wideband digital array radar (WDAR). In this paper, by comprehensively considering the characteristics of WDAR and the current hardware and software capabilities for radar in engineering, several practical WDBF technologies based on accurate digital true time delay (TTD) are studied. WDBF technology at radio frequency (RF) is applied and tested on a WDAR test-bed. Besides, WDBF technologies at baseband by implementing fractional delay filers at element level and subarray level are presented and simulated.

1. INTRODUCTION

As a further development of narrowband digital array radar [1–4], wideband digital array radar (WDAR) uses wideband signal and has advantages in areas such as range resolution and anti-interference ability [5]. Wideband beamforming is a fundamental and key technology for WDAR and other wideband arrays [6]. Narrowband beamforming assumes that the complex envelopes of array signals are approximately the same; however, the complex envelopes of wideband array signals cannot be assumed approximately the same because they vary rapidly over time; therefore, WDAR must resolve the aperture effect caused by aperture transit time to achieve wideband digital beamforming (WDBF). Generally, there are about two ways to solve the aperture effect and realize WDBF. The first way is to change wideband signal into narrowband signal by Discrete Fourier Transforming (DFT) before DBF, which is classified as wideband DBF in frequency domain [7, 8]; the other one is to compensate the aperture transit time for array signals, which can be classified as wideband DBF in time domain. Besides, the space-time adaptive processing (STAP) based on Frost architecture can also achieve adaptive WDBF; however, the heavy computation complexity makes it difficult for STAP to be realized in real-time according to the recent engineering capability.

Nowadays, WDAR is mainly in the experimental stage. Some test beds implement stretch processing as WDBF method [5, 9], which can only be applied to the receiving of linear frequency modulation (LFM) signal. In addition, a combination of analog and digital true time delay (TTD) devices with 6-bit RF phase shifters to achieve wideband beamforming is also applied in WDAR test-bed [10]. To compensate the aperture fill time of any wideband radar signal in receiving or transmitting mode, TTD units should be implemented, and they can be realized by digital methods or analog devices [11–13].

WDAR requires high level of digitization, with the fast development of software & hardware platform and digital processers such as analog to digital converter (ADC)/digital to analog converter (DAC), FPGA and DSP chip [14], and adopting complete digital processing to achieve receiving and transmitting wideband beamforming of WDAR becomes more feasible and attractive. In the article, we

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firstly present a WDBF technology based on digital phase shift and digital integer delay at RF band which has been applied on our WDAR test bed in Section 2. Section 3 covers WDBF technologies by implementing digital fractional delay (FD) filters at element level and subarray level. In Section 4, the article is concluded, and some possible future work is introduced.

2. WDBF TECHNOLOGY BASED ON DIGITAL INTEGER DELAY

For an arbitrary array with N array elements and working in receiving mode, assume that the signal on the reference array element which is farthest away from the far-field target is given by

$$x_1(t) = \mathrm{rect}\left(\frac{t}{T_p}\right) u(t) e^{j2\pi f_0 t} \tag{1}$$

$$\mathrm{rect}\left(\frac{t}{T_p}\right) = \begin{cases} 1, & |t| \leq T_p/2 \\ 0, & \text{others} \end{cases} \tag{2}$$

where $u(t)$ is the complex envelope of $x_1(t)$, and f_0 is the carrier frequency. Accordingly, the signal on array element i $(i = 1, 2, \cdots, N)$ can be written as

$$x_i(t) = \mathrm{rect}\left(\frac{t+\tau_i^\theta}{T_p}\right) u\left(t+\tau_i^\theta\right) e^{j2\pi f_0\left(t+\tau_i^\theta\right)} \tag{3}$$

τ_i^θ is the time delay between the signal on reference array element and the i-th array element. For a uniform line array (ULA), $\tau_i^\theta = (i-1)d\sin\theta/c$, where c is the speed of light, and θ is the array pointing angle. If direct RF digitization is implemented, then the digitized array signal at RF can be given by

$$x_i(n) = \mathrm{rect}\left(\frac{n+\alpha_i^\theta}{T_p}\right) u\left(n+\alpha_i^\theta\right) e^{jw_0\alpha_i^\theta} e^{jw_0 n} \tag{4}$$

$$\alpha_i^\theta = \frac{\tau_i^\theta}{T_s} = G_i^\theta + g_i^\theta \tag{5}$$

where $T_s = 1/f_s$ is the AD sampling interval, $w_0 = 2\pi f_0/f_s$, $G_i^\theta = \mathrm{round}(\tau_i^\theta/T_s)$, and $\mathrm{round}(x)$ denotes the integer closest to x. After intermediate frequency (IF) converting and Zero-IF processing (DDC processing) including down conversion and decimation, we obtain the digital baseband signal

$$s_i(m) = \mathrm{rect}\left(\frac{m+\beta_i^\theta}{T_p}\right) u\left(m+\beta_i^\theta\right) e^{jw_0\beta_i^\theta} \tag{6}$$

$$\beta_i^\theta = \frac{\tau_i^\theta}{T_b} = L_i^\theta + l_i^\theta \tag{7}$$

where $T_b = 1/f_b$ is the digital baseband signal interval and $L_i^\theta = \mathrm{round}(\tau_i^\theta/T_b)$.

To align array signals and achieve sum beamforming at RF, $x_i(n)$, $i = 1, \ldots, N$ need to be processed as

$$x_d(n) = \mathrm{rect}\left(\frac{n}{T_p}\right) u(n) e^{jw_0 n} \tag{8}$$

Then the desired response of array processing in the i-th channel can be given as

$$H_i(w_s) = \frac{e^{jw_0 n}\sum_n \mathrm{rect}\left(\frac{n}{T_p}\right) u(n) e^{-jw_s n}}{e^{jw_0 n}\sum_n \mathrm{rect}\left(\frac{n+\alpha_i^\theta}{T_p}\right) u\left(n+\alpha_i^\theta\right) e^{jw_0\alpha_i^\theta} e^{-jw_s n}} = e^{-jw_0\alpha_i^\theta} e^{-jw_s\alpha_i^\theta} = e^{-jw_0\alpha_i^\theta} e^{-jw_s G_i^\theta} e^{-jw_s g_i^\theta} \tag{9}$$

where $w_s = 2\pi f/f_s$. The desired response of array processing in the i-th channel carries out complex digital phase shift $-w_0\alpha_i^\theta$ and digital delay α_i^θ to the sampled array signal at RF $x_i(n)$. As $|g_i^\theta| \leq 0.5$, if the part $e^{-jw_s g_i^\theta}$ of $H_i(w_s)$ is ignored, the response of array processing to finish a sum beamforming is

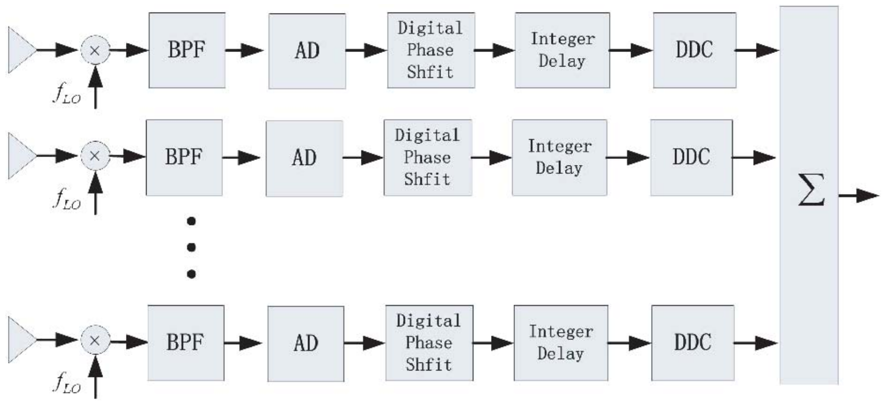


Figure 1. The diagram of WDBF for an S-band WDAR test bed.

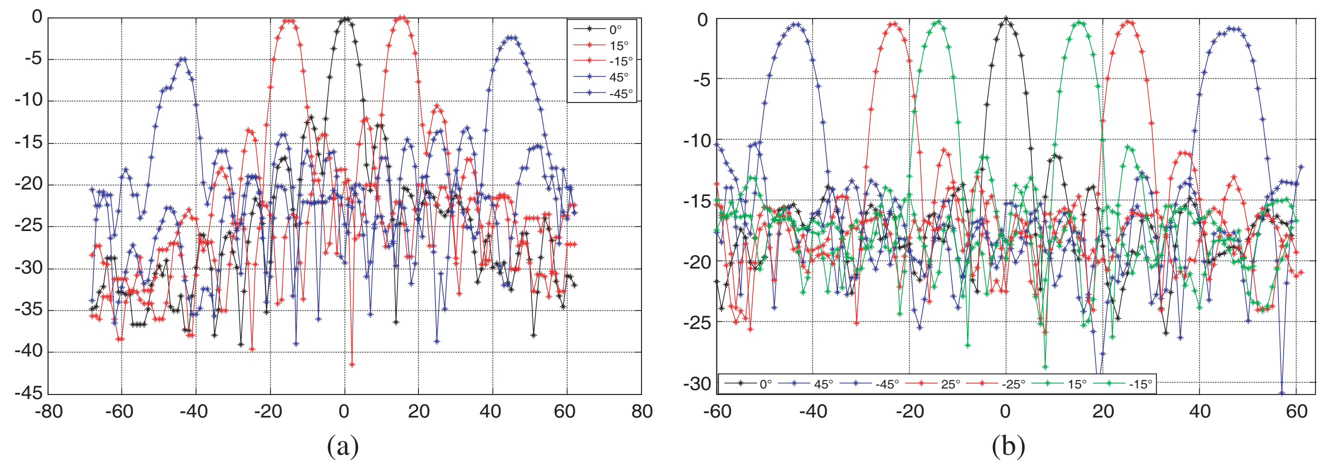


Figure 2. The measured beam patterns. (a) Transmitting beam patterns, (b) receiving beam patterns.

to carry out integer sample intervals delay and digital phase shift which can be simply realized by using digital shift register (DSR) and digital phase weighting respectively in engineering. A receiving WDBF diagram based on digital integer delay and digital phase shift for a fully digitized WDAR is shown in Fig. 1. For transmitting mode, conducting digital phase shift and digital integer delay to digital signal generated by direct digital waveform synthesis (DDWS) at RF can also achieve transmitting WDBF. On an S-band WDAR test bed digitized at RF and having a 16-elements uniform line array (ULA), LFM signal with bandwidth of 200 MHz is adopted to test the receiving and transmitting WDBF technology discussed above. The measured transmitting and receiving beam patterns are given in Fig. 2.

To digitize WDAR at RF, high-performance ADC/DAC is necessary. Currently, ADC/DAC is capable of conducting GHz-level and over 8-bits AD/DA conversion [14], so it is more and more feasible and attractive to carry out WDBF at RF. Besides, WDBF at RF is less affected by channel mismatch and provides more accurate TTD than that at IF or baseband. However, ignoring the fractional part of the array channel delay not only brings feasibility in engineering but also causes random delay error which has negative effects on array gain and sidelobe level. Just as shown in Fig. 2, the mean sidelobe level and peak sidelobes of the beam patterns are relatively high. Although more advanced ADC/DAC with smaller sample interval can decease the random delay error, cost of high-performance ADC/DAC may make it unaffordable in practice especially when there are hundreds even thousands of array elements; therefore, other WDBF technologies with more accurate TTD are proposed in the next part.

3. WDBF TECHNOLOGY BASED ON DIGITAL FRACTIONAL DELAY FILTER

3.1. WDBF Methods

To align array signals and achieve sum beamforming at baseband, $s_i(m)$, $i = 1, ..., N$ need to be processed as

$$s_d(m) = \text{rect}\left(\frac{m - m_0}{T_p}\right) u(m - m_0) \tag{10}$$

where m_0 is the time when $s_i(m)$, $i = 1, ..., N$ are aligned. For the previous beamforming technology at RF, $m_0 = 0$. In this case, the desired response of array processing in the i-th channel can be given by

$$\begin{aligned} Q_i(w_b) &= \frac{\sum\limits_m \text{rect}\left(\frac{m - m_0}{T_p}\right) u(m - m_0) e^{-jw_b m}}{\sum\limits_m \text{rect}\left(\frac{m + \beta_i^\theta}{T_p}\right) u\left(m + \beta_i^\theta\right) e^{jw_0\beta_i^\theta} e^{-jw_b m}} = e^{-jw_0\beta_i^\theta} e^{-jw_b\beta_i^\theta} e^{-jw_b m_0} \\ &= e^{-jw_0\beta_i^\theta} e^{-jw_b L_i^\theta} e^{-jw_b\left(m_0 + l_i^\theta\right)} \end{aligned} \tag{11}$$

where $w_b = 2\pi f/f_b$. In the desired channel response $Q_i(w_b)$, $e^{-jw_0\beta_i^\theta}$ and $e^{-jw_b L_i^\theta}$ mean complex digital phase shifting $-w_0\beta_i^\theta$ and integer digital delay L_i^θ, respectively. For the part $e^{-jw_b(m_0+l_i^\theta)}$ in $Q_i(w_b)$, assume that $m_0 = (N1 - 1)/2$ is an integer, then $N1$th-order digital finite impulse response (FIR) filter whose transfer function approximates $e^{-jw_b(m_0+l_i^\theta)}$ under certain criteria can be used to realize it in engineering. There are many methods to design the FIR filter, but considering the feasibility of engineering application and the performance of filter, we choose LS (least squares) criteria to design the $N1$th-order FIR filter that can be expressed as

$$\min_{H_{\text{iFIR}}(w_b)} \frac{1}{2\pi} \int_{-\alpha\pi}^{\alpha\pi} \left| H_{\text{iFIR}}(w_b) - e^{jw_b[(N1-1)/2 + l_i^\theta]} \right|^2 dw_b \tag{12}$$

$$H_{\text{iFIR}}(w_b) = \sum_{k=1}^{N1+1} h(k) e^{-jw_b k} \tag{13}$$

According to [15], the impulse response vector of the designed filter can be denoted as:

$$\mathbf{h} = \mathbf{P}^{-1}\mathbf{p_1} \tag{14}$$

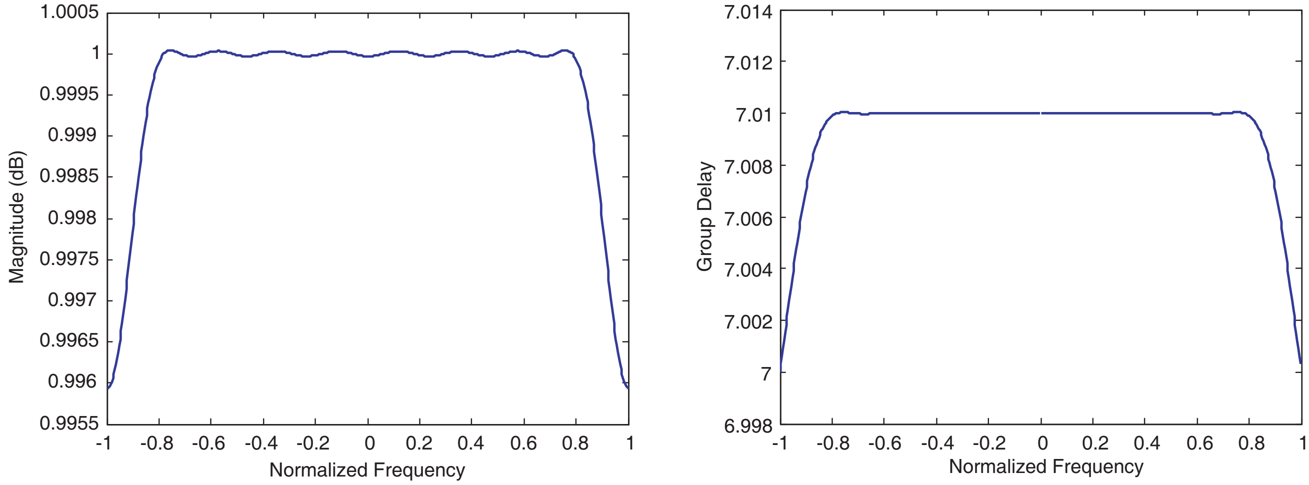
where $\mathbf{h} = [h(1), ..., h(N1+1)] \in C^{(N1+1)\times 1}$. $\mathbf{P} \in C^{(N1+1)\times(N1+1)}$ and $\mathbf{p_1} \in C^{(N1+1)\times 1}$ are denoted as

$$\mathbf{P}(k, j) = \alpha \sin c[\alpha(k - j)] \quad k, j = 1, ..., N1 + 1 \tag{15}$$

$$\mathbf{p_1}(k) = \alpha \sin c\left[\alpha\left(k - \frac{N1 - 1}{2} - l_i^\theta\right)\right] \quad k = 1, ..., N1 + 1 \tag{16}$$

where $\sin c(x) = \sin x/x$, $\alpha \in [0, 1]$ is the approximation bandwidth parameter. The digital FIR filter is called digital fractional delay (FD) filter because it can realize fractional interval delay to digital signal within the approximation bandwidth $[-\alpha\pi, \alpha\pi]$. As shown in Figs. 3(a) and (b), the 15th-order FD filter designed by LS criteria with $\alpha = 0.8$ has a flat magnitude and phase delay response within the normalized bandwidth $[-0.8\pi, 0.8\pi]$. In practice, the magnitude and phase delay response of the filter could be used to help setting the length of the FD filter. Once the order of the filter is given, since $\mathbf{P}$ is independent of the delay value, it can be calculated as mentioned previously and only needs to be inverted once. According to Eq. (11), we need to carry out complex digital phase shift, integer digital delay, and fractional delay to the digital baseband signal in each channel to achieve WDBF at baseband, where lower data rate facilitates reducing computation load, so a practical WDBF method applying FD FIR filters at baseband and element level is presented in Fig. 4.

Practically, to employ FD filters at elemental level may be too expensive in cost and system resource, especially for WDAR with a large aperture. Instead, it is more realistic to employ TTD units including integer delay and fractional delay at subarray level. Generally, in order to obtain a balance between cost



(a) (b)

Figure 3. The frequency response of 15th-order FD filter designed by LS algorithm. (a) Magnitude response, (b) group delay response.

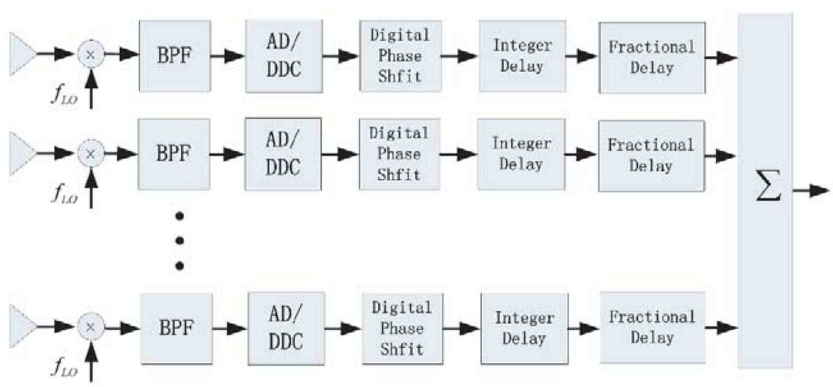


Figure 4. The diagram of WDBF technology by applying FD FIR filter at baseband and element level.

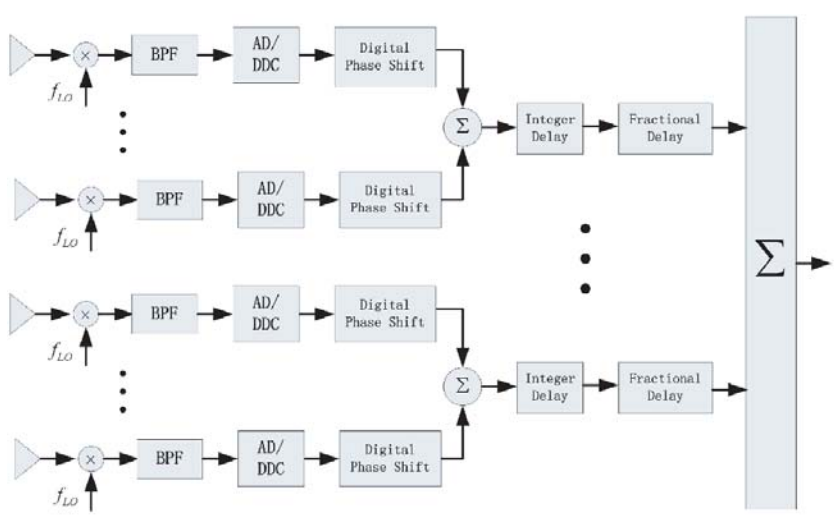


Figure 5. The diagram of WDBF technology by applying FD FIR filters at baseband and subarray level.

and performance, the aperture transit time of the subarray should be less than 80% of the reciprocal of signal bandwidth [15]. Besides, irregular rather than regular subarray layout can be used to avoid severe grating sidelobes. Therefore, the diagram of WDBF technology based on implementing FD FIR filters at baseband and subarray level is given in Fig. 5.

3.2. Simulation Results

Considering a 64-elements ULA array, the distance of array element $d = \lambda_{\min}/2$, $\lambda_{\min} = 3 \times 10^8/(f_0 + B/2)$, carry frequency $f_0 = 2.6\,\text{GHz}$, the bandwidth of the wideband LFM signal $B = 300\,\text{MHz}$, and pulse duration $T_p = 40\,\mu\text{s}$. Baseband signal data rate $f_b = 1/T_b = 350\,\text{MHz}$, array pointing angle $\theta = 60°$, and the FD filters are designed by LS algorithm.

In Fig. 6, compared with the ideal pattern obtained by phase shift and ideal delay processing, the traditional phase shift only processing cannot form correct beam pattern because of aperture effects, while the beam patterns by implementing FD filters according to WDBF technology in Fig. 4 have very close mainlobes and first sidelobes, and the overall sidelobe level can also be controlled better when the order of the FD filter is 15 rather than 7, so we can get nearly ideal beam pattern with a properly designed FD filter.

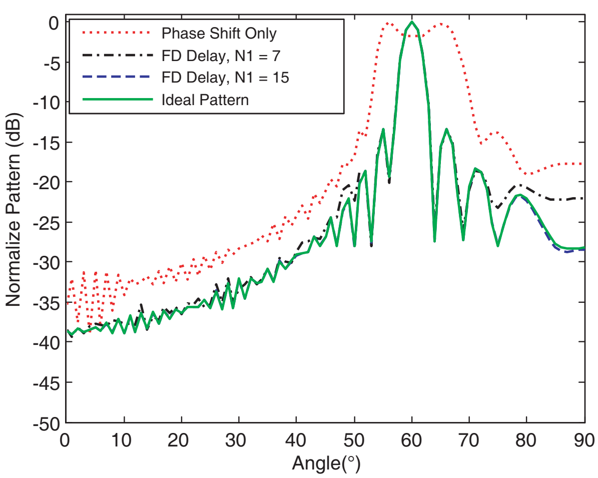


Figure 6. The beam patterns of WDBF technology by applying FD FIR filters at baseband and element level (when $N1 = 7$, $\alpha = 0.6$; when $N1 = 15$, $\alpha = 0.8$).

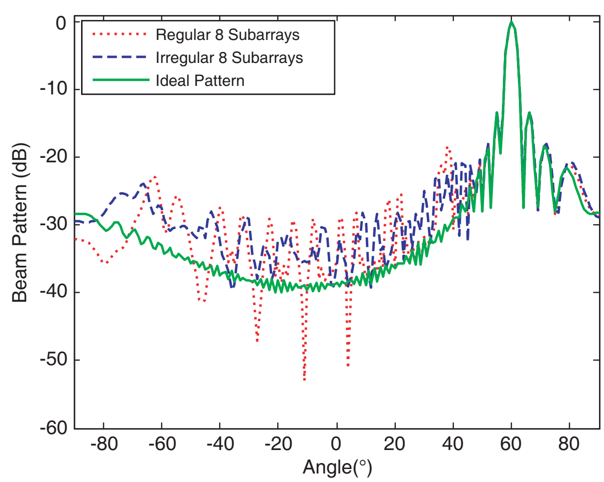


Figure 7. The beam patterns of WDBF technology by applying FD FIR filters at baseband and subarray level ($N1 = 15$, $\alpha = 0.8$).

In Fig. 7, compared with the ideal pattern defined before, the mainlobes and first sidelobes of the beam patterns with different subarray partitions are almost the same as that of ideal pattern; however, there are relatively serious grating lobes in the beam pattern obtained by a regular subarray partition while an irregular subarray partition is able to avoid severe grating sidelobes. According to the simulation results, we can come to the conclusion that the two WDBF technologies based on FD filter delay are able to form nearly ideal mainlobes and first sidelobes with correct pointing angle, and the sidelobes can also be controlled to an desired level when the FD filter is properly designed, and irregular subarray layout can effectively avoid serve gating lobes.

4. CONCLUSIONS

In this paper, practical WDBF technologies based on accurate digital delay are presented. The WDBF technology by implementing digital phase shift and integer delay at RF band has been tested on our WDAR test bed, which is easy to be realized in engineering but needs high-performance ADC/DAC to control sidelobes. Another two WDBF technologies by applying FD filters at baseband and element or subarray level are also studied. When FD filters are properly designed and used at element level, ignoring other array errors, a nearly ideal beam pattern can be obtained. Considering cost and complexity in engineering, implementing FD filters at subarray level can also form a beam pattern with nearly ideal mainlobes and first sidelobes, and serious grating lobes can be avoided if an irregular subarray partition is adopted. Our future work may be conducted to test the WDBF technologies based on FD filters on the WDAR test bed.

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